AN ABSTRACT OF THE THESIS OF

Dongsheng Zhou for the degree of Doctor of Philosophy in Electrical and Computer Engineering presented on October 30, 1995. Title: Dynamic Control of Brushless Doubly-Fed Machines.

Redacted for Privacy

Abstract Approved: ________

René Spée

This thesis presents the development of dynamic control strategies for the brushless doubly-fed machine (BDFM). A simple open-loop scalar controller is first investigated and its response is found to be oscillatory. Using the speed feedback signal, an improved closed loop scalar control algorithm is designed. Steady state performance is controlled by the magnitude of the BDFM control winding current. Simulation and experimental results demonstrate that the closed loop algorithm has a slow response and is suitable for low performance drive applications. A model reference adaptive control algorithm is investigated in simulation, attempting to improve the BDFM dynamic response and assure its robustness against system parameter variations.

Further investigations reveal that the field orientation principle for conventional induction machines can be adapted for the BDFM. This enables the design of a rotor flux oriented control algorithm, based on a newly established synchronous reference frame model. Simulation results illustrate the algorithm's fast dynamic response and
robustness against parameter variations.

The verification of various control algorithms is carried out on a laboratory system consisting of an experimental BDFM, a power converter and associated control hardware. An Intel 80196Kr microprocessor is used to implement inverter switching and current regulation for the BDFM control winding. The rotor flux oriented control algorithm is implemented using an Intel 80960KB floating point microprocessor, achieving a control bandwidth in the kHz-order.

Evaluation of a BDFM synchronous angle shows its significance in control design, and it is incorporated into the later control algorithm development in order to eliminate electric torque estimation. This simplifies control algorithm design and is verified experimentally. Consequently, the control algorithm for the BDFM can approach the simplicity of equivalent induction machine control techniques.
Dynamic Control of Brushless Doubly-Fed Machines

by

Dongsheng Zhou

A THESIS

submitted to

Oregon State University

in partial fulfillment of
the requirements for the
degree of

Doctor of Philosophy

Completed October 30, 1995
Commencement June 1996
Acknowledgements

The first thank goes to my major advisor Dr. René Spée, whose thorough and patient review of this thesis is highly appreciated. I would like to express my deepest gratitude toward professor Gerald C. Alexander and Alan K. Wallace for their generous supports and many helpful suggestions. I would also like to thank professor P. Tadepalli, W. J. Kolodziej and S. Kiaei for serving on my graduate committee.

All my fellow graduate students deserve special thanks for their helps and moments shared with me when loud sounds or big smoke occurred in the lab. Among them are Bhanu Gorti, Shibashis Bhowmik and Ernesto Wiedenbrüg. Thank Dr. Ruqi Li at FaAA Electrical, Dr. Ashok Ramchandran at EPC, Brian Wiley at Kenetech for their valuable discussions.

The ECE staff members have been very kind and helpful. I am indebted to Rita Wells, Lynn O’Hare, April Melton, Kim Rowe, Suzy Rogers and Tom Lieullen for their supports.

I would like to thank all my friends and their family, Huaisheng Chen, Yuepeng Zheng, Jinfeng Zhuang, Lisa Yang for giving me so many joyful times. I would also like to thank Terry Alexander, Cindy Spée and Pat Wallace for their kindness and friendship.

Finally, a very special thank goes to my family: Chun Zhang, Duyi Zhou, Shuxiang Fang, Dongyun Zhou and Dongwen Zhou for their understanding and supports.
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**NOMENCLATURE**

- \( V_{6a}, V_6 \): Instantaneous and rms phase voltages on the power winding
- \( V_{q6}, V_{d6} \): Power winding dq voltages, on rotor reference frame
- \( V_{q2}, V_{d2} \): Control winding dq voltages, on rotor reference frame
- \( V_{qr}, V_{dr} \): Rotor dq voltages, on rotor reference frame
- \( V_{q6e}, V_{d6e} \): Power winding dq voltages, on power winding synchronous frame
- \( i_{q6}, i_{d6} \): Power winding dq currents, on rotor reference frame
- \( i_{q2}, i_{d2} \): Control winding dq currents, on rotor reference frame
- \( i_{qr}, i_{dr} \): Rotor dq currents, on rotor reference frame
- \( i_{qr6}, i_{dr6} \): Components of the rotor dq currents which correspond to power winding excitation, on rotor reference frame
- \( i_{q2}, i_{d2} \): Components of the rotor dq currents which correspond to control winding excitation, on rotor reference frame
- \( i_{q6e}, i_{d6e} \): Power winding dq currents, on power winding synchronous frame
- \( i_{q2e}, i_{d2e} \): Control winding dq currents, on control winding synchronous frame
- \( i_{qr6e}, i_{dr6e} \): Components of the rotor dq currents which correspond to power winding excitation, on power winding synchronous frame
- \( i_{qr2e}, i_{dr2e} \): Components of the rotor dq currents which correspond to control winding excitation, on control winding synchronous frame
- \( \psi_{q6}, \psi_{d6} \): Power winding stator dq fluxes, on rotor reference frame
- \( \psi_{q6e}, \psi_{d6e} \): Power winding stator dq fluxes, on power winding synchronous frame
- \( \psi_{qr2e}, \psi_{dr2e} \): Control subsystem rotor dq fluxes, on control winding synchronous frame
- \( f_{r,m} \): Rotor shaft mechanical frequency
- \( f_p, f_o; f_c, f_2 \): Power and control winding excitation frequencies
- \( \omega_p, \omega_o \): Power and control winding frequencies in rad/second
- \( \omega_r \): Rotor mechanical speed in rad/second
- \( \theta_r \): Rotor angle displacement in mechanical degrees
- \( \theta_{6e} \): Power winding angle between rotor and synchronous reference frames
\( \theta_{2e} \) Control winding angle between rotor and synchronous reference frames

\( \theta_{62e} \) The BDFM synchronous angle

\( \theta_6 \) Power winding voltage excitation angle

\( \theta_2 \) Control winding current excitation angle

\( T_e, T_L \) Electric and load torques

\( p \) Differential operator \( p = \frac{d}{dt} \)

\( r_6 \) Power winding stator phase resistance

\( r_2 \) Control winding stator phase resistance

\( r_r \) Rotor resistance in dq domain

\( L_{s6} \) Power winding stator phase inductance in dq domain

\( L_{s2} \) Control winding stator phase inductance in dq domain

\( L_r \) Rotor inductance in dq domain

\( M_{6} \) Power winding to rotor mutual inductance in dq domain

\( M_{2} \) Control winding to rotor mutual inductance in dq domain
Dynamic Control of Brushless Doubly-Fed Machines

Chapter 1

Introduction

The objective of this thesis is to present control developments for brushless doubly-fed machines (BDFM). In this chapter, a background for the BDFM is given. Two BDFM models and the associated dynamic equations needed for control design are then described. A literature survey and a thesis outline are also included.

1.1 Background

Conceptually the BDFM can be derived from a system of two induction machines on the same shaft with cascaded rotor windings. This self cascaded induction machine system can operate as an adjustable speed drive by varying one machine stator winding frequency [1]. Such a system requires wound rotor machines for the rotor connection. Generally speaking, the system is bigger than a single cage rotor induction machine. It is desirable to combine these two machines into one machine structure and substitute the wound rotors with a cast cage. Hunt [2] showed in 1907 that by proper stator and rotor winding design it is possible to combine these two machines into one frame. Creedy [3] demonstrated that further improvement would be possible if a more effective stator and
The idea of using a cage structure rotor was introduced by Broadway [4] in the 1970's.

Although the concept of the single frame BDFM originated 90 years ago, it didn't gain much attention until recently, when its potential in variable speed systems was shown, in which it only requires a power converter of reduced rating [1]. Fig. 1.1 shows a BDFM system with a power converter connected to one set of three phase windings. As illustrated, the single frame BDFM stator consists of two sets of three phase windings which are usually excited with different frequencies. Equal pole pair numbers for these two windings are to be avoided in order to prevent direct transformer coupling. Specific pole pair combinations can be selected to assure balanced power transfer between stator and rotor. The laboratory prototype, which will be described in detail in chapter 5, uses

---

**Figure 1.1** BDFM System
a 3/1 pole pair combination. The BDFM stator windings are designed such that the winding connected directly to the grid will process the main portion of total power; while the winding connected to the power converter only has to process the small portion necessary for control purposes. Hence, they are referred to as power and control winding, respectively. Since the power converter only processes a small portion of the total power, its rating and cost can be reduced. The rotor is of a cage structure, where all conductors are connected to an end ring on one end, while forming several "nests" on the opposite end, as shown in Fig. 1.1. The number of nests is the sum of the pole pair numbers for both stator windings. Comparisons can be drawn between BDFM adjustable speed systems and other drive systems [5]:

1. The BDFM is more robust and costs less than a wound rotor induction machine;
2. The power converter needed for a BDFM is smaller and cheaper than the converter required for conventional induction or synchronous machines;
3. Due to the smaller power converter, harmonic pollution is reduced for a BDFM system;
4. A failure of the power converter will only affect the control winding in a BDFM system. The BDFM can still process power via the power winding for reduced performance operation. This high tolerance makes the BDFM suitable for critical industrial applications.

The stator windings are excited with different frequencies and possibly different phase sequences. In general, these two excitations generate two air gap electromagnetic fields which in turn lead to rotor currents of two frequencies. When the rotor speed is such that these two rotor frequencies become equal, both air gap electromagnetic fields
"lock" into each other, enabling full power transfer from the stator to the rotor or vice versa. This condition is referred as the synchronous mode of operation for the BDFM \cite{6,7}. The rotor mechanical frequency can be related to stator frequencies and pole pair numbers as

\[ f_{r,m} = \frac{f_p \pm f_c}{P_p + P_c} \]  \hspace{1cm} (1.1)\

where the indices p and c refer to power and control winding, respectively. The BDFM can also operate in the so-called induction mode, where the two induced rotor frequencies are different. Since the two electromagnetic fields are not locked into each other, only a portion of rated power can be transferred between stator and rotor, hence the BDFM efficiency decreases. Operation into the induction mode should be avoided for normal operation \cite{7}.

BDFM research activities have covered several aspects. Modeling and analysis have shed light on machine design principles and have enhanced overall machine performance \cite{8}. Practical application studies have shown new areas for the BDFM, such as car alternators and wind generator systems \cite{9}. This thesis deals with dynamic control issues and attempts to improve BDFM dynamic performance \cite{10-13}.

1.2 Literature Survey

The decoupled dq-axis representation of electrical machines has become a universal tool for developing dynamic control strategies. As shown in Fig. 1.1, the unique structure of the BDFM needs to be considered in the dynamic control design.
process. R. Li [5] discarded conventional approaches taken by previous researchers, who considered the BDFM as two magnetically separate wound rotor induction machines with electrically connected rotors. He established a new dq model from direct mathematical transformation of a detailed machine design model [14]. A. Ramchandran [15,16] developed an off-line frequency domain parameter estimation method to obtain the dq model parameters from dynamic measurements. These models and parameter estimation techniques are needed for subsequent controller development efforts.

Since BDFM research started several years ago, its dynamic performance in a drive system has continuously been investigated. R. Li [6,7] used simple constant volts/hertz control to simulate BDFM response to changes in control frequency, which is directly related to rotor speed in synchronism. Constant current control [13,17,18] has been investigated as an alternative to constant volts/hertz control. It has been shown that both constant volts/hertz and current control methods are acceptable for slow speed response applications such as pumps or fans [13,19].

W.R. Brassfield [20,21] developed a direct torque control algorithm which is based on correcting the electric torque and control winding flux errors. The algorithm uses the full sixth-order electrical differential equations of the BDFM voltage source model to predict all system quantities and generate control outputs.

A machine similar to the BDFM is the doubly-fed reluctance machine (DFRM) [22-24]. Its stator also consists of two sets of three phase windings. The DFRM uses a reluctant rotor structure to replace the nested cage structure in the BDFM. It is still in its early development stage and no dynamic control implementations have been reported in the literature.
Many control methods developed for induction or synchronous machines can be modified for the BDFM. Scalar control methods have been used extensively to control electric machines. In a scalar controller, the control objectives are achieved by adjusting the magnitude of the controller input variables. Machine flux may or may not be disturbed during the control process, depending on controller design. In general, simple scalar controllers will affect machine flux, and thus, torque stability. Independent torque and flux control design is also possible, but may grow too complicated compared with a field oriented controller [25].

Since the principle of field orientation was applied to both induction and synchronous machines more than two decades ago [26], the use of induction machines in drive systems has increased significantly. Field orientation, which is often called vector control, is a torque control algorithm where electric torque can be controlled separately while machine flux remains unchanged [27-33]. Based on this principle, many control objectives, such as speed or position control can be achieved via separate control loops placed outside of the field oriented controller [25]. Control methods such as model reference control [34-37], variable structure control [38] or self-tuning control [25] can be applied.

The direct torque control method (DTC) combines machine torque and inverter control into one control unit. In a three phase, six switch inverter configuration there are only six valid switch combinations. Correct sequencing of these combinations can achieve torque regulation while maintaining machine flux level [39-41].

Model reference adaptive control (MRAC) can be used on line to estimate rotor resistance, which is the critical parameter for indirect field oriented controllers [37]. It
can also be used to form an outer speed control loop with a field oriented controller [34,35]

The field oriented control concept can also be applied to variable speed generator systems [42]. Instead of controlling speed in a drive system, torque and flux commands can be used to control active and reactive power flows through the generator system. This method assures fast response and improves system stability.

1.3 BDFM Model Representations

As mentioned previously, dq models and electrical parameter estimation techniques are needed before any BDFM dynamic controller design can proceed. Fig. 1.1 illustrates that the power winding is directly connected to the grid and only the control winding quantities are controllable. Depending on nature of the control winding excitation, there are two dq model representations for BDFM.

The BDFM dynamic model was developed on the rotor reference frame [5]. The voltage source model is shown in Eqn. (1.2), where $V_{qr} = V_{dr} = 0$ and control is applied via $V_{q2}$ and $V_{d2}$. All variables in this thesis are defined in the Nomenclature.

$$
\begin{bmatrix}
V_{q6} \\
V_{d6} \\
V_{q2} \\
V_{d2} \\
V_{qr} \\
V_{dr}
\end{bmatrix} =
\begin{bmatrix}
r_6 + L_{s6}p & 3L_{s6}\omega_r & 0 & 0 & M_6p & 3M_6\omega_r \\
-3L_{s6}\omega_r & r_6 + L_{s6}p & 0 & 0 & -3M_6\omega_r & M_6p \\
0 & 0 & r_2 + L_{s2}p & L_{s2}\omega_r & -M_2p & M_2\omega_r \\
0 & 0 & -L_{s2}\omega_r & r_2 + L_{s2}p & M_2\omega_r & M_2p \\
M_6p & 0 & -M_6p & 0 & r_r + L_r p & 0 \\
0 & M_6p & 0 & M_2p & 0 & r_r + L_r p
\end{bmatrix}
\begin{bmatrix}
i_{q6} \\
i_{d6} \\
i_{q2} \\
i_{d2} \\
i_{qr} \\
i_{dr}
\end{bmatrix}
$$

(1.2)
If the control winding is fed with a current source, a current forced model can be derived from Eqn. (1.2) [18] where the control inputs are $i_{q2}$ and $i_{d2}$:

\[
\begin{bmatrix}
V_{q6} \\
V_{d6} \\
M_{2pi_{q2}} \\
-M_{2pi_{d2}}
\end{bmatrix} =
\begin{bmatrix}
 r_6 + L_{so} p & 3L_{so} \omega_r & M_p & 3M_6 \omega_r \\
-3L_{so} \omega_r & r_6 + L_{so} p & -3M_6 \omega_r & M_p \\
M_6 p & 0 & r_r + L_{ip} p & 0 \\
0 & M_6 p & 0 & r_r + L_{ip} p
\end{bmatrix}
\begin{bmatrix}
 i_{q6} \\
i_{d6} \\
i_{qr} \\
i_{dr}
\end{bmatrix}
\]  
\( (1.3) \)

The BDFM electric torque can be expressed as

\[
T_e = 3M_6 (i_{q6} i_{dr} - i_{d6} i_{qr}) + M_2 (i_{q2} i_{dr} + i_{d2} i_{qr})
\]  
\( (1.4) \)

The system mechanical equations are

\[
\frac{d\theta_r}{dt} = \omega_r \]  
\( (1.5) \)

\[
J \frac{d\omega_r}{dt} = T_e - T_L - K_d \omega_r \]  
\( (1.6) \)

where $J$ is the inertia and $K_d$ is the viscous damping coefficient.

An inverter is used to control the excitation of the BDFM control winding as shown in Fig. 1.1. It should be noted that, if a controller is designed based on the BDFM voltage source model, the inverter should be controlled in the voltage source mode. The current source mode is needed if the controller design is based on the current source model.

1.4 Thesis Outline

As indicated by the literature survey, unique strategies and algorithms need to be formulated for dynamic control of the BDFM because of its specific structure and different model representation when compared with conventional machines. The
objective of this thesis is to develop a simple control algorithm which can be implemented on a laboratory prototype for evaluation. Different methods have been evaluated for this purpose.

Chapter 2 presents simple scalar control methods for the BDFM. Experimental and simulation results are given and evaluated. An MRAC controller is designed in chapter 3 based on a direct torque controller developed earlier [20,21]. Chapter 4 extends the rotor flux field oriented control method for conventional induction machines to the BDFM. A decoupled control algorithm is described. Chapter 5 evaluates experimental results for different control algorithms based on the rotor flux oriented control method. The experimental system used to verify the control algorithms is detailed in chapter 6. Finally, conclusions and recommendations are given in chapter 7.
Chapter 2

Scalar Control Methods For Brushless Doubly-Fed Machines

During synchronous steady state operation, the mechanical rotor frequency, \( f_{r,m} \), of brushless doubly-fed machines can be expressed as

\[
f_{r,m} = \frac{f_p \pm f_c}{P_p + P_c}
\]

(2.1)

where \( f_p \) and \( P_p \), \( f_c \) and \( P_c \) are power and control winding excitation frequencies and pole pair numbers, respectively. Since the power winding is connected to the fixed frequency grid, BDFM rotor speed corresponds directly to the control winding excitation frequency, \( f_c \). Therefore, scalar control methods can easily be applied by varying \( f_c \) in variable speed BDFM systems.

The control excitation magnitude can also be controlled to regulate a performance index, which may comprise any of the BDFM steady state performance characteristics, such as efficiency or power winding power factor.

In this chapter, following a comparison between two commonly used scalar control methods (constant volts/hertz and constant current controls), the constant current control method is applied to the BDFM. Simulation and experimental results are presented.
2.1 Constant Volts/Hertz Versus Constant Current Control

In a variable speed control system, it is desirable to maintain machine air gap flux to provide maximum torque control sensitivity with stator current, similar to a dc machine. There are two sources for air gap flux in the BDFM, one excited from the power winding and the other from the control winding. Because the power winding voltage is maintained by the grid, the air gap flux due to power winding excitation will be maintained constant over the entire operating range. As in an induction machine, the air gap flux due to control winding excitation is approximately related to the volts/hertz ratio (machine terminal voltage / excitation frequency) or the control current magnitude. Thus, for variable speed operation, it is desirable to maintain either a constant volts/hertz ratio or a constant control current magnitude.

The open loop volts/hertz and current control methods for the BDFM are illustrated in Fig. 2.1 and Fig. 2.2, respectively. For the volts/hertz control, the frequency signal, f, is multiplied with a constant gain, G, to obtain the magnitude signal V. In the current control case, the I signal is simply kept constant while f is varied.

![Figure 2.1 Constant Volts/Hertz Control](image1)

![Figure 2.2 Constant Current Control](image2)
At low frequency, a relatively high percentage of the applied voltage in the constant volts/hertz controller has to be used to compensate for the stator resistive voltage drop. Thus, an initial voltage boost is needed in order to maintain a constant air gap flux level over the entire speed range. Eqns. (1.2) and (1.3) represent the dynamic equations for the BDFM under voltage and current mode excitations. The electrical equations for current control are of lower order than those for voltage control. Control systems based on current control will be less complex than the equivalent voltage control systems. Therefore, only current control schemes will be discussed.

It is important for stable BDFM operation that the electric torque remains within the maximum torque envelope at all operating points. Fig. 2.3 shows the simulated maximum torque envelope for the laboratory BDFM prototype with constant current

![Figure 2.3 Maximum Torque Envelope for Constant Current Control.](image-url)
control method. The prototype has 6-pole power and 2-pole control windings. Fig. 2.3 displays a zero torque production point at 1200 l/min, which corresponds to the synchronous speed of the 6-pole power winding. Two experimental points are also obtained for this type of operation. The laboratory BDFM prototype is described in section 6.1.

2.2 Current Mode Scalar Control

2.2.1 Open Loop Control

The open loop current control scheme for the BDFM is shown in Fig. 2.2. A converter capable of accepting both I and f commands excites the control winding of the BDFM. Eqn. (2.1) shows that the BDFM rotor speed can be controlled via the frequency (f) command. Other performance parameters, such as efficiency or power winding power factor, can be regulated via the current (I) command. The laboratory experimental set up for the BDFM open loop control is the same as described in section 3.4 where both I and f commands can be issued from a personal computer (PC).

Figs. 2.4 to 2.6 illustrate simulation and experimental results for the BDFM open loop speed response to a step change in frequency command from 5 to 8 Hz. The power winding power factor response to a step current command (from 2 to 5A) is shown in Figs. 2.7 through 2.9. The results show that both responses are oscillatory. The simulation results agree with the experimental results. Power factor control is only an example. The control winding current magnitude can also be used to control other performance parameters, such as efficiency or reactive power.
Figure 2.4 Open Loop Speed Response to a Step Change in Control Frequency Command (Simulation).

Figure 2.5 Control Current Waveform for a Step Change in Frequency Command (Simulation).
Figure 2.6 Open Loop Response to a Step Change in Control Frequency Command (Experiment). Upper: Shaft Speed; Lower: Control Current (2A/div).

Figure 2.7 Open Loop Power Winding Power Factor Response to a Step Change in Control Current Magnitude Command (Simulation).
Figure 2.8 Control Current Waveform for a Step Change in Magnitude Command (Simulation).

Figure 2.9 Open Loop Response to a Step Change in Control Current Command (Experiment). Upper: Power Winding Power Factor; Lower: Control Current (4A/div).
2.2.2 Closed Loop Scalar Control

As discussed previously, it is desirable to operate the BDFM in synchronism where the rotor speed is given as in Eqn. (2.1). Scalar control of speed, torque and other performance characteristics are feasible via controlling the magnitude and frequency of control winding excitation [6,7]. The block diagram shown in Fig. 2.10 illustrates two primary loops to control the BDFM speed and power winding power factor, respectively. The experimental set up described in section 3.4 is also used to implement the closed loop scalar control. It is obvious that although a DSP56001 microprocessor is used here, closed loop control computations only require a low cost microprocessor, such as 8051 or 80196, when practical systems are implemented.

\[ \text{Figure 2.10 Closed Loop Scalar Control Block Diagram.} \]
2.2.2.1 Speed Control

The proportional-integral control loop shown in Fig. 2.10 is used to alleviate the stability problems during acceleration and deceleration illustrated by Figs. 2.5 through 2.10 in the previous section. In this case, the microprocessor uses rotor speed information and Eqn. (2.1) to calculate the appropriate converter frequency \( f_2 \). The simple PI-controller improves the stability margin of the drive system and allows for a smoother speed transfer. This is illustrated in Figs. 2.11 through 2.13 for a step change in the speed command from 780 to 825 \text{1/min}.

![Figure 2.11](image)

**Figure 2.11** Closed Loop Speed Response to a Step Change in Speed Command (Simulation).
Figure 2.12 Control Current Waveform for a Step Change in Speed Command (Simulation).

Figure 2.13 Closed Loop Speed Response to a Step Change in Speed Command (Experiment). Upper: Shaft Speed; Lower: Control Current (4A/div).
This simple control algorithm is augmented by a synchronization loop, which further assures stability. If synchronism is about to be lost, the calculated value for $f_{2r}$ will deviate significantly from the converter input frequency $f_{2*}$. If this difference persists, the control algorithm based on Eqn. (2.1) no longer guarantees successful speed control. In this case, the DSP adjusts $f_{2*}$ close to $f_{2r}$ and updates the PI parameters. This assures synchronous operation and is followed by establishment of the desired speed. The resulting control philosophy is easy to implement on low cost microprocessors and is quite acceptable for low performance industrial applications such as pump or fan drives [10, 13].

2.2.2.2 Power Winding Power Factor Control

The converter output voltage or current magnitude can be used to control the power winding power factor. As illustrated in Fig. 2.10, only rms quantities associated with the power winding are measured. The remaining quantities needed in the power factor control system are estimated based on the BDFM steady state equivalent circuit shown in Fig. 2.14; where $p$ refers to the power winding (six poles in the experimental prototype) and $c$ refers to the control winding (two poles). Using a desired six pole power factor $\cos\phi^*$, two equations can be derived from the stator equivalent circuit (subscripts $r$ and $i$ denote real and imaginary parts, respectively):

$$V_6 = r_6 I_6 \cos\phi^* + X_{se} I_6 \sin\phi^* - X_{me} I_i$$  \hspace{1cm} (2.2)
Figure 2.14 BDFM Steady State Equivalent Circuit.

\[ 0 = X_{so}I_6\cos\phi_6^* - r_0I_6\sin\phi_6^* + X_{m6}I_r \]  

where \( V_6 \) and \( I_6 \) are determined from measurements and \( X_{so} = X_{sl} + X_{m6} \). \( I_r \) and \( I_\tau \) are the two rotor current components in the phasor diagram shown in Fig. 2.15.

Eqns. (2.2) and (2.3) can be solved for \( I_r \) and \( I_\tau \) as

\[
I_\tau = \frac{-X_{so}I_6\cos\phi_6^* + r_0I_6\sin\phi_6^*}{X_{m6}} 
\]

\[
I_r = \frac{r_0I_6\cos\phi_6^* + X_{so}I_6\sin\phi_6^* - V_6}{X_{m6}} 
\]

Since the converter is operated in the controlled current mode, only the rotor equivalent circuit is needed to estimate the desired \( I_2 \).

\[ 0 = X_{m6}I_6\cos\phi_6^* - X_{m2}I_2 + \frac{r_rI_r}{s_r} + X_{m2}I_2 \]  

\[ 0 = X_{m6}I_6\sin\phi_6^* + X_{m2}I_2 + \frac{r_rI_r}{s_r} - X_{m2}I_2 \]

where \( X_r = X_{tr} + X_{mr}, s_r = (\omega_6 - 3\omega_0)/\omega_6 \) and \( I_2r \) and \( I_2l \) are defined in Fig. 2.15.
Figure 2.15 BDFM Steady State Phasor Diagram.

From Eqns. (2.6) and (2.7) and Fig. 2.14 it is obvious that

\[
I_{2r} = \frac{X_m I_0 \cos \phi^* + \frac{r}{s} I_{ri} + X_r I_{rr}}{X_{m2}} \quad (2.8)
\]

\[
I_{2i} = \frac{-X_m I_0 \sin \phi^* - \frac{r}{s} I_{ri} + X_r I_{ri}}{X_{m2}} \quad (2.9)
\]

and the resulting control current magnitude

\[
I_2 = \sqrt{I_{2r}^2 + I_{2i}^2} \quad (2.10)
\]

As illustrated in Fig. 2.10, the above calculations are implemented in the microprocessor using estimated machine parameters [15]. A PI loop translates the power factor error into a current control signal. Fig. 2.16 to 2.18 show the resulting system response for a step change in power factor command from 0.625 to 0.375. As shown, the power factor error is quickly brought to zero. In industrial applications, power factor can
be commanded to a constant value and the controller will update the control current magnitude in the presence of operating condition changes.

**Figure 2.16** Closed Loop Power Factor Error Response to a Step Command (Simulation).

**Figure 2.17** Control Current Waveform for a Step Change in Power Factor Command (Simulation).
Figure 2.18 Closed Loop Response to a Step Change in Power Winding Power Factor Command (Experiment).
Upper: Power Factor Error; Lower: Control Current (4A/div).

It is evident that simple control algorithms based on steady state system equivalents can enhance the BDFM mechanical and electrical performance for relatively undemanding industrial applications such as pumps and fans. However, the speed response under scalar closed loop control is still very slow for a large step speed command. If fast dynamic control is desired, scalar control methods are not sufficient. Alternate methodologies have to be developed.
Chapter 3

Model Reference Adaptive Speed Control For Brushless Doubly-Fed Machines

As stated in Chapter 1, brushless doubly-fed machines are suitable for variable speed applications. Although open loop or scalar controls can easily control steady state performance characteristics, such as efficiency or power factor, by controlling either voltage or current magnitude, they fail in providing good dynamic performance. Alternative methods are needed for improved dynamic response. The proposed methods should be robust with respect to both mechanical and electrical system parameters.

This chapter will discuss a model reference adaptive speed control (MRAC) based on a direct torque control algorithm developed for BDFM.

3.1 Direct Torque Control Algorithm [20, 21]

The BDFM system shown in Fig. 1.1 can be represented by the dq equivalent circuit illustrated in Fig. 3.1, with the voltage, current and torque relationships in the rotor reference frame outlined in Eqns. (1.2) through (1.4) [5,18]. In Fig. 3.1 the only control variables are the terminal voltages on the control winding, i.e., $V_{q2}$ and $V_{d2}$.

Traditional self control systems for conventional induction machines are based on the fact that each particular inverter voltage vector is known to produce a decrease or increase in torque and flux. This does not involve machine parameters and a sliding mode control based on torque and flux error can be formulated in a relatively straightforward manner. A direct formulation is not available in the BDFM, since the
Figure 3.1 Two-Axis Model for BDFM
voltage vectors driving torque and flux in the correct direction are not known. However, one direct torque control method can be developed as follows. The rates of change in electric torque and control winding flux can be expressed as in Eqns. (3.1) and (3.2) [21].

\[ T_e = 3(i_{q0}, \lambda_{q0} + i_{d0}, \lambda_{d0} - i_{q0}, \lambda_{d0} - i_{d0}, \lambda_{q0}) \]

\[ + (i_{q2}, \lambda_{q2} + i_{d2}, \lambda_{d2} - i_{q2}, \lambda_{q2} - i_{d2}, \lambda_{d2}) \]

\[ \lambda_{q2}, \lambda_{q2} + \lambda_{d2}, \lambda_{d2} = \lambda_{2q}^2 + \lambda_{2d}^2 \]  

Although Eqns. (1.2), together with the mechanical equations (1.5) and (1.6) form an eighth order system, measurement and prediction of both power and control winding terminal voltages and currents allow a combination of the six electrical equations with Eqns. (3.1) and (3.2). This yields [21]

\[ T_e^* = \beta_1 V_{q2} + \beta_2 V_{d2} + \beta_3 \]  

\[ \lambda_2^* = \gamma_1 V_{q2} + \gamma_2 V_{d2} + \gamma_3 \]

where the \( \beta_i \) and \( \gamma_i \) are functions of machine parameters and predicted or measured electrical quantities. Eqns. (3.3) and (3.4) can be manipulated to be of a second order linear relationship between the control inputs and the rates of change in torque and control winding flux:

\[ \begin{bmatrix} V_{q2} \\ V_{d2} \end{bmatrix} = \frac{1}{\beta_1 \gamma_2 - \beta_2 \gamma_1} \begin{bmatrix} \gamma_2 & -\beta_2 \\ -\gamma_1 & \beta_1 \end{bmatrix} \begin{bmatrix} T_e^* \\ \lambda_2^* \end{bmatrix} \]  

The block diagram of the controller is shown in Fig. 3.2. An estimator accepts all eight machine terminal quantities and shaft information to calculate the machine electric torque at the moment of data sampling. The machine parameters are estimated off line using frequency domain estimation methods [16]. The control algorithm implemented in Eqn. (3.5) will take a period of time to finish. However, the control output \( V_{q2} \) and \( V_{d2} \) are
only suitable at the moment of data sampling. A predictor has to be designed to give correct machine terminal quantities at the time \( V_{q2} \) and \( V_{d2} \) are applied to the control winding. An error trap loop is also included to prevent the control matrix in Eqn. (3.5) from going singular.

As reported in [20] and [21], the direct torque control in Fig. 3.2 shows significant improvement of the BDFM dynamic performance over the open loop case. The method relies on machine parameters and terminal quantities for the calculations in the control loop. Robustness with respect to electrical parameter variations has also been reported. As shown in Fig. 3.2, the development of the direct torque control so far has only emphasized the inner electrical control loop. The outer speed and flux loops are simple PI-regulators. A model reference adaptive speed loop to improve mechanical system robustness and eliminate the over- and undershoots associated with the PI-regulators in Fig. 3.2 is investigated here [11].

![Figure 3.2 Direct Torque Control Block Diagram (from Ref [20])](image)
### 3.2 Model Reference Adaptive Speed Control Development

The mathematical process of model reference adaptive control algorithm with a high order machine model needs extensive computations and hence slows down the response, denying the desired dynamic performance. Present day microprocessors are not capable of economically implementing these strategies in real time. While the electrical performance of the BDFM is described by a set of six differential equations, the structure of direct torque control is only of second order. Considering the capabilities of modern microprocessors for real time implementation, a first order model reference adaptive controller speed loop is chosen. The block diagram for the model reference adaptive controller is shown in Fig. 3.3, where it replaces the PI-speed loop located in the upper left corner of Fig. 3.2.

![Figure 3.3 MRAC Block Diagram](image-url)
Assuming that all electrical transients in BDFM under direct torque control are much faster than the mechanical speed transient, the BDFM system can be represented by a first order speed state space equation

\[
\frac{d\omega_r}{dt} = A_p\omega_r - B_p\Delta\omega_r^*
\] (3.6)

where \(\Delta\omega_r^*\) is the command input to the BDFM plant system. \(A_p\) and \(B_p\) are coefficients which may vary with load and environmental situations.

The reference model is chosen as [36]

\[
\frac{d\omega_m}{dt} = A_m\omega_m + B_m\omega_r^*
\] (3.7)

where the \(\omega_m\) is the model output. \(A_m\) and \(B_m\) are coefficients which can be adjusted to achieve the desired system performance, as the system output will follow the reference model output in this case.

The block diagram in Fig. 3.3 shows the system error as

\[ e = \omega_r - \omega_m \] (3.8)

and the input to the BDFM plant is

\[ \Delta\omega_r^* = K_f\omega_r^* - K_b\omega_r \] (3.9)

Taking the derivative with respect to time in Eqn. (3.8) and combining with Eqns. (3.6) and (3.7) yields

\[
\frac{de}{dt} = \frac{d\omega_r}{dt} - \frac{d\omega_m}{dt}
\]

\[
= A_m e + (A_p - A_m - B_p K_b)\omega_r + (B_p K_f - B_m)\omega_r^*
\]

In order to derive the adaptive mechanism of \(K_f\) and \(K_b\), the following Lyapunov function is introduced
\[ V(e, K_f, K_b) = \frac{1}{2} [e^2 + \frac{1}{B_p \gamma} (B_p K_b + A_m - A_p)^2 \]
\[ + \frac{1}{B_p \gamma} (B_p K_f - B_m)^2] \]

where \( \gamma \) is a constant such that \( B_p \gamma > 0 \). The Lyapunov function \( V(e, K_f, K_b) \) is positive definite in a global sense. In order for the model reference adaptive controller to be stable, the following condition must be met

\[ \frac{dV(e, K_f, K_b)}{dt} < 0 \quad \forall e \neq 0 \] (3.12)

\[ \frac{dV(e, K_f, K_b)}{dt} = 0 \quad \forall e = 0 \] (3.13)

From Eqns. (3.10) and (3.11)

\[ \frac{dV(e, K_f, K_b)}{dt} = A_m e^2 + \frac{1}{\gamma} (A_m - A_p + B_p K_b) \frac{dK_b}{dt} - \gamma e \omega_r \]
\[ + \frac{1}{\gamma} (B_p K_f - B_m) \frac{dK_f}{dt} + \gamma e \omega_r^* \] (3.14)

It is obvious that if the following conditions are satisfied, then \( \frac{dV(e, K_f, K_b)}{dt} \) will satisfy Eqns. (3.12) and (3.13) and ensure the stability of the model reference adaptive controller.

\[ A_m < 0 \] (3.15)
\[ B_p \gamma > 0 \] (3.16)
\[ \frac{dK_b}{dt} = \gamma e \omega_r \] (3.17)
\[ \frac{dK_f}{dt} = -\gamma e \omega_r^* \] (3.18)

Equations (3.15) and (3.16) will be satisfied by properly selecting parameters \( A_m \) and \( \gamma \). The adaptive law of the model reference adaptive controller is described by Eqns. (3.17) and (3.18).
3.3 Simulation Results

As developed, the model reference adaptive control algorithm can improve system dynamic performance for changes in mechanical parameters such as inertia and load torque. Simulation is carried out to prove the concept where the model reference and direct torque controls are updated at 1ms intervals.

The response of the model reference adaptive controller as compared with the PI-based controller of Fig. 3.2 is shown in Fig. 3.4 for a step change in speed command (+100 1/min) while maintaining a constant load torque of 8 Nm. In order to allow for a valid comparison, the model reference adaptive controller parameters are tuned such that it has the same time response as the PI-based controller. As illustrated, the system response is improved. The use of the measured speed signal in place of the predicted speed value in the model reference adaptive control yields oscillations at the beginning.

Speed response for a doubling of system inertia is shown in Fig. 3.5. The model parameters $A_m$ and $B_m$ are not changed. Under model reference adaptive control, the BDFM exhibits the same time response as illustrated for the undisturbed case in Fig. 3.4. For the PI-based controller, however, the time response is doubled.

Fig. 3.6 shows the response when the load torque is increased by 25%. The model reference adaptive control response shows less peak-to-peak variation (about 0.5 r/min) than the PI-controller. The time responses are similar as both control systems are tuned for identical inertia values.
Figure 3.4 System Response to a Step Change in Speed Command
(a): PI-Controller; (b): MRAC
Figure 3.5 System Response to a Step Change in Inertia and Speed Command
(a): PI-Controller (inertia doubled at t = 2.75 s)
(b): MRAC (inertia doubled at t = 5.25 s)
Figure 3.6 System Response to a Step Change in Load Torque
(a): PI-Controller; (b): MRAC
As illustrated, the model reference adaptive control implemented is very effective with regard to parameter variations in the BDFM mechanical system. The direct torque control itself is quite robust for electrical parameter variations as reported in [20] and [21]. Since the model reference adaptive control system is built around the inner direct torque controller, it is also robust with respect to electrical parameter variations. This is illustrated by the model reference adaptive controller speed response in Fig. 3.7, where rotor resistance in the prediction algorithm of the direct torque controller is underestimated by 20%. Satisfactory dynamic performance can still be maintained. Here the criteria for testing system robustness with respect to electrical parameters and inertia

**Figure 3.7** MRAC Response to a Step Change in Speed Command with 20% Rotor Resistance Error in the Direct Torque Controller
are different. A test of robustness with respect to electrical parameter variation determines the boundary of parameter error range where the control system becomes unstable. However, robustness against inertia means how well the control system maintains acceptable dynamic performance as required by most industry applications when the system inertia is changed. The robustness of the model reference adaptive controller against inertia is clearly superior to a PI regulator as demonstrated in Fig. 3.5.

3.4 Experimental Set Up and Remarks

Fig. 3.8 shows the experimental set up intended for laboratory implementation of the direct torque control and model reference adaptive speed control. The set up is equipped with four instantaneous current and voltage transducers with 5kHz bandwidth. Since both BDFM stator windings are Y-connected, only two currents and voltages are measured. The set up also has a 12-bit absolute position encoder which provides shaft position information via a position interface card. All measurements can be sampled and brought into the microcontroller within 7μs.

The DSP56001 microprocessor runs on a 27MHz clock and is equipped with some powerful functions and one cycle instructions. After collecting data from A/D and position interface boards, the DSP56001 performs the following steps in a control cycle:

1. Transform measured quantities from the abc domain to the dq reference frame
2. Direct torque control and MRAC algorithm calculations
3. Transform the control quantities from the dq to the abc reference frame
Three high speed D/A boards convert three digital control signals from DSP56001 to appropriate analog signals which are acceptable to the series resonant power converter used in the experimental setup.

After hardware installation and debugging, the direct torque control algorithm is implemented on the DSP56001 microprocessor. It is found that one control cycle takes about 85ms, which is far slower than the 1ms update period assumed in simulation. The slow control cycle is due to the intensive calculations required by the estimator and the predictor in Fig. 3.2, both of which are derived from the sixth order BDFM electrical differential equations. Because the DSP56001 is a fixed point processor, the floating
point operations and trigonometric functions associated with the control algorithm slow down the calculation speed to about one third of that achievable with floating point microprocessors.

The main function of the estimator in Fig. 3.2 is to obtain the electric torque at the sampling moment. Due to the long control update time (85ms), the calculated value of BDFM electric torque does not represent the actual machine torque. Consequently, the predictor inside the direct torque controller is not able to correctly predict values and implementation of the DTC algorithm becomes impossible. The direct torque control algorithm for the BDFM is still too complex to be economically implemented on modern microprocessors. This has lead to the investigation of alternative control strategies and microprocessors as discussed in the following chapters.
Chapter 4

Synchronous Reference Frame Model and Control Developments

Although the simulations in Chapter 3 show that direct torque control (DTC) and model reference adaptive control (MRAC) improve BDFM dynamic performance, they are too complicated to implement because their control algorithms need to calculate the sixth order BDFM differential equation model twice to complete a control cycle. Simpler and faster control algorithms are desirable.

As indicated in Chapter 1, the BDFM can be represented by either a voltage source or a current forced model depending on the nature of the control winding excitation. Chapter 2 shows that the current forced model is of lower order than the voltage source model, which will simplify the dynamic control design for the BDFM. Hence, current excitation is chosen for the control developments discussed here.

By analyzing DTC and MRAC methods in Chapter 3, there are two reasons why they need more microprocessor processing time. The first reason is that both DTC and MRAC use the BDFM voltage source model which is four orders higher than a conventional induction machine model. The BDFM consists of a power winding and a control winding. Since the power winding is directly connected to the power grid, it is not controllable. In order to avoid high order model manipulation, the whole BDFM system has to be broken down into subsystems, which will be discussed in this chapter. The second reason is that both DTC and MRAC use a BDFM model based on the rotor reference frame. On this frame, all BDFM dq variables appear as sinusoids during steady state operation. It is well known that sinusoidal quantities need three elements to be
solely determined and hence are much more complicated to be controlled than any DC quantity. It is desirable to introduce different reference frames to transform such rotor reference frame quantities to their DC values for ease of subsequent controller design.

In this chapter, the BDFM current forced model is first divided into two subsystems representing power and control windings. Different synchronous frames are then applied to these two subsystems to obtain DC value variables. A control algorithm based on control winding rotor flux orientation is then derived from this approach and its results are presented.

### 4.1 Subsystem Classifications

When BDFM control winding is excited by a current source, the machine model can be represented as in Eqn. (1.3) on the rotor reference frame. For ease of reference, Eqn. (1.3) is repeated here as Eqn. (4.1).

\[
\begin{bmatrix}
V_{q6} \\
V_{d6} \\
M_{2p}i_{q2} \\
-M_{2p}i_{d2}
\end{bmatrix} =
\begin{bmatrix}
r_6 + L_{s6}p & 3L_{s6} \omega_r & M_6p & 3M_6 \omega_r \\
-3L_{s6} \omega_r & r_6 + L_{s6}p & -3M_6 \omega_r & M_6p \\
M_6p & 0 & r_r + L_r p & 0 \\
0 & M_6p & 0 & r_r + L_r p
\end{bmatrix}
\begin{bmatrix}
i_{q6} \\
i_{d6} \\
i_{qf} \\
i_{df}
\end{bmatrix}
\]  

(4.1)

The electric torque is expressed as

\[
T_e = 3M_6(i_{q6}i_{dr} - i_{d6}i_{qf}) + M_2(i_{q2}i_{dr} + i_{d2}i_{qf})
\]

(4.2)

During normal operation, the BDFM power winding is connected to the fixed frequency (50 or 60 Hz) utility grid. At this frequency, the power winding magnetizing
Impedance is much higher than the winding resistance and the power winding resistive voltage drop is negligible. Due to the constant grid voltage, the power winding stator flux contribution is approximated as constant.

Defining power winding stator fluxes as

\[ \psi_{q6} = L_{s6}i_{q6} + M_6i_{qr} \]  
\[ \psi_{d6} = L_{s6}i_{d6} + M_6i_{dr} \]

the power winding equations in the current forced model (4.1) can be expressed as

\[
\begin{bmatrix}
V_{q6} \\
V_{d6}
\end{bmatrix} = p \begin{bmatrix}
\psi_{q6} \\
\psi_{d6}
\end{bmatrix} + 3\omega_r \begin{bmatrix}
\psi_{q6} \\
-\psi_{q6}
\end{bmatrix}
\]

In terms of power winding stator fluxes and control winding currents, the rotor equations are:

\[
\left[ r_r + (L_r - \frac{M_6^2}{L_{s6}})p \right] i_{qr} = M_2pi_{q2} - \frac{M_6^2p}{L_{s6}} \psi_{q6}
\]  
\[
\left[ r_r + (L_r - \frac{M_6^2}{L_{s6}})p \right] i_{dr} = -M_2pi_{d2} - \frac{M_6}{L_{s6}} \psi_{d6}
\]

Terms on the right hand side of Eqns. (4.6) and (4.7) can be treated as inputs (forcing functions), due to power winding flux \( \psi_{q6} \) and control winding current \( i_{q2} \). As discussed in Chapter 1, on the rotor reference frame these terms have different frequencies in the induction mode of operation. During synchronous operation, these frequencies are identical, forcing rotor currents of a single frequency.

Neglecting rotor saturation, Eqns. (4.6) and (4.7) can be separated into two subsystems. The rotor currents can be expressed as
\[ i_{qr} = i_{qr6} + i_{qr2} \quad (4.8) \]
\[ i_{dr} = i_{dr6} - i_{dr2} \quad (4.9) \]

where \( i_{qr6} \) and \( i_{dr6} \) correspond to power winding and \( i_{qr2} \) and \( i_{dr2} \) correspond to control winding excitation forcing functions. Eqns. (4.6) and (4.7) can now be rewritten as

\[
\begin{bmatrix}
  r_r + (L_r - \frac{M_6^2}{L_{s6}}) p \\
\end{bmatrix} i_{qr6} = -\frac{M_6}{L_{s6}} p \psi_{q6} 
\]

\[
\begin{bmatrix}
  r_r + (L_r - \frac{M_6^2}{L_{s6}}) p \\
\end{bmatrix} i_{dr6} = -\frac{M_6}{L_{s6}} p \psi_{d6} 
\]

for the power winding contribution and

\[
\begin{bmatrix}
  r_r + (L_r - \frac{M_6^2}{L_{s6}}) p \\
\end{bmatrix} i_{qr2} = M_2 p i_{q2} 
\]

\[
\begin{bmatrix}
  r_r + (L_r - \frac{M_6^2}{L_{s6}}) p \\
\end{bmatrix} i_{dr2} = M_2 p i_{d2} 
\]

for the control winding contribution. With these modifications, the torque equation (4.2) becomes

\[
T_e = 3 \frac{M_6}{L_{s6}} (\psi_{q6} i_{dr6} - \psi_{d6} i_{qr6}) + M_2 (i_{dr6} i_{q2} + i_{qr6} i_{d2}) 
\]

\[-3 \frac{M_6}{L_{s6}} (\psi_{q6} i_{dr2} + \psi_{d6} i_{q2}) + M_2 (i_{qr2} i_{d2} - i_{dr2} i_{q2}) \]

In all, the BDFM model of Eqns. (4.1) and (4.2) is now expressed by Eqns. (4.5), (4.10), (4.11), (4.12), (4.13) and (4.14). Eqns. (4.5), (4.10) and (4.11) are referred to as the power winding subsystem and Eqns. (4.12) and (4.13) form the control winding subsystem.
4.2 Synchronous Reference Frame Model

The BDFM model equations derived in the previous section are based on the rotor reference frame. As stated before, all stator electrical quantities are sinusoidal during steady state. Sinusoidal quantities are more difficult to control than DC quantities and thus inconvenient for controller design. Synchronous reference frames are desirable for transformation of the electrical quantities to yield constants for the steady state. It is evident that Eqns. (4.5), (4.10) and (4.11) (power winding subsystem) are only related to power winding excitation, while Eqns. (4.12) and (4.13) (control winding subsystem) are due to control winding excitation. Apparently all electrical subsystem quantities will become constant during steady state operation when transformed onto their appropriate synchronous reference frames. Thus the complexity of controller design will be reduced.

Fig. 4.1 illustrates the resulting transformations. All quantities on the power winding synchronous reference frame are assigned a superscript "6e" and all control winding synchronous reference frame values are represented by a superscript "2e".

The transformation between the rotor and the power winding reference frames is expressed as

\[
\begin{bmatrix}
V_{q6} \\
V_{d6}
\end{bmatrix}
= \begin{bmatrix}
\cos(\theta_{6e}) & \sin(\theta_{6e}) \\
-sin(\theta_{6e}) & \cos(\theta_{6e})
\end{bmatrix}
\begin{bmatrix}
V_{q6}^{6e} \\
V_{d6}^{6e}
\end{bmatrix}
\]

\[ (4.15) \]

\[
\begin{bmatrix}
\psi_{q6} \\
\psi_{d6}
\end{bmatrix}
= \begin{bmatrix}
\cos(\theta_{6e}) & \sin(\theta_{6e}) \\
-sin(\theta_{6e}) & \cos(\theta_{6e})
\end{bmatrix}
\begin{bmatrix}
\psi_{q6}^{6e} \\
\psi_{d6}^{6e}
\end{bmatrix}
\]

\[ (4.16) \]
Figure 4.1 Rotor and Stator Winding Synchronous Reference Frames

\[
\begin{bmatrix}
i_{q6} \\
i_{d6}
\end{bmatrix} = \begin{bmatrix}
\cos(\theta_{6e}) & \sin(\theta_{6e}) \\
-\sin(\theta_{6e}) & \cos(\theta_{6e})
\end{bmatrix} \begin{bmatrix}
i_{q6e} \\
i_{d6e}
\end{bmatrix}
\]  \hfill (4.17)

\[
\begin{bmatrix}
i_{q6} \\
i_{d6}
\end{bmatrix} = \begin{bmatrix}
\cos(\theta_{6e}) & \sin(\theta_{6e}) \\
-\sin(\theta_{6e}) & \cos(\theta_{6e})
\end{bmatrix} \begin{bmatrix}
i_{q6e} \\
i_{d6e}
\end{bmatrix}
\]  \hfill (4.18)

where \( \theta_{6e} = \int_0^t \omega_6 dt - 3\theta_r = \theta_6 - 3\theta_r \). \( \theta_6 \) defines the power winding voltages in abc domain as
\[
\begin{bmatrix}
V_{6a} \\
V_{6b} \\
V_{6c}
\end{bmatrix} = \sqrt{2} V_6 \begin{bmatrix}
\cos \theta_6 \\
\cos(\theta_6 - 120^\circ) \\
\cos(\theta_6 + 120^\circ)
\end{bmatrix}
\]

\(V_6\) is the rms value of the phase voltages. The power winding subsystem is expressed on its synchronous frame as

\[
\begin{bmatrix}
V_{q6}^e \\
V_{d6}^e
\end{bmatrix} = p \begin{bmatrix}
\varphi_{q6}^e \\
\varphi_{d6}^e
\end{bmatrix} + \omega_6 \begin{bmatrix}
\psi_{d6}^e \\
-\psi_{q6}^e
\end{bmatrix}
\]

\[
(L_r - \frac{M_6}{L_{s6}}) \begin{bmatrix}
i_{q6}^e \\
i_{d6}^e
\end{bmatrix} + r_{ir} \begin{bmatrix}
i_{q6}^e \\
i_{d6}^e
\end{bmatrix} + (\omega_6 - 3\omega_r)(L_r - \frac{M_6^2}{L_{s6}^2}) \begin{bmatrix}
i_{q6}^e \\
i_{d6}^e
\end{bmatrix}
\]

\[
= -\frac{M_6}{L_{s6}} p \begin{bmatrix}
\psi_{q6}^e \\
\psi_{d6}^e
\end{bmatrix} - (\omega_6 - 3\omega_r) \frac{M_6}{L_{s6}} \begin{bmatrix}
\psi_{d6}^e \\
-\psi_{q6}^e
\end{bmatrix}
\]

If the power winding synchronous frame d-axis is aligned with the total winding flux, i.e.

\[
\begin{bmatrix}
\psi_{q6}^e \\
\psi_{d6}^e
\end{bmatrix} = \begin{bmatrix}
\frac{3}{\sqrt{2}} \\
\frac{V_6}{\omega_6}
\end{bmatrix}
\]

where \(\sqrt{\frac{3}{2}}\) results from the transformation from 3 phase stationary frame to the dq synchronous frame [5]. The power winding equations (4.21) becomes
The transformation between rotor and control winding reference frames is expressed as

\[
\begin{bmatrix}
    i_{q2} \\
    i_{d2}
\end{bmatrix}
= \begin{bmatrix}
    \cos(\theta_2e) & \sin(\theta_2e) \\
    -\sin(\theta_2e) & \cos(\theta_2e)
\end{bmatrix}
\begin{bmatrix}
    i_{q2}^e \\
    i_{d2}^e
\end{bmatrix}
\] (4.24)

\[
\begin{bmatrix}
    i_{qr2} \\
    i_{dr2}
\end{bmatrix}
= \begin{bmatrix}
    \cos(\theta_2e) & \sin(\theta_2e) \\
    -\sin(\theta_2e) & \cos(\theta_2e)
\end{bmatrix}
\begin{bmatrix}
    i_{qr2}^e \\
    i_{dr2}^e
\end{bmatrix}
\] (4.25)

\[
\begin{bmatrix}
    \psi_{qr2}^e \\
    \psi_{dr2}^e
\end{bmatrix}
= \begin{bmatrix}
    \cos(\theta_2e) & \sin(\theta_2e) \\
    -\sin(\theta_2e) & \cos(\theta_2e)
\end{bmatrix}
\begin{bmatrix}
    \psi_{qr2}^e \\
    \psi_{dr2}^e
\end{bmatrix}
\] (4.26)

where \( \theta_2e = -\int_0^t \omega_2 dt - \theta_r = -\theta_2 - \theta_r \). \( \theta_r \) defines the control winding currents in abc domain as

\[
\begin{bmatrix}
    i_{2a} \\
    i_{2b} \\
    i_{2c}
\end{bmatrix}
= \sqrt{2} I_2 \begin{bmatrix}
    \cos \theta_2 \\
    \cos(\theta_2 + 120^\circ) \\
    \cos(\theta_2 - 120^\circ)
\end{bmatrix}
\] (4.27)

\( I_2 \) is the rms value of the phase currents. The control subsystem equations (4.12) and (4.13) now become
Using both synchronous reference frame transformations, the equation for electric torque (4.14) can now be written as

\[
T_e = 3 \frac{M_6}{L_{s6}} (-\psi_{q6}^e i_{q6} + M_2 (i_{q2}^e i_{d2}^e - i_{d2}^e i_{q2}^e)) \\
+ M_2 (-i_{q6}^e i_{q2}^e + i_{d6}^e i_{d2}^e) \sin(\theta_{d6} + \theta_{d2}) \\
+ M_2 (i_{q6}^e i_{d2}^e + i_{d6}^e i_{q2}^e) \cos(\theta_{d6} + \theta_{d2}) \\
- 3 \frac{M_6}{L_{s6}} (\psi_{d6}^e i_{q2}^e \cos(\theta_{d6} + \theta_{d2}) + \psi_{d6}^e i_{d2}^e \sin(\theta_{d6} + \theta_{d2}))
\]

(4.29)

4.3 A Rotor Flux Decoupled Control Algorithm

4.3.1 Power Winding Subsystem Predictor

The equations for the power winding quantities (4.22) and (4.23) clearly indicate that this subsystem is not directly influenced by the control winding quantities. By measuring the supply voltage to the power winding and the rotor shaft speed, a predictor can be designed for the power winding forcing function related rotor currents, i.e., \(i_{q6}^e\) and \(i_{d6}^e\) can be predicted using Eqn. (4.23).
### 4.3.2 Control Subsystem Flux Alignment

Defining control subsystem rotor flux quantities as

\[
\psi_{qr}^{2e} = (L_r - \frac{M_6^2}{L_{s6}})i_{qr}^{2e} - M_2^2i_{q2}^{2e}
\]  
(4.30)

\[
\psi_{dr}^{2e} = (L_r - \frac{M_6^2}{L_{s6}})i_{dr}^{2e} - M_2^2i_{d2}^{2e}
\]  
(4.31)

eqn. (4.28) can be re-written in terms of rotor flux as

\[
\begin{bmatrix}
\psi_{qr}^{2e} \\
\psi_{dr}^{2e}
\end{bmatrix} + \frac{r_r}{L_r - \frac{M_6^2}{L_{s6}}} \begin{bmatrix}
\psi_{qr}^{2e} \\
\psi_{dr}^{2e}
\end{bmatrix} - (\omega_2 + \omega_r) \begin{bmatrix}
\psi_{dr}^{2e} \\
-\psi_{qr}^{2e}
\end{bmatrix} + \frac{r_r M_2}{L_r - \frac{M_6^2}{L_{s6}}} i_{d2}^{2e} = 0
\]  
(4.32)

If the control reference frame d-axis is chosen such that it aligns with total control flux, we obtain

\[
\psi_{qr}^{2e} = 0
\]  
(4.33)

\[
\frac{d\psi_{qr}^{2e}}{dt} = 0
\]  
(4.34)

Consequently, Eqn. (4.32) can be simplified as

\[
\omega_2 = \frac{r_r M_2}{\psi_{dr}^{2e}(L_r - \frac{M_6^2}{L_{s6}})} i_{q2}^{2e} - \omega_r
\]  
(4.35)

\[
\begin{bmatrix}
p \psi_{dr}^{2e} + \frac{r_r}{L_r - \frac{M_6^2}{L_{s6}}} \psi_{dr}^{2e} + \frac{r_r M_2}{L_r - \frac{M_6^2}{L_{s6}}} i_{d2}^{2e} = 0
\end{bmatrix}
\]  
(4.36)
4.3.3 Electric Torque Expression and Decoupled Torque Control

Based on the assumptions made and the reference frame alignments discussed above, the electric torque is expressed from eqn. (4.29) as

\[
T_e = \frac{3M_6}{L_{s6}}(-\psi_{d6}^e i_{q6}^e + \frac{M_2}{M_6}(-\psi_{dr2}^e i_{q2}^e)} + L_r - \frac{M_6}{L_{s6}} + M_2 [i_{d6}^e \cos(\theta_{6e} + \theta_{2e}) + i_{d6}^e \sin(\theta_{6e} + \theta_{2e})] i_{q2}^e \\
+ M_2 [i_{q6}^e \cos(\theta_{6e} + \theta_{2e}) + i_{d6}^e \sin(\theta_{6e} + \theta_{2e})] i_{d2}^e \\
- \frac{3M_6}{L_{s6}(L_r - \frac{M_6}{L_{s6}})} \psi_{d6}^e \psi_{dr2}^e \sin(\theta_{6e} + \theta_{2e}) \\
- \frac{3M_6 M_2}{L_{s6}(L_r - \frac{M_6}{L_{s6}})} \psi_{d6}^e [\cos(\theta_{6e} + \theta_{2e}) i_{q2}^e + \sin(\theta_{6e} + \theta_{2e}) i_{d2}^e] \\
\]

(4.37)

where \( \theta_{6e} + \theta_{2e} = \theta_{dr2} = \omega_2 t - 4 \theta - \int_0^t \omega_2 dt \). \( \theta_{dr2} \) is referred to as the BDFM synchronous angle hereafter. \( \psi_{d6}^e, i_{q6}^e \) and \( i_{d6}^e \) can be predicted from the power winding subsystem.

Eqn. (4.36) shows that \( \psi_{dr2}^e \) is only related to \( i_{q2}^e \) and can be kept constant by appropriate control of \( i_{d2}^e \). Using the relationship between \( i_{q2}^e \) and \( \omega_2 \) defined in Eqn. (4.35), the electric torque equation (4.37) can be expressed as a nonlinear function of \( i_{q2}^e \). Thus, \( T_e \) can be controlled by appropriate control of \( i_{q2}^e \). Separate control of \( i_{q2}^e \) and \( i_{d2}^e \) allows for decoupled control of \( T_e \) and \( \psi_{dr2}^e \). In order to find the appropriate value of \( i_{q2}^e \) for a desired torque command value, \( T_e^* \), equation (4.37) needs to be solved. This can be achieved by using a single step Newton-Raphson algorithm.
Defining \( f(i_{q2}^{2e}) \) as:
\[
f(i_{q2}^{2e}) = T_e^* - T_e
\]
(4.38)

the current command \( i_{q2}^{2e} \) can be chosen as
\[
i_{q2}^{2e} = i_{q2}^{2e} - \frac{f(i_{q2}^{2e})}{df(i_{q2}^{2e})}
\]
(4.39)

where
\[
\frac{df(i_{q2}^{2e})}{di_{q2}^{2e}} = -A + \frac{r_i M_2 \delta t}{\psi_{dr2}(L_r - \frac{M_6^2}{L_{66}})} \left[-B \sin(\theta_{6e} + \theta_{2e}) + C \cos(\theta_{6e} + \theta_{2e}) \right]
\]
(4.40)

\[
-(D \sin(\theta_{6e} + \theta_{2e}) + E \cos(\theta_{6e} + \theta_{2e}))i_{q2}^{2e} - [D \cos(\theta_{6e} + \theta_{2e}) - E \sin(\theta_{6e} + \theta_{2e})]
\]

with
\[
A = \frac{M_2}{L_r - \frac{M_6^2}{L_{66}}} (-\psi_{dr2}^{2e})
\]
(4.41)
\[
B = M_2^2 \psi_{dr6}^{2e} \psi_{d2}^{2e}
\]
(4.42)
\[
C = M_2^2 \psi_{dr6}^{2e} \psi_{d2}^{2e} - \frac{3M_6^2}{L_{66}(L_r - \frac{M_6^2}{L_{66}})} \psi_{dr2}^{2e} - \frac{3M_6^2 M_2}{L_{66}(L_r - \frac{M_6^2}{L_{66}})} \psi_{6e}^{2e}
\]
(4.43)
\[
D = M_2 \psi_{dr6}^{6e} - \frac{3M_6^2 M_2}{L_{66}(L_r - \frac{M_6^2}{L_{66}})} \psi_{d6}^{6e}
\]
(4.44)
\[
E = M_2 \psi_{qr6}^{6e}
\]
(4.45)
4.3.4 Simulation Results

Fig. 4.2 shows a speed control system using the rotor flux decoupled control algorithm. A predictor is used to predict non-measurable quantities for the power winding subsystem. The control winding flux current command is determined from Eqn. (4.36). Using a one step Newton-Raphson algorithm, the torque command current $i_q^{2e*}$ is obtained using terminal measurements, power winding subsystem predictor results and the desired torque and control winding flux commands. A position encoder provides accurate rotor position for the reference frame transformations. The laboratory BDFM prototype parameters are listed in Table 6.2.

![BDFM Rotor Flux Decoupled Controller Block Diagram](image)

**Figure 4.2** BDFM Rotor Flux Decoupled Controller Block Diagram

Figs. 4.3 through 4.8 show simulation results for a step speed command change from 600 r/min to 700 r/min, while the drive is supporting a constant load torque (8 Nm). Fig. 4.3 shows that a PI regulator in the speed loop can lead to an acceptable speed response. Fig. 4.4 illustrates the commanded control winding phase current. Both
frequency and phase angle adjustments are needed, which are equivalent to three phase instantaneous current control. The current control algorithm is implemented in simulation using an ideal current source representation. It will be shown in chapter 6 that the actual BDFM machine currents can follow their command values closely with a 10 kHz PWM inverter. Thus, performance degradation and loss of dynamic range expected for current controlled voltage source inverter implementation can be kept to a minimum. Figs. 4.5 and 4.6 illustrate power winding and rotor current responses to a step change in speed command. The power winding flux response shown in Fig. 4.7 illustrates the effect of stator resistance in the machine, which has been neglected during control development. It is evident that this stator resistance contributes to the small value in the q-axis flux in Fig. 4.7, which is assumed to be zero in the controller. The flux oscillations couple into the dynamic torque response and the resulting torque pulsations are visible on the speed response of Fig. 4.3. Overall dynamic power winding flux errors are within ±5%, which seems reasonable considering the assumptions made. As the BDFM machine rating is increased, the power winding resistance decreases compared with the magnetizing impedance, hence the resulting pulsations will be alleviated. Improvements are also possible by appropriate compensation techniques. Control winding fluxes are maintained decoupled as illustrated in Fig. 4.8. Again, a realistic current control algorithm may lead to a slight degradation in this response.

Fig. 4.9 shows a speed response for a step load torque change from 8 Nm to 16 Nm. It is comparable to results obtained with field oriented controllers for conventional induction machine[31]. Fig. 4.10 illustrates the increase in rotor current magnitude to
accommodate the torque change. Fig. 4.11 shows flux response in both power and control windings.

![Speed response graph](image)

**Figure 4.3** Speed Response to a Step Change in Speed Command

![Control winding phase current graph](image)

**Figure 4.4** Commanded Control Winding Phase Current for a Step Change in Speed Command
Figure 4.5  Power Winding Phase Current Response for a Step Change in Speed Command

Figure 4.6  Rotor Current Response for a Step Change in Speed Command
Figure 4.7 Power Winding Flux Response for a Step Change in Speed Command

Figure 4.8 Control Winding Flux Response for a Step Change in Speed Command
Figure 4.9 Speed Response for a Step Load Torque Change

Figure 4.10 Rotor Current Response for a Step Load Torque Change
Figure 4.11 Flux Response for a Step Load Torque Change

Extensive simulations have been carried out in order to evaluate the robustness of the controller with regard to the electrical machine parameters required for control computations. As shown in Table 4.1, the allowable parameter range for which the controller is stable is quite wide, especially for rotor resistance which, as in conventional cage rotor induction machines, is difficult to estimate. It should be noted that in the equivalent circuit model for the BDFM [5], all rotor quantities can be referred to the stator, where the referred rotor resistance adds to the power winding resistance. Hence, neglecting power winding resistance during decoupled control development is equivalent to underestimating rotor resistance in the controller.
Table 4.1 Allowable Parameter Errors in Controller

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Positive Error Range</th>
<th>Negative Error Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>6 pole to rotor mutual</td>
<td>$M_6$</td>
<td>24%</td>
<td>100%</td>
</tr>
<tr>
<td>2 pole to rotor mutual</td>
<td>$M_2$</td>
<td>5%</td>
<td>16%</td>
</tr>
<tr>
<td>rotor resistance</td>
<td>$r_r$</td>
<td>100%</td>
<td>87%</td>
</tr>
<tr>
<td>rotor inductance</td>
<td>$L_r$</td>
<td>23%</td>
<td>52%</td>
</tr>
</tbody>
</table>

Fig. 4.12 illustrates the speed response when rotor resistance in the decoupled control algorithm is overestimated by 50%. Current and flux response are about the same as those obtained with a correct rotor resistance estimation. It is evident that controller performance is acceptable even in the presence of large errors in rotor resistance. In this regard, the decoupled BDFM controller compares very favorably with many conventional field oriented controllers for induction machines.
Figure 4.12  Speed Response for a Step Change in Speed Command With 50% Rotor Resistance Error in the Controller
Chapter 5

Experimental Evaluation of Rotor Flux Oriented Control Algorithms

The synchronous reference frame dq model developed for the BDFM in chapter 4 is universally applicable for other applications such as the BDFM steady state and dynamic analyses, which will not be discussed in this thesis. Here the application focuses on BDFM dynamic control. Chapter 4 has presented a rotor flux oriented control algorithm as a design example. The model significantly reduces the complexity of BDFM controller designs. The control algorithm was successfully tested in simulation.

In this chapter the process of experimentally validating the rotor flux oriented control algorithms is described. First the correct measurement of the synchronous angle $\theta_{s2e}$ is justified. A rotor flux oriented control algorithm utilizing electric torque estimation and $\theta_{s2e}$ is successfully implemented. This verification also validates the example given in chapter 4. Further controller design focuses on its simplification to reduce cost and hence to increase its economic viability and reliability. The importance of $\theta_{s2e}$ is assured by a simplified control method. Further investigation reveals that the measured synchronous angle $\theta_{s2e}$ can be used to represent the BDFM electric torque. Subsequent controller designs can hence be simplified by not estimating the electric torque, which also shortens the control update cycles. All control algorithms developed in this chapter are justified with experimental results.
5.1 A Rotor Flux Oriented Control Algorithm With $T_e$ Estimation

5.1.1 Control Algorithm Development

In order to implement the decoupled control proposed in section 4.3, the electric torque $T_e$ in Eqn. (4.37) has to be estimated. This in turn requires that the BDFM synchronous angle $\theta_{62e}$ to be measured. $\theta_{62e}$ can be expanded into:

$$\theta_{62e} = \theta_{6e} + \theta_{2e} = \theta_{6} - \theta_{2} - 4\theta_r$$  \hspace{1cm} (5.1)

where four variables $\omega_6$, $\omega_2$, $t$ and $\theta_r$ have to be measured in real time.

Measurement of $\omega_6$, $\omega_2$, $t$ and $\theta_r$ depends on the actual hardware setup for the control implementation, details of which will be described in chapter 6. A 14 bit position encoder mounted on the shaft provides adequate resolution for $\theta_r$. $\omega_2$ is commanded from the controller. In chapter 6 it will be shown that the BDFM control winding currents can be forced to closely follow the control current signals by a 10 kHz inverter. This results in adequate resolution for $\omega_2$. Since the BDFM power winding is connected to a rigid grid, it is initially assumed that $\omega_6$ will remain at 60Hz. Thus, $\theta_6$ in Eqn. (5.1) can be calculated as the summation of each $\omega_6 \Delta t$, i.e., $\theta_6 = \omega_6 \Sigma \Delta t$ where $\Delta t$ is the control update period and is preset to be 1ms. However, the steady state synchronous angle $\theta_{62e}$ calculated in this manner is oscillating, instead of constant. The electric torque $T_e$ estimated by using this oscillating $\theta_{62e}$ in Eqn. (4.37) is not a valid representation for the BDFM. Hence the derivation of $i_{q2}^{2e}$ in Eqn. (4.39) can not be used for the controller output signal generation.
Since $\omega_2$ is the command signal and $\theta_e$ comes from the shaft-mounted position encoder, the respective resolutions are adequate. The oscillatory behavior in $\theta_{6e}$ may be caused by the assumption of a constant 60Hz for $\omega_6$ and possible timer error in the microprocessor. The summation process of $\omega_6 \Sigma \Delta t$ further accumulates errors. Consequently, both the assumption of a constant 60Hz for $\omega_6$ and the summation process have to be discarded in order to improve the $\theta_6$ measurement.

An alternative method of obtaining $\theta_6$ is based on Eqn. (4.19). After a power winding phase voltage is sampled by an A/D converter into the microprocessor, $\theta_6$ can be obtained as

$$
\theta_6 = \cos^{-1}\left(\frac{V_{6a}}{\sqrt{2} V_6}\right)
$$

(5.2)

Here $\theta_6$ ranges from 0 to $2\pi$. Since there are two $\theta_6$ values corresponding to one $V_{6a}$ value for a cycle of sinewave, the rate of change in $V_{6a}$ is needed to help determine a unique $\theta_6$ for each $V_{6a}$. Experimental evaluation of this technique yields a stable and constant $\theta_{6e}$ during steady state BDFM synchronous operation, as expected. This enables further controller designs.

Rearranging the BDFM electric torque equation (4.37) yields

$$
T_e = H + I \cos \theta_{6e} + J \sin \theta_{6e}
$$

(5.3)

where
As illustrated in 4.3, quantities on the power winding side can be predicted and those on the control winding side can be measured. Hence it is easy to conclude from Eqn. (5.3) that if the desired BDFM electric torque $T_e^*$ is given, the corresponding desired synchronous angle $\theta_{62e}^*$ will be

$$\theta_{62e}^* = \sin^{-1} \left( \frac{T_e^*}{L_6} - \tan^{-1} \left( \frac{I}{J} \right) \right)$$  \hspace{1cm} (5.7)$$

The objective of the controller design is to generate the control current signal inputs to the power inverter. These signals can be specified by $i_{q2}^{6e}$, $i_{d2}^{6e}$ on the synchronous frame and $\theta_2$. $i_{d2}^{6e}$ can be calculated by Eqn. (4.36) with specification of the desired rotor flux level on the control winding side, $\psi_{d2}^{6e}$. Usually $\psi_{d2}^{6e}$ is used to regulate the BDFM steady state performance and kept constant during speed and load transients. For this case $i_{d2}^{6e}$ relates to $\psi_{d2}^{6e}$ simply as
The \( \theta_2 \) control signal can be derived from Eqn. (5.1) by replacing the real \( \theta_{62e} \) with its desired value,

\[
\theta_2 = \theta_6 - \theta_{62e} - 4\theta_r
\]  

(5.9)

where \( \theta_6 \) and \( \theta_r \) come from measurements. Finally, Eqn. (4.35) can be used to obtain \( i_{q_2}^{2e} \) from known \( \theta_2 \). Integration of Eqn. (4.35) over time \( t \) gives the change of \( \theta_2 \) over one control cycle as

\[
\Delta \theta_2 = \frac{r_r M_2}{\psi_{ad2}(L_r - \frac{M_6}{L_{s6}})} i_{q_2}^{2e} \Delta t - \Delta \theta_r
\]  

(5.10)

\( \Delta t \) is the time for a control cycle. Rearrangement gives \( i_{q_2}^{2e} \) with \( \Delta \theta_2 \) calculated using Eqn. (5.9).

\[
i_{q_2}^{2e} = \frac{\psi_{ad2}(L_r - \frac{M_6}{L_{s6}})(\Delta \theta_2 + \Delta \theta_r)}{r_r M_2 \Delta t}
\]  

(5.11)

5.1.2 Experimental Results

Fig. 5.1 gives the block diagram of a speed control system based on the rotor flux oriented algorithm with \( T_e \) estimation. A predictor is implemented based on Eqn. (4.23) to predict power winding quantities needed for electric torque estimation. The \( T_e \)
estimation also needs control winding quantities which are already available inside the microprocessor; as they are control signals. The desired rotor flux level on the control winding gives its flux current component \( i_{d2e} \). The desired torque command signal \( T_e^* \) is generated by the summation of the estimated torque and the speed error signals. Eqns. (5.7), (5.9) and (5.11) then give the desired synchronous angle \( \theta_{d2e}^*, \theta_2 \) and \( i_{q2e}^2 \) in sequence. Then transformation between the synchronous reference frame and the abc domain is performed to give the abc domain control currents \( i_{d2}^*, i_{b2}^* \) and \( i_{c2}^* \).

The above algorithm is implemented using the setup described in chapter 6. Two tests are performed. One is illustrated in Figs. 5.2 to 5.5, showing BDFM behavior in
response to a 100 1/min step speed command. The other is given in Figs. 5.6 to 5.9 to show the BDFM response for a load torque change.

The BDFM speed response is taken under no load condition, since no dynamic load controller for the DC machine is available in the laboratory to provide fast load torque regulation. Fig. 5.2 shows the BDFM speed response for the 100 1/min step change command. BDFM speed settles on the new desired speed in about 300ms. Fig. 5.3 shows how currents in the control and power windings behave during the speed response. Fig. 5.4 illustrates the desired torque calculated inside the controller according to the step speed command and the BDFM speed feedback. Although the desired torque signal in Fig. 5.4 displays a small oscillation after returning to zero, its effect on the shaft speed is certainly negligible as shown in Fig. 5.2. The desired BDFM synchronous angle $\theta_{q2e}^*$ and $i_{q2e}$ are plotted in Fig. 5.5. Values for Figs. 5.4 and 5.5 are taken from the 80960KB microprocessor and exhibit some noise. As compared to the results obtained from the closed loop scalar control method described in chapter 2, the speed response in Fig. 5.2 is much faster.
Figure 5.2 Speed Response to a Step Change in Speed Command

Figure 5.3 The BDFM Currents for a Step Change in Speed Command
Figure 5.4 The BDFM Desired Torque for a Step Change in Speed Command

Figure 5.5 The BDFM Desired Synchronous Angle $\theta_{62e}^*$ and $i_{q2}^{2e}$ for a Step Change in Speed Command
Fig. 5.6 illustrates the speed response for a load torque change from no load to 8Nm. This test uses using a DC load machine, which is mechanically connected to the BDFM shaft. The DC machine has a separate field winding which is excited by a constant voltage. The load torque change is implemented by switching a load resistor into the DC machine armature circuit. Armature current is used in Fig. 5.6 to indicate when the load torque change takes place. The small oscillation on the armature current shows the effect of the commutator. As shown, the speed returns to its desired value within 400ms. Currents in both the control and power windings are illustrated in Fig. 5.7 where the slight cycle time reduction in the control winding current can be observed during the transient. The reduction compensates for the drop in speed. Fig. 5.8 shows that the desired torque value moves to a new level corresponding to the new load torque value. Fig. 5.9 depicts the desired synchronous angle $\theta_{62e}^*$ and $i_{q2e}^2$. The values plotted in Figs. 5.8 and 5.9 are obtained from the control 80960KB microprocessor.

Careful observation reveals a slight difference between control algorithms implemented in Figs. 4.2 and 5.1. Although both algorithms are rotor flux oriented, Fig. 4.2 emphasizes utilization of $i_{q2e}^2$ in the control, while Fig. 5.1 focuses on $\theta_{62e}$. However, since $i_{q2e}^2$ and $\theta_{62e}$ are related to each other by Eqns. (5.1) and (5.11), both algorithms are basically the same. Consequently, the experimental results also justify the control algorithm presented in Fig. 4.2.

Comparison of the desired electric torque $T_e^*$ and the desired synchronous angle $\theta_{62e}^*$ in both step speed response and load torque change tests exposes their similarities. This fact will be further exploited in section 5.3.
Figure 5.6  Speed Response to a Change in Load torque

Figure 5.7  The BDFM Currents for a Change in Load Torque
Figure 5.8 The BDFM Desired Torque $T_e^*$ for a Change in Load Torque

Figure 5.9 The BDFM Synchronous Angle $\theta_{62e}^*$ and $i_{q2}^{2e}$ for a Change in Load Torque
5.2 A Simplified Control Algorithm Without $\theta_{62e}$ Measurement

5.2.1 Control Algorithm Derivation

Section 5.1 verifies the rotor flux oriented control algorithm based on electric torque estimation and $\theta_{62e}$ information. It is desirable to reduce the complexity of the algorithm for actual applications. Since section 5.1 uses $\theta_{62e}$ extensively, an investigation into the importance of the BDFM synchronous angle $\theta_{62e}$ is desirable. A simplified control algorithm is proposed here, assuming a constant angle during transients to eliminate the complexity of using $\theta_{62e}$. Thus, the transient behavior of $\theta_{62e}$ is ignored and no measurement of $\theta_{62e}$ is required.

As evident from Eqns. (4.23) and (4.30) to (4.36), the above assumption does not affect the power winding prediction and the control winding flux alignment described in section 4.3.1 and 4.3.2. However, the derivation of $i_{q2}^{2e}$ needs to be adjusted accordingly. Eqn. (4.37) can now be arranged as:

\[
T_e = P - M_2 \dot{\theta}_{6e}^{2e} - \frac{M_2}{L_r} \dot{\psi}_{d2}^{2e} \\
= - \frac{M_2}{L_r} \dot{\psi}_{d2}^{2e} - \frac{M_2}{L_r} \dot{\psi}_{q2}^{2e}
\]

(5.12)

where
\[
F = 3 \frac{M_6}{L_{s6}} (-\psi_{d6}) i_{q6}^{6e} - \frac{3M_6}{L_{s6}(L_r - \frac{M_6}{L_{s6}})} \psi_{d6}^2 \psi_{dr2}^{2e} + M_2 i_{d6}^{6e} i_{d2}^{2e} \\
+ \frac{3M_6 M_2}{L_{s6}(L_r - \frac{M_6}{L_{s6}})} \psi_{d6}^{6e} i_{d2}^{2e}
\]  

(5.13)

\(\psi_{dr2}^{2e}\) will be a constant if the flux component \(i_{d2}^{2e}\) of the control current remains the same during the transients. The other quantities can be obtained from the BDFM power winding predictor. Eqn. (5.12) shows that the torque command \(i_{q2}^{2e}\) of the control current can be obtained as

\[
i_{q2}^{2e} = \frac{F - T_e^*}{M_2 (I_{q6}^{6e} + \frac{\psi_{dr2}^{2e}}{L_r - \frac{M_6}{L_{s6}}})}
\]

(5.14)

**Figure 5.10** Simplified Control Without \(\theta_{62e}\) Measurement for BDFM
A block diagram of the simplified control algorithm is shown in Fig. 5.10. The power winding quantities are predicted by Eqn. (4.23). Eqns. (4.35) and (4.36) generate the $\omega_2$ and $i_{q2}^2$ commands separately. Only Eqn. (4.39) needs to be replaced by (5.14). As shown, the simplified control does not require rotor position information, and the one step Newton-Raphson iteration method used in the decoupled control algorithm is also eliminated. Therefore, the hardware resources required for the simplified controller are reduced significantly.

5.1.2 Simulation and Experimental Results

A simulation is carried out to examine the controller robustness with respect to controller parameter variations. The criterion for testing the robustness of the controller in simulation is whether the controller can maintain the BDFM in synchronism and bring the speed to its desired value if a new speed is commanded. Table 5.1 shows the allowable range for the BDFM parameters based on the simulation results.

Figs. 5.11 and 5.12 show simulation and experimental results of the BDFM speed response with a 100 r/min step command. The laboratory setup for the simplified control algorithm can be found in Chapter 6.
Table 5.1 Simplified Controller Robustness Against Parameter Variations

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Positive Error Range</th>
<th>Negative Error Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>$r_r$</td>
<td>3800%</td>
<td>90%</td>
</tr>
<tr>
<td>$M_6$</td>
<td>30%</td>
<td>9%</td>
</tr>
<tr>
<td>$M_2$</td>
<td>40%</td>
<td>25%</td>
</tr>
<tr>
<td>$L_r$</td>
<td>15%</td>
<td>75%</td>
</tr>
<tr>
<td>$L_{s6}$</td>
<td>20%</td>
<td>45%</td>
</tr>
</tbody>
</table>

Figure 5.11 Simulation Result for a 100 1/min Speed Command Change
Figure 5.12 Experimental Result of the Speed Response for a 100 l/min Speed Command Change

From Figs. 5.11 and 5.12, it is evident that the simulation result confirms the experimental result. Although both figures show that the simplified control algorithm can achieve relatively fast dynamic response, an undesirable oscillatory response occurs during transients. This indicates that because the BDFM electric torque expression is simplified to avoid calculation of the synchronous angle $\theta_{62e}$, decoupled control is not achievable.

It can be concluded from the above investigation that $\theta_{62e}$ plays an important role in establishing BDFM dynamic response. As illustrated in Fig. 5.11 and 5.12, the BDFM dynamic response is negatively impacted when $\theta_{62e}$ is not used. Therefore, further efforts need to be directed toward using $\theta_{62e}$. Since $\theta_{62e}$ is critical for controlling the BDFM transient behavior, it has to be closely involved in any controller design, even more than
ig22e, which is the primary variable in the controller design in section 4.3. This philosophy is embodied in the rest of this chapter.

5.3 A Rotor Flux Oriented Control Algorithm Without Tₐ Estimation

5.3.1 Control Algorithm Development

Section 5.1 describes a typical process for BDFM dynamic controller design. Before any dynamic controller can be designed for the BDFM, its electric torque behavior during transients has to be studied and understood well. Then a torque estimator is built into the controller to provide adequate electric torque information for derivation of the control signals. However, since the torque estimation is derived from the full BDFM dq model, a large amount of data processing time is needed. Thanks to the fast floating point unit in the Intel 80960KB microprocessor used in the laboratory setup, the controller can complete one cycle of data processing within 1ms. Nevertheless, since the BDFM is constructed to be as simply as a conventional induction machine, there is always a desire to formulate the BDFM control algorithms as simple as possible such that the whole BDFM control system can compete with those for conventional induction machines.

The electric torque estimation process can be avoided if an easily controlled variable can be found to have a sole relationship with the electric torque. The synchronous angle θₑₑₑ is the natural choice since section 5.1 reveals the similarity between Tₑ* and θₑₑₑ* responses during transients. Moreover, θₑₑₑ can be easily controlled.
The relationship between $\theta_{62e}$ and $T_e$ is illustrated in Figs. 5.13 and 5.14 at 700 and 825 1/min, respectively. Both simulation and experiment are conducted with the control winding under vector control and at constant flux. The simulation results are given by the BDFM electric torque Eqn. (4.37) or (5.3). Obviously the experimental results validate the simulations. As depicted in the figures, the BDFM torque angle characteristic is similar to that of conventional synchronous machines, except that the offsets are different. Similarly, the BDFM can only operate on the rising slope, as indicated in the figures, while operation on the falling slope is unstable. It has to be mentioned that the results in Figs. 5.13 and 5.14 are based on lower control current magnitudes and hence give lower maximum torques than shown in the maximum torque simulation in Fig. 2.3.

Both Figs. 5.13 and 5.14 manifest that the synchronous angle $\theta_{62e}$ is solely related to the electric torque $T_e$ over the stable rising slope. Hence, the BDFM electric torque

![Figure 5.13 The BDFM Electric Torque Versus Synchronous Angle](image-url)
Figure 5.14 The BDFM Electric Torque Versus Synchronous Angle

can be represented by $\theta_{62e}$ over the same slope. Operation in the unstable region (falling slope) can be prohibited by a limiter designed for $\theta_{62e}$. Because $\theta_{62e}$ can be measured and controlled quite easily, the controller design will be simple.

5.3.2 Experimental Results

Fig. 5.15 shows the block diagram of a controller design using $\theta_{62e}$ instead of $T_e$ estimation. Because the power system predictor is only needed for the $T_e$ estimation process, it can be eliminated to further simplify the controller structure and speed up the data processing time inside the microprocessor. As a result, the control update cycle can be shortened to 0.5ms. The $\theta_{62e}^*$ signal is generated by a simple proportional and integral regulator using the speed error signal. The $\theta_2$ and $i_{q2}^{2e}$ signals are then calculated from the $\theta_{62e}^*$ command. The desired control winding flux level gives $i_{d2}^{2e}$. 
Figs. 5.16 to 5.19 demonstrate the BDFM response for a 100 1/min step change in speed command under no load operation. The control update cycle is preset at 1ms. Fig. 5.16 is the shaft speed response, which has an overshoot but returns to its desired value within 700ms. Both currents in the BDFM control and power winding are shown in Fig. 5.17. The desired synchronous angle $\theta_{62e}^*$ is depicted in Fig. 5.18. Fig. 5.19 displays the $i_{q2}^{2e}$ behavior. Based on Eqns. (5.9 and (5.11), it can be shown that a small error between $\theta_{62e}^*$ and $\theta_{62e}$ can cause significant regulation on $i_{q2}^{2e}$.

Figs. 5.20 to 5.23 illustrate the BDFM response for a load torque change from no load to 8Nm. The DC machine armature current is used to indicate when the load torque is changed in Fig. 5.20. The speed returns to its desired value within 600ms. Fig. 5.21 shows both control and power winding currents. Fig. 5.22 is the desired synchronous angle $\theta_{62e}^*$. $\theta_{62e}^*$ moves to a new value corresponding to the new load torque. Fig. 5.23
displays $i_{q2}^{2e}$. Quantities plotted in Figs. 5.18, 5.19, 5.22 and 5.23 are again obtained from the outer loop 80960KB microprocessor.

**Figure 5.16** Speed Response to a Step Change in Speed Command

**Figure 5.17** The BDFM Currents for a Step Change in Speed Command
Figure 5.18 The BDFM Desired Synchronous Angle $\theta_{o2e}^*$ for a Step Change in Speed Command

Figure 5.19 $i_{q2e}^{2e}$ for a Step Change in Speed Command
Figure 5.20  Speed Response to a Change in Load Torque

Figure 5.21  The BDFM Currents for a Change in Load Torque
Figure 5.22 The BDFM Desired Synchronous Angle $\theta_{a2e}^*$ for a Change in Load Torque

Figure 5.23 $i_{q2}^{2e}$ for a Change in Load Torque
5.4 Improvement Over The Rotor Flux Oriented Control Algorithm Without $T_e$ Estimation

The elimination of the BDFM electric torque estimation and the power system predictor in Fig. 5.15 simplifies the controller designs and can certainly speed up the dynamic control implementation. This implies that cheap and moderate microprocessors can be used to replace the 80960KB. However, the use of a simple PI regulator to generate $\theta_{62e}^*$ from the speed error signal does not necessarily guarantee better results than those obtained from the control algorithm with $T_e$ estimation, as evident by comparing Figs. 5.16 to 5.23 and Figs. 5.2 to 5.9.

Because the control algorithm is implemented in the 80960KB, an improvement can be easily achieved by replacing the PI regulator with an advanced controller as illustrated in Fig. 5.24. The advanced controller can range from simple PI plus limiter control to adaptive algorithms such as model reference or self-tuning control. Here an example of such control is formulated as follows:

(1) Whenever a speed change command is received, the controller commands the desired synchronous angle $\theta_{62e}^*$ to its preset limit, positive for increasing speed and negative for decreasing speed. The $\theta_{62e}^*$ limits can be obtained from Eqn. (4.37) or (5.3) with knowledge of the BDFM parameters, as illustrated in Figs. 5.13 and 5.14. Forcing $\theta_{62e}^*$ to its limit ensures maximum speed response. This is called the desired synchronous angle $\theta_{62e}^*$ limit controller.

(2) Once the BDFM shaft speed reaches the new desired speed value, a normal PI regulator is used to settle the shaft speed to its new value.

The other control blocks in Fig. 5.24 are the same as those in Fig. 5.15.
Figure 5.24 Improvement Over Rotor Flux Oriented Control Without $T_e$ Estimation

Figs. 5.25 to 5.28 present the BDFM response to a 100 l/min step change in speed command under no load condition. The control update cycle is preset at 1ms. Fig. 5.25 shows that the dynamic response can be completed within 250ms. Fig. 5.26 illustrates the currents on both windings. The advanced control is clearly depicted by the desired synchronous angle $\theta^{\alpha2e}_d$ in Fig. 5.27. Fig. 5.28 displays $i_{q2}^{2e}$. Values plotted in Figs. 5.27 and 5.28 are obtained from the outer loop microprocessor.

The advanced control algorithm in Fig. 5.24 can only improve the speed response corresponding to the speed change command. Once the shaft speed reaches the new value, it is regulated by a PI controller, equivalent to Fig. 5.15. This means that the control algorithm in Fig. 5.24 will respond the same as that in Fig. 5.15 for any load torque change.
Figure 5.25 Speed Response to a Step Change in Speed Command

Figure 5.26 The BDFM Currents for a Step Change in Speed Command
Figure 5.27 The BDFM Desired Synchronous Angle $\theta_{62e}^*$ for a Step Change in Speed Command

Figure 5.28 $i_{q2}^{2e}$ for a Step Change in Speed Command
Chapter 6

Experimental System for Control Implementation

The experimental system for the BDFM control implementation is shown in Fig. 6.1. An Intel 80960KB floating point computation intensive microcontroller is used to replace the initial DSP56001 fixed point microprocessor to accelerate the data processing. This includes the control algorithm as well as the required transformation and predictor calculations, based on command signals issued via a personal computer. To allow for modular design, another microprocessor Intel 80196Kr is used to control the power inverter. Three high speed A/D converters (6μs, 12 bit) provide for current feedback information. Another high speed A/D converter is utilized to obtain the phase voltage signal from the power winding used to determine $\theta_6$.

Before applying the control algorithm with $T_e$ estimation, a predictor algorithm is used to solve for power winding quantities. The Intel 80960KB microcontroller is equipped with a floating point unit, resulting in sufficient speed to support a control update cycle of 1ms. Position feedback information is provided by a 14 bit absolute encoder.

Although control winding quantities can be transformed into different rotating reference frames for ease of controller design, the winding is physically fixed on the stationary frame. Therefore, all control quantities have to be transformed back to the stationary frame before they can physically be realized. After this transformation to the stationary frame, the desired control winding currents are controlled by the fixed-update-
time PWM current control algorithm [27] implemented in a conventional hard switched IGBT inverter. The inverter switching frequency is preset at 10 kHz.

In this chapter, detailed descriptions of experimental system hardware and implementation are presented. Circuit schematics are given in the Appendices. Fig. 6.2 shows the experimental setup for the data processing and control output section. This section mainly consists of a power converter, two Intel evaluation boards for the 80960KB and 80196Kr microprocessors, which are housed on a rack. Two personal computers (PCs) are used to communicate with these evaluation boards for development and operator control purposes.

![Data Processing and Control Section for BDFM Dynamic Control](image)

**Figure 6.2** Data Processing and Control Section for BDFM Dynamic Control
6.1 The Prototype Brushless Doubly-Fed Machine

The laboratory system connects the BDFM with a DC machine on the same shaft for various test purposes [43]. Fig. 6.3 shows a picture of the laboratory BDFM prototype at the front. The DC machine is at the far upper end. Since it is desirable to compare the BDFM with an induction motor in terms of operating performance for a given frame size, the laboratory prototype was constructed in an existing induction

![Figure 6.3 The Laboratory BDFM Prototype](image-url)
machine frame, using stator laminations of a four pole induction motor with a custom designed rotor. The stator was rewound to give two sets of three phase windings for power and control purposes. The physical dimensions of the BDFM are given in Table 6.1.

### Table 6.1 Physical Dimensions of the Laboratory BDFM Prototype

<p>| | | | | |</p>
<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Stator Frame Size</strong></td>
<td>445</td>
<td><strong>Stator Slots</strong></td>
<td>72</td>
<td><strong>Airgap</strong></td>
</tr>
<tr>
<td><strong>Power Winding:</strong></td>
<td></td>
<td><strong>Control Winding:</strong></td>
<td></td>
<td><strong>Rotor:</strong></td>
</tr>
<tr>
<td><strong>Voltage</strong></td>
<td>230V</td>
<td><strong>Voltage</strong></td>
<td>230V</td>
<td><strong>Diameter</strong></td>
</tr>
<tr>
<td><strong>Frequency</strong></td>
<td>60Hz</td>
<td><strong>Frequency</strong></td>
<td>0-60Hz</td>
<td><strong>Slots</strong></td>
</tr>
<tr>
<td><strong>Coil Pitch</strong></td>
<td>15/18</td>
<td><strong>Coil Pitch</strong></td>
<td>5/6</td>
<td><strong>Nests</strong></td>
</tr>
<tr>
<td><strong>Winding Layer</strong></td>
<td>2</td>
<td><strong>Winding Layer</strong></td>
<td>2</td>
<td><strong>Loops/Nest</strong></td>
</tr>
<tr>
<td><strong>Turns/Coil</strong></td>
<td>10</td>
<td><strong>Turns/Coil</strong></td>
<td>12</td>
<td><strong>Slot Pitch</strong></td>
</tr>
</tbody>
</table>

In order to proceed with any control development, the BDFM parameters have to be measured or estimated. An off-line estimation scheme was developed in the frequency domain to suit the specific structure of BDFM [15]. The electrical parameters for the prototype are listed in Table 6.2.
Table 6.2 Laboratory BDFM Prototype Parameters [15]

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_p$ ($\Omega$)</td>
<td>0.7</td>
</tr>
<tr>
<td>$R_c$ ($\Omega$)</td>
<td>1.8</td>
</tr>
<tr>
<td>$L_p$ (H)</td>
<td>0.049</td>
</tr>
<tr>
<td>$L_c$ (H)</td>
<td>0.247</td>
</tr>
<tr>
<td>$M_p^2/R_r$</td>
<td>0.0027</td>
</tr>
<tr>
<td>$M_c^2/R_r$</td>
<td>0.0338</td>
</tr>
<tr>
<td>$L_r/R_r$</td>
<td>0.194</td>
</tr>
</tbody>
</table>

6.2 Power Converter Circuit

The power circuit built for control winding excitation consists of an input diode bridge, a DC link and an output IGBT inverter. A simple diode bridge is built to expedite the BDFM control implementation. Six IR16F60 diodes are used, enabling the bridge to process up to 16A input current. The DC link uses a 2400µF electrolytic capacitor as the energy storage component. Because an electrolytic capacitor presents a high impedance path for high frequency noise associated with switch-mode power electronic circuits, two 0.47µF ceramic high frequency capacitors are used in parallel with the electrolytic capacitor for noise filtering purpose.

Because the diode bridge can only process power in one direction from grid to DC link, a resistor dumping circuit was built to dissipate the excess energy in the link once the BDFM control winding starts regenerating. The dumping circuit block diagram is shown in Fig. 6.4; a detailed schematic is included in Appendix A. A resistor network scales down the DC link voltage to an appropriate low voltage and feeds it to a
Figure 6.4 Dumping Circuit Block Diagram

comparator. Once the voltage is higher than a preset threshold value, the comparator issues a gate signal to turn on the dumping IGBT switch. The IGBT then connects four 50Ω resistors in series to DC bus; the system is capable of dissipating up to 900W continuously. Since the dumping circuit directly uses the DC link negative bus as its common, which is different from the other control circuits, no isolation circuitry is needed. To prevent the dumping IGBT switch from tripping on and off due to noise on the bus voltage signal, a hysteresis band is also provided. Although the combination of a diode bridge and a dumping circuit provides for fast and simple implementation, undoubtedly it restricts system efficiency whenever the control winding enters the regeneration mode. A regenerative rectifier stage can be used to overcome this restriction [44].

The inverter stage consists of six IXYS IXSN35N100U1 IGBT packaged with integral diodes. Each package has a continuous current rating of 25A [45]. The stage regulates the three phase BDFM control winding instantaneous currents. Fig. 6.5 shows a picture of the power inverter.
6.3 Intel 80196Kr Control Sub-System

6.3.1 EV80C196KR Core System [46]

As noted in Chapter 4, the current forced model of the BDFM is used to derive the control algorithm. In order for the inverter to generate control winding currents, appropriate switching functions have to be generated for the IGBTs. An EV80C196KR evaluation board is selected to perform this function and give enough flexibility for later modifications.
Fig. 6.8 shows the block diagram of the EV80C196KR board. The board incorporates several peripherals around the 16 bit 80196Kr microprocessor and is ready for application specific implementation. An 8K 16 bit word erasable programmable read only memory (EPROM) monitor program residing on board uses a dedicated external universal asynchronous serial controller (UART) to communicate with a host personal computer for software development. RAMs are formed into an 8K word and an 8K byte segments to be used for application program storage and execution. Expansion boards are connected to the memory I/O expansion bus on board at specific memory addresses to provide command and feedback signals. After processing the signals, the 80196Kr uses six channels from its Event Processor Array (EPA) unit to output gate signals to the inverter IGBTs.
6.3.2 Current Feedback Path

In order for the inverter to be capable of regulating the output current waveforms, the 80196Kr microprocessor needs current feedback information. Fig. 6.7 is the block diagram of the current feedback path, which performs three phase current sensing, analog signal conditioning, fast analog to digital conversion and interfacing with the microprocessor. It is divided into three functional units: current sensing, A/D conversion and interfacing boards. Detailed schematics are included in Appendices B through D.

![Figure 6.7 Block Diagram of the Current Feedback Path](image)

Fig. 6.8 shows the block diagram of the current sensing and conditioning board. Three Micro Switch CSLA1CD hall effect current sensors are used to measure the three phase currents of the BDFM control winding. The purpose of providing three current sensors for the inverter current regulation is to assure its flexibility for the other applications, such as 4 wire system. Each sensor is capable of measuring up to 57A peak current with a delay of 3 μs. The sensors are supplied with a single 12V DC source and give 6V output at zero current. Because the sensor output voltage ranges from 0 to 12V and has limited driving capability, operational amplifiers (LM324) are used to condition the voltage to a 0 to 10V range, which is appropriate as input to the A/D board. The LM324 also assures enough capability for driving the next stage.
Although the 80196Kr has a built-in 10 channel A/D unit, each channel is limited to a 10 bit bandwidth with a conversion time of up to 40 \( \mu s \). A multiplexing technique is used within the microprocessor and a single current measurement comprising three A/D conversions may take up to 120 \( \mu s \). Simultaneous sampling of the three phase currents is not possible using the on-chip A/D converters. In order to improve the A/D performance, three external 12 bit fast A/D converter boards are provided. Fig. 6.9 illustrates the block diagram of one A/D board. A sample and hold chip (HA5320) is used to hold the analog signal during the conversion process. The A/D converter (AD578KN) is capable of completing a 12 bit A/D conversion in 4.5 \( \mu s \) by using a successive approximation technique. Combined with the 1 \( \mu s \) sampling time of the HA5320, the entire A/D process takes 5.5\( \mu s \). The logic in the external A/D boards is configured such that sampling and conversion commence at the same time for all boards. The converted digital values are then held in output data buffers for sequential output to the 80196Kr microprocessor.
An interface board enables selection of different inputs based on commands from the 80196Kr, as illustrated in Fig. 6.10. The microprocessor data bus is interfaced with the outputs of two data buffers. One buffer has a data width of 12 bits and is interfaced with three current A/D boards. The other buffer is a full 16 bits wide and is used to interface with the hardware buffer, which in turn interfaces with the outer loop microcontroller (80960KB). According to the given address, the data bus of the 80196Kr microprocessor can interface to different A/D boards or hardware buffer ports selected by the address decoder. The hardware buffer ports are built between the 80196KR and the main 80960KB microprocessors for data transfers. Each port is 16 bit wide and can transfer one 16 bit variable between the two microprocessors. Two ports can be combined into a 32 bit port and transfer a 32 bit variable, if needed. The 80196KR system only does current regulation such that θ₁ information is not needed. Measurement of θ₁ will be discussed in the 80960KB system.
6.3.3 Pulse-Width-Modulation Implementation

As stated earlier in section 6.3.2, the configuration for the inverter control is for general purpose applications, which may include 4 wire unbalance load application. Three current sensors are built to provide such flexibility. However, when the inverter is used for a motor load, such as the BDFM control winding, two sensors may only be needed. The control algorithm developed here are based on three sensor circuitry.

The BDFM control algorithms as discussed in chapters 4 and 5 are developed based on a current forced model. Thus, the inverter IGBTs have to be switched such that output currents follow the command requirements. Three main methods for current regulation are available, as discussed in the following.
Hysteresis current control, also referred to as tolerance band control, is illustrated in Fig. 6.11. For each phase arm, the error between reference and the measured current values is compared with a preset hysteresis band around the reference value. This comparison leads to three conditions where appropriate actions are taken to switch that particular phase arm. If the measured current exceeds the positive hysteresis band, the lower switch in the phase arm will be turned on, implying a complementary turn-off of the upper switch. Control action is opposite if the measured current drops below the negative band. No actions are taken as long as the current remains inside the hysteresis band.

Fig. 6.11 does not show any interconnection between the switching of different arms. Phase arm switching may not be synchronized. This in turn may require more
logic devices and may lead to unbalanced output currents. It is desirable to synchronize the switching of the three arms. This leads to a maximum error PWM hysteresis control, where the maximum error of all three phase currents, as compared with their respective reference values, is monitored. If the maximum error is detected to exceed the tolerance band, appropriate switching for the inverter is instigated.

The switching frequency of such control depends on how fast the current changes from the lower to the upper limits and vice versa. Such changes are dependent on load and inverter DC link voltage. Moreover, the switching frequency does not remain constant but varies with the current waveform. When the inverter load is not inductance dominated, the output filter design will be difficult due to this variation in switching frequency. However, the control algorithm reduces the number of switching transitions.

6.3.3.2 Fixed Frequency Control [47]

A fixed frequency switching algorithm is illustrated in Fig. 6.12. The error between reference and measured currents is fed through a PI regulator. The output of the regulator $V_{\text{cont}}$ is compared with a fixed frequency triangular waveform, $V_{\text{tri}}$, which determines the inverter switching frequency. Positive error will control the inverter to increase the phase current in order to match the reference value. The opposite action will take place for negative error values.

Because the inverter is switching at a fixed frequency, the harmonics at the output are easily defined, which simplifies output filter design. However, the inverter needs to be switched every cycle, resulting in higher switching losses.
6.3.3.3 Fixed-Update-Time Control [27]

The block diagram for a fixed-update-time control is shown in Fig. 6.13. As illustrated, a timer is implemented and at the end of each time interval all three phase currents are measured and compared with their respective reference values. A decision making block, which can comprise a comparison logic or a microprocessor, then decides on an appropriate switching function for the inverter to bring the currents close to their reference values. Each inverter phase arm does not need to switch for every cycle. Indeed it only needs to switch when the current error of the particular phase changes sign, i.e., the error changes from positive to negative or vice versa. This makes the fixed-update-time control comparable to hysteresis control in terms of reducing switching losses in the inverter.
As stated before, the inverter switching functions are implemented in a 80196Kr microprocessor which allows flexibility for later modifications. The 80196Kr is equipped with two embedded timers and can easily satisfy the requirements of the fixed-update-time control method. Therefore, compared with hysteresis and fixed frequency methods, the fixed-update-time control is the easiest to be implemented in terms of available resources and inverter switching loss minimization.

6.3.3.4 Inverter Switching Function Generation

In the laboratory implementation, the fixed-update-time method is used to generate the inverter switching function, as illustrated in the flowchart of Fig. 6.14.

At power up, the microprocessor starts to execute the inverter code from a specific address location. An initialization process is carried out to configure various on chip resources such as timers and Event Processor Array (EPA) unit. After setting up
Figure 6.14 Inverter Switching Function Generation Flowchart

interrupts and operating mode, the microprocessor starts the timer for a fixed-update-time interrupt. The hardware buffer is examined for updated command values at a fixed time cycle which is preset at 100μs. The command values are the instantaneous current signals output from the outer loop microprocessor 80960KB.

After inputing phase current references, the actual currents are sampled, converted from analog to digital values and read into the 80196Kr microprocessor. Comparison between the respective phase reference and measured current values yields the switching functions, i.e., if the reference value is higher than the measured current, the upper switch is turned on and the lower switch is turned off; if the reference value is lower than the measured current, the opposite switching sequence takes place. Once the 80196Kr
microprocessor finishes one control cycle, it waits until the next timer overflow to start with another control cycle.

### 6.3.4 Isolation and IGBT Drivers

There should be at least one isolation layer between microprocessor and main power inverter circuitry to prevent any controller damage from faults in the inverter power stage. Such isolation can be provided either by optocouplers, fiber optic cables or high frequency isolation transformers. For a 230V application, optocouplers are adequate. As shown in Fig. 6.15, one optocoupler HPCL2232 has two channels, which are connected in such way that at most one output can be active at any time. This prevents both upper and lower IGBTs in a phase arm from being turned on at the same time and protects the DC bus against shoot-through faults. The switching function generator provides for a 5 ms blanking time between complementary IGBTs turning off and on. In addition, the HPCL2232 configuration provides for a hardware blanking mechanism to further ensure safe inverter operation.

![Figure 6.15 Inverter IGBT Driver for a Phase Arm](image-url)
A combination of high side (IXYS IXBD4411) and low side (IXBD4410) drivers are used per phase arm [45]. The power supply ground of the lower side driver is tied to the DC link negative bus. Because of the high voltage between high and low side circuits, a separate power supply is provided for the upper switches, using an oscillator, a pulse transformer and a secondary rectifier circuit. Since the output stage of the optocoupler HCPL2232 is also relative to the low side ground, both control signals need to be first fed into the low side driver. A communication link comprised of small pulse transformer with an isolation level of 1200V is provided to allow the high side driver to receive its control signal from the low side driver. The link also transforms the fault signal detected by the high side driver back to the low side for output to the protection logic.

When an IGBT overcurrent occurs, the voltage across the IGBT increases. In this case, the IGBT gate signal has to be removed as soon as possible in order to protect the device from damage. This is implemented via desaturation circuits for both high and low side IGBTs. The described circuits use a resistance network to monitor the IGBTs forward voltage and send out signals to turn off the IGBTs when the on-state voltage exceeds a certain threshold.

6.3.5 Implementation Results

Fig. 6.16 illustrates one measured open loop phase current of the control winding when the power inverter is fed with three phase sinusoidal reference signals. The resulting current waveform in the control winding is satisfactory, aided by the high
inductance of the 2-pole winding. Figs. 6.17 and 6.18 show the commanded input current signal and the closed loop inverter output current during a transient for a 100 l/min step change in speed command. These two currents are purposely split into two figures to demonstrate their similarities. A switching frequency of 10kHz is sufficient to regulate the BDFM control winding currents based on the commanded values from the outer loop microprocessor 80960KB. Due to the outer loop speed control, the control winding currents shown in Figs. 6.17 and 6.18 are no longer pure sinusoids. The currents shown in Figs. 6.17 and 6.18 are obtained from the 80196KR microprocessor.

Figure 6.16 Measured Control Winding Current $i_{a2}$ Realized by the Inverter
Figure 6.17 Commanded $i_{a2}$ Signal From 80960KB

Figure 6.18 Measured $i_{a2}$ From the BDFM Control Winding
6.4 Hardware Buffer

The most popular inter-microcontroller communication method is the RS232 protocol. The serial nature of RS232 communication makes it very slow and a prime obstacle in obtaining a high control bandwidth when large amounts of data need to be transferred. Instead of using an RS232 interface, the laboratory inverter uses a hardware buffer between the 80960KB and 80196Kr microprocessors for data and command transfers. A block diagram is shown in Fig. 6.19 and the detailed schematic is given in Appendix E. The buffer has four 16 bit ports which can store command values issued from the 80960KB, and one 16 bit port for one variable feedback from 80196Kr.

![Figure 6.19 Hardware Buffer Block Diagram](image-url)
The 80960KB uses three ports to output instantaneous control current signals for the power inverter to control the currents in the control windings. The 80196Kr uses one port to feed back a measured BDFM control winding phase current to 80960KB for data acquisition purposes. As stated earlier, although Fig. 6.19 shows 5 hardware buffers, any two of them can be combined for a 32 bit variable transfer. When the 80960KB needs to store an updated command value in the buffer, its address bus drives the 80960KB decoder logic block to select the correct port for the command. In the meantime, it disables the 80196Kr from accessing the appropriate port during data transfer. This process takes place within a 0.5 µs window, which is very small compared to the 10kHz PWM update frequency in the 80196Kr. The 80196Kr continuously monitors the buffer ports for updated command values.

6.5 80960KB Microcontroller Subsystem

6.5.1 QT960 Core System [48]

The QT960 evaluation board is illustrated in Fig. 6.20. The on board 80960KB is an embedded full 32-bit processor with an integrated floating-point unit, originally developed as a central computing unit for embedded control system applications. The 80960KB design also features RISC, on-chip instruction cache and a large memory address space. Unlike conventional microcontroller designs which integrate all slow I/O and other peripherals inside the chip, the 80960KB only incorporates the high speed core computing unit. The QT960 evaluation board uses an Intel 82380 DMA controller to
support all system peripherals. When the 80960KB needs to interact with any slow peripheral, a wait state generator in the 82380 is programmed to insert enough wait signals to the 80960KB local bus for the slow peripheral to catch up. Then the 80960KB resumes its high speed calculations.

![Figure 6.20 QT960 Evaluation Board Block Diagram [48]](image)

A monitor program resides in an 128K byte on board EPROM and uses an Intel 82510 UART to communicate with a host personal computer for downloading and debugging purposes. Because the 80960KB is running at a very high speed, both wire wrap area and local bus connectors are provided on the evaluation board to minimize any noise problem caused by circuit expansions.
6.5.2 Position Encoder and Interface Circuit

A Teledyne Gurley 25/04S 14 bit absolute position encoder capable of providing a 1MHz parallel data output rate is used for BDFM shaft position feedback. Fig. 6.21 shows the block diagram of the interface circuit between encoder and the 80960KB microprocessor. Appendix F shows the detailed circuit schematic. To obtain position information, the 80960KB addresses the appropriate encoder memory location to initiate conversion and subsequently obtain the output value. The encoder is pictured in Fig. 6.22 at the lower right corner connected to the shaft of the DC machine.

![Figure 6.21 Position Encoder Interface Block Diagram](image)

6.5.3 \( \theta_e \) Measurement Circuit

The BDFM dynamic control algorithms design described in chapter 5 require information of the synchronous angle \( \theta_{e2e} \), which in turn necessitates measurement of \( \theta_e \) or \( \Sigma \omega_e \Delta t \). Measurement of \( \theta_e \) is preferred over the summation process of \( \Sigma \omega_e \Delta t \), which may accumulate errors caused by the measurements of \( \omega_e \) and \( \Delta t \). As indicated by Eqn.
(5.2), $\theta_6$ can be obtained by measuring the phase voltage, $V_{6a}$ of the power winding. An instantaneous voltage transducer transforms the high voltage $V_{6a}$ into a low voltage signal. High frequency noise components are filtered out by a low pass filter. A fast A/D board, which is the same as those described in Fig. 6.9, converts the analog $V_{6a}$ signal to its digital form. The 80960KB microprocessor then reads the digital $V_{6a}$ value at a preset control cycle time and calculates $\theta_6$ according to Eqn. (5.2). The instantaneous transducer and the low pass filter may cause a phase shift on the $V_{6a}$ signal. Because this phase shift is constant for a 60Hz signal, it can be easily compensated for by an offset within the 80960KB microprocessor. Figs. 6.23 and 6.24 show the $V_{6a}$ signals before and after the low pass filter. Apparently the low pass filter is effective in filtering out the high frequency noise caused by a six step DC machine drive connected at the same grid point.
Fig. 6.25 illustrates the $V_{6a}$ signal after the sample and hold chip (HA5320) on the A/D board. The sample and hold moments are synchronized to the control cycle of 1ms.

**Figure 6.23** The Power Winding Phase Voltage $V_{6a}$ Signal Before Being Filtered

**Figure 6.24** The Power Winding Phase Voltage $V_{6a}$ After Filter
6.5.4 BDFM Control Algorithm Implementation

Fig. 6.26 shows the flowchart for implementing the BDFM control algorithms discussed in chapter 5. After start up and initialization, a timer in the Intel 82380 DMA is set to run at 1 ms interrupt rate as a time stamp for real time calculation. Initial values for the control algorithm (i.e. BDFM machine parameters) are also loaded into the microcontroller. The 80960KB interrupt priority is reset from its supervisor mode to user mode to allow any programmed interrupts to be acknowledged.

Once initialization is finished, the microprocessor automatically uses the position encoder output to calculate rotor speed at every timer interrupt request. If a control algorithm requires the estimation of electric torque, $T_e$, the power winding subsystem estimation and the subsequent $T_e$ calculation will use the speed information and nominal
power winding quantities (230V and 60Hz). If the control loop is commanded to be closed by a signal from the host personal computer, control commands (three phase instantaneous current signals for the control winding) are calculated using the synchronous angle $\theta_{se}$ and information from the power winding estimator, speed command and measured speed signals. In open loop operation, the control command values (frequency and magnitude) can be input via the host personal computer. The new commands are then sent out to the hardware buffer. This completes a control loop and the 80960KB goes back to the power winding estimation calculation, if it is necessary.
The loop control update frequency can be up to 2kHz for algorithms with no $T_e$ estimation.
Chapter 7
Conclusions and Recommendations

This thesis describes the development of dynamic control algorithms for the BDFM. This chapter gives a conclusion of the thesis and recommendations for future work.

7.1 Conclusions

Because BDFM synchronism possesses a straightforward relationship between the shaft speed and the control frequency, simple scalar controls such as constant volts/hertz and current controls are feasible. Both methods are limited by their maximum torque curves, beyond which stable synchronous operation is not possible. The controller design based on the constant current control method is easier to implement, since the complexity of the dynamic equations is reduced compared to the constant volts/hertz control method.

Simulation and experimental results show that the open loop constant current control for the BDFM has a poor dynamic response. Closed loop control can prevent the BDFM from going into the induction mode of operation, which offers a lower torque margin and reduces BDFM efficiency. The control winding current magnitude can be utilized to control a steady state performance index such as machine efficiency, minimum control power, power winding power factor or a combination thereof. This control method is only suitable for low performance industrial systems like pump or fan drives.
A first order model reference adaptive control method is developed based on a direct torque control algorithm [20, 21] to further expand the system robustness against changes in load and inertia. Simulation results show its effectiveness. The BDFM voltage source model is replaced with its current forcing model to simplify subsequent control designs and implementations.

A BDFM synchronous reference frame dq model is derived in this thesis to obtain DC quantities. The BDFM can mathematically be divided into two subsystems based on the assumption of a magnetically linear rotor circuit. These two subsystems are referred to their own synchronous frames. The electric torque can now be expressed in terms of quantities on these synchronous frames and the BDFM synchronous angle, which consists of the angles between these frames and the rotor. The synchronous angle stays constant during steady state operation.

A rotor flux field oriented control method is derived based on the proposed synchronous reference frame model. As a result, the BDFM electric torque can be controlled via the torque component of the control winding current, while control winding flux is maintained by the flux component. Simulation results show excellent dynamic response and robustness against electrical parameter variations. The assumption of negligible power winding stator resistance contributes to small oscillations on the speed response. In the laboratory implementation, estimated electric torque information is used to formulate a control algorithm, and experimental results justify the BDFM synchronous reference frame model development.

Further investigation shows that the BDFM synchronous angle can be utilized to represent the electric torque behavior during transients. Two control algorithms are
constructed using the angle, yielding a control update cycle of 0.5ms. The experimental results illustrate that a 100 1/min step speed response can be achieved in about 200ms.

Since the control algorithms using the BDFM synchronous angle are as simple as those for conventional induction machines in terms of required computational and hardware resources, modern economic microprocessors can be used to build practical control systems for the BDFM. This, in combination with its other advantages, makes the BDFM system competitive with its conventional induction machine counterpart.

7.2 Recommendations For Future Work

This thesis has formulated the concept of field oriented control for the BDFM. Simulations and experiments have been used to verify speed control algorithms based on the rotor flux oriented method. This section identifies several additional applications and control implementations.

7.2.1 BDFM Generator System

Although the BDFM vector control theory is verified only in motor control applications in this thesis, it is applicable to generator controls. In a generator system, the multiplication of the BDFM shaft torque and its speed determines its active input power. The BDFM shaft torque has to correspond to the input power which may be determined by the prime mover or demanded by the load. Therefore, the BDFM desired
torque command can be generated by characteristics of either the prime mover or the load.

If a bi-directional converter is used to excite the BDFM control winding, it can be controlled in a way that only active power is processed. This leaves the reactive power being processed through the power winding side. Since there are two control variables, $i_{q2e}$ and $i_{d2e}$, on the control winding synchronous reference frame, they can be used to control the BDFM reactive power flow via the power winding and its electric torque. The electric torque corresponds to the active power which balances either the prime mover input power or the load demand.

### 7.2.2 BDFM Rotor Flux Oriented Control with A Voltage Source Converter

When a voltage mode converter is used in place of a current mode converter to control the BDFM, the control commands should be derived in terms of voltages instead of currents. Although the BDFM rotor flux oriented control requires current control on the control winding, the winding terminal voltages can be derived from these current commands using the winding voltage equations. These winding terminal voltages are then sent to the voltage mode converter for voltage regulation.

### 7.2.3 A BDFM Stator Flux Oriented Control Algorithm

Careful examination of Eqns. (4.30) and (4.31) reveals $\psi_{q2e}$ and $\psi_{dr2e}$ to be rotor fluxes. Subsequent control algorithm development is based on rotor flux orientation.
If the control winding stator fluxes are defined instead, a stator flux oriented control can be formulated. Compared with rotor flux oriented control, stator flux is easier to be measured and controlled.

7.2.4 A Sensorless Speed Observer for BDFM

The BDFM speed control systems as presented require a position encoder on the rotor shaft to feed back speed or position information. The encoder increases the complexity and cost of the whole system. It would be desirable to replace the encoder with a speed observer. When a BDFM drive does not need fast dynamic response, rms values from the machine terminals can be used to estimate the rotor shaft speed. This will make the low performance BDFM drive more compact and less costly.

7.2.5 Remarks

This thesis verifies the concept of field oriented control for the BDFM with simulation and experimental results over a limited speed range. In order for the control to be practical for the BDFM, future work has to be done for the entire speed range, in which some electrical parameters in the BDFM may vary more. The BDFM dq model does not consider machine saturation, which may cause more parameter variations. Therefore, the robustness of the described control algorithms against parameter variations has to be investigated.
The inverter built for the laboratory verification of the BDFM field oriented control algorithms can only process power in one direction. Since power flow associated with the BDFM control winding can be positive or negative during normal synchronous operation, a resistive breaking circuit was built to dump the output energy from the control winding. This leads to a low efficiency for the entire BDFM system. A converter equipped with regeneration capability is needed for better power transfer.

Further work needs to address controller optimization and packaging to maximize the performance/cost ratio. Integration of controller and converter will also yield more compact and economic systems. Replacement of the absolute position encoder with an incremental one will also help bring down the cost.

Control algorithm development in this thesis has mainly focused on how to control the BDFM electric torque. This forms the basis for development of outer loops to control speed, position or another performance parameter. Outer control loops can be based on simple PI regulators or advanced control techniques, such as model reference, fuzzy or even neural-networks.
Bibliography


Appendices

- Circuit Schematics
NOTE: FOR R1=R3 AND R2=R4,
V0 = Vin * R2 / R1
FOR VCC=12V, Vin = 6V WHEN NO CURRENT FLOWS
R1 AND R2 SHOULD BE SELECTED THAT V0 = 5V
IN ORDER TO INTERFACE WITH 0-10V A/D CONVERTERS
NOTE: JP2 MEMORY I/O EXPANSION CONNECTOR ON EV90C196GP BOARD IS RE-DEFINED AS JP1 HERE.

CONNECTOR TO AID BOARD 1: DEFINED AS JP2
CONNECTOR TO THE HARDWARE BUFFER 1: DEFINED AS JP3

CONF: SIGNAL FOR ALL A/D BOARDS TO DO CONVERSION AT THE SAME TIME (ADDRESS: E000H)
AID 1: READ 1ST AID BOARD RESULT (ADDRESS: 0400H)
AID 2: READ 2ND AID BOARD RESULT (ADDRESS: 0000H)
AID 3: READ 3RD AID BOARD RESULT (ADDRESS: 0000H)
BUFFERS: READ BUFFER PORT 1 (ASSIGNED TO CONTROL_Magnitude, 0000H)
BUFFERS 2, 3, & 4: READ BUFFER PORT 2 & 3 (ASSIGNED TO CONTROL_FREQUENCY, 0000H)
BUFFERS 5, 6, & 7: READ BUFFER PORT 4 & 5 (ASSIGNED TO CONTROL_VOLTAGE, 0000H)
BUFFER 8: SEND DATA TO BUFFER PORT 8 (ADDRESS: 0000H)

Appendix D: 80C196KR Interface