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Vector control of an induction machine

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VECTOR CONTROL OF AN INDUCTION MACHINE

A Thesis

Presented to

The Faculty of the Department of Electrical Engineering

San Jose State University

In Partial Fulfillment

of the Requirements for the Degree

Master of Science

by

John David Loud

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ABSTRACT

VECTOR CONTROL OF AN INDUCTION MACHINE

by John David Loud

Precision motor control has until recently only been accomplished by use of DC or synchronous machines. While the performance attained with these machines is good, they are expensive to manufacture, require regular maintenance and have a limited range of speed. Only with the advent of high-power, fast-switching electronic devices in conjunction with fast microprocessors to perform real-time control has it become possible to achieve precise control of an induction motor.

The goal for this project was to review the esoteric theoretical literature to glean sufficient practical information to construct a controller for a real-world application. This controller was developed for United Defense LP. Using a high-performance, digital-signal-processing (DSP) floating-point microprocessor and inexpensive power electronics, the vector control theories were implemented and evaluated. It was demonstrated that it is possible to achieve servo-motor-like control of a two-horsepower induction motor over an extremely wide range of speed. Control strategies for attaining high efficiency were also implemented.

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CHAPTER 1

History and Definitions

1.1 Control of Electro-Mechanical Machines

Since its invention in the late 1880s by Nikola Tesla, the induction motor has become industry's workhorse. Compared to other types of motors, induction motors are rugged, low-cost and capable of handling high power [Ryff and Platnick, 264]. Until recently, these motors could only be used in constant-speed applications, such as driving pumps or fans, connected directly to a utility power line. For applications requiring a discrete alternate speed, adjustment was achieved by mechanical means such as a gear box. Adjustable speed was achieved with an electronic control unit commonly referred to as a driver or inverter. Only in recent years have these inverters contained more than a handful of power semiconductor devices such as SCRs. The performance of such drivers was limited due to their slow-switching power devices and crude control circuitry. For these reasons, DC motors or synchronous machines were used for applications that required precise motion control.

Early inverters used an electronic unit capable of producing a variable frequency voltage. These devices usually controlled the motor speed in an open-loop

fashion. The transient response and low-speed performance of this type of controller was poor even when the control loop was closed via a speed measurement device. These performance deficiencies resulted from basing the controller on the steady-state characteristics of the motor.

With the advent of modern drivers and the recent development of “vector control theory” [Vas, 2], induction motors can be controlled just as precisely and accurately as DC or synchronous motors. With the application of this new control theory and the falling cost of power semiconductors, vector-controlled induction motors have become an attractive alternative to DC and synchronous motors.

The most important merit of vector control is its ability to precisely and instantaneously control output torque well beyond the rated speed of the motor. In addition, vector control allows full and stable control of torque down to zero speed. Because of these properties, a vector-controlled induction motor can be used in conjunction with a relatively simple feedback control scheme (such as PID or state feedback) to achieve a high level of performance. The real advantage of vector control is that all the merits of induction motors (low cost, efficiency, ruggedness etc.) can now be realized in demanding motor control applications. The following is a partial list of applications in which vector-controlled induction motors are presently used [Baldor]:

1.1.1 Line Regeneration Applications

- testing of turbine impellers - 90kW line regeneration - up to 25000rpm - installed in Taiwan;
- test stand for jet aircraft engines - 160kW line regeneration - installed in BMW plant in Germany;

1.1.2 Precise Speed and Torque Regulation Applications

- food processing plant - Keebler O'Boisies Potato Chips - eight vector drives moving 5000lbs of potatoes per hour;
- mining conveyors - for speed control and line regeneration on long downhill runs - 10hp through 150hp;
- motor machining equipment - 10hp feed and speed controllers installed in all Baldor manufacturing plants;
- used in dynamometers - loading and testing automotive transmissions on assembly lines;

1.1.3 Precise Position Control Applications

- Phelps Dodge Copper Mine/Smelter - control of a 60-foot diameter anode casting wheel for acceleration/deceleration and position control for each pour;
 - US Army Corps of Engineers - used to tow barges into position in the Mississippi lock and dam system;
-

- wood products industry - veneer peeling machine - vector control proved to be the only viable means of reliably achieving high torque at very low speeds;
- elevator drives - providing regeneration and torque down to zero speed making the mechanical brake redundant instead of necessary;
- Dolco Plastics, Dallas, Tex. - plastic molding machines - 10hp and 15hp;
- eddy current drives - glass plant - large conveyor control;

1.1.4 High Reliability Applications

- New United Manufacturing Inc. (NUMI) - automobile manufacturing plant - among other applications, control of the main assembly line (1car/min);
- electric automobiles - all major manufacturers have chosen to use a vector-controlled induction motor to achieve high torque from zero to rated speed, thus eliminating the need for a transmission.

A few issues must be evaluated prior to deciding whether to implement a vector drive system. One is that a moderately priced power electronics inverter with feedback controller is required for operation. This is to some extent offset by the benefits of an induction motor. Further issues are stability and reliability. As with all feedback control systems, there is a potential for instability in the event of a component failure in the controller or feedback systems though with the intended application in mind, steps can be taken to ensure a robust and partially fault-tolerant system [Liu, Fu and Lipo].

1.2 Symbols and Abbreviations Defined

The following symbols will be used throughout this document and are defined as follows:

a : Subscript a denotes that this quantity applies to the armature

a : The number of paths for the armature current

AC: Alternating current

ASD: Adjustable speed drive

AWG: American wire gauge

a/d: Analog to digital converter

b : Subscript b denotes that this quantity is measured with the rotor blocked

B : Magnetic flux density

c : Subscript c denotes that this quantity applies to the core

d : Subscript d refers to the direct component

d/a: Digital to analog converter

DC: Direct current

DSP: Digital signal processor

e : Subscript e denotes that it is the electrical aspects of this quantity that are referenced

f : Stator supply voltage frequency

F : Force in newtons

FOCA: Flux optimization control algorithm

GE: General Electric

Hz: Cycles/second

i or I : Current in amperes

IGBT: Insulated gate bipolar junction transistor

I/O: Input/output

j : A complex number defined to be $\sqrt{-1}$

K : A DC motor parameter constant of proportionality

ℓ : Subscript ℓ denotes leakage (as in leakage flux)

l : Subscript l denotes loss (as in power loss)

L : Subscript L denotes that this value quantity applies to an inductor

L : Inductance in henrys

lbf: Slug

m : Subscript m denotes that this quantity is mutually related

MRAS: Model reference adaptive system

n_s : Synchronous speed in $\frac{\text{radians}}{\text{sec}}$

N : The number of turns in the armature winding

Nm: Newton-meters

PD: Proportional derivative

PI: Refers to proportional-integral control

PID: Refers to proportional-integral-derivative control

PWM: Pulse-width modulated

r : Subscript r denotes that this quantity applies to the rotor

r : Superscript r denotes that this is expressed in the rotor reference frame

r^Ψ : Superscript r^Ψ denotes that this quantity is expressed in the rotor-flux reference frame

ρ : C code equivalent of ρ

P : Number of stator poles

q : Subscript q refers to the quadrature component

rpm : Revolutions/minute

R : Resistance in ohms (Ω) or reluctance in $\frac{1}{\text{henry}}$

RAM: Random access memory

s : Slip

s : Subscript s denotes that this quantity applies to the stator

s : Superscript s denotes that this is expressed in the stator reference frame

S : Laplace operator ($j\omega$)

T : Torque in newton-meters (Nm)

x : Subscript x states that this quantity is in the x direction.

V : Electric potential in volts

Wb : Webers

X : Complex reactance

y : Subscript y states that this quantity is in the y direction.

Z : Impedance

Ψ : Flux in webers

ρ : Rotor flux angle in radians

$\dot{\rho}$: Rotor flux speed in $\frac{\text{radians}}{\text{sec}}$

θ : Rotor position with respect to the stator in radians

$\dot{\theta}$: Rotor angular speed in $\frac{\text{radians}}{\text{sec}}$

ω_r : Rotor angular velocity (also $\dot{\theta}$)

Ω : Ohms

μ : 10^{-6}

k : 10^3

$\approx$: Approximately equal to

$^\circ$: Degrees

CHAPTER 2

Induction Motors

2.1 Mechanical Structure

The most common type of induction motor is the three-phase, squirrel cage rotor induction motor [Ryff and Platnick, 264]. This type of motor is comprised of three major parts: the stator, the rotor and the end caps with bearings. Figure 2.1 shows a cross-sectional view of a three-phase induction motor with, for simplicity, the stator windings shown as a single conductor. In reality, the windings are distributed in the stator and provide flux similar to what would be expected from current through the single conductor windings shown.

2.1.1 The Stator The stator is made up of thin laminations of a highly permeable steel that together yield high magnetic flux density and low core losses. The stator windings are wound and placed into slots in the stator, 120° apart in space. These are wound so as to create stator magnetic poles when current flows through them. The number of stator poles in conjunction with the frequency of the applied three-phase power determine the speed of rotation of the stator's magnetic field and thereby the speed of the induction motor [Ryff and Platnick, 267].

Figure 2.1: Cross Section of an Induction Machine

Figure 2.2: Resultant Vector Representation of the Three Phase Currents

Figure 2.2 shows a static picture of the flux and current directions when only winding A of the stator is energized. While normal operation of a wye-connected induction machine precludes having current through only one winding, this drawing illustrates the resultant flux vector. During normal operation, three currents, oriented 120 degrees apart in time, are applied to the three windings and can be combined to form a $\overrightarrow{MMF}$ stator vector. The stator flux vector $\overrightarrow{\Psi}$ is found to be $\overrightarrow{\Psi} = \overrightarrow{MMF}/R$ where R is defined to be the magnetic circuit reluctance [Chai]. It is important to note that a balanced three-phase current applied to the stator windings results in a rotating stator current and therefore a rotating flux vector. The magnitude of the $\overrightarrow{MMF}$ stator vector is constant as it rotates around the stator [Ryff and Platnick, 267]. Since the stator current vector rotates within a plane, it can be represented by two components: x and y . Analysis can be greatly simplified if we consider that the stator current has only these two components and represented it as:

$$I_s = \begin{bmatrix} I_{sx} \\ I_{sy} \end{bmatrix} \quad (2.1)$$

2.1.2 The Squirrel Cage Rotor The squirrel cage rotor (so named for the conductor's resemblance to the exercise wheel found in a squirrel cage) is also made up of thin laminations of a highly permeable steel that together yield high magnetic flux density and low core losses. The rotor bars, or windings, are placed in the rotor slots and shorted together at each end as shown in figure 2.3. The rotor is mounted

Figure 2.3: Conducting Rotor Bars without the Rotor

on a shaft so that it can rotate and is placed inside the stator. The number, size, skew and depth of recess of the rotor conductors influence motor performance.

2.1.3 End Caps and Bearings The end caps and bearings serve to position the rotor and allow it to rotate inside the stator. Since air has very high magnetic reluctance, the rotor must be positioned very close to the stator to achieve maximum flux coupling and high efficiency. The bearings, which are secured to the end bells, must keep the rotor centered and able to spin freely inside the stator.

2.2 Basic Concepts of Induction Motor Operation

Electrical power is delivered to the induction motor through the stator windings. The three-phase power induces a rotating magnetic flux in the stator. This rotating flux passes through the rotor and cuts the rotor bars. As a result, a current is induced in the rotor bars, limited only by their resistance. The interaction

Figure 2.4: Torque vs Rotor Speed/Synchronous Speed

of the rotor current and the flux produces torque, and the rotor rotates towards the flux. With no load, the rotor spins at almost the same speed as the rotating stator field. Since the speed differential between the rotor and stator is small, the induced voltage in the rotor bars is small, and the rotor currents are small. As more torque is demanded from the rotor, the speed of the rotor is reduced and, as more flux lines cut rotor bars, higher rotor current is induced. Higher rotor current results in greater torque. In general, it is this speed differential, or slip, between the speed of rotation of the rotor and the speed of rotation of the stator magnetic field that determines motor torque. A typical torque-speed curve of an induction motor is shown in figure 2.4 where the horizontal axis is the speed of the rotor expressed as a percentage of the speed of rotation of the flux (the synchronous speed). Note that an induction motor must have slip to develop torque. No flux lines would be cut by the rotor conductors were they rotating at the synchronous speed [Ryff and Platnick, 264-271].

As shown in equation 2.2, the synchronous speed n_s depends on the stator

supply voltage frequency f , and the number of stator poles P in the following relationship.

$$n_s = \frac{f}{P/2} \quad (2.2)$$

The actual rotor speed n_r is dependent on the load and can vary from a value just below n_s at no load to zero for a locked rotor. Slip s is defined as

$$s = \frac{n_s - n_r}{n_s} \quad (2.3)$$

and can vary between 1 for a locked rotor to near 0 for no-load operation.

Current flows in the rotor by induction – hence the name induction motor. Most other motors (DC or synchronous) require an electrical connection between the rotating assembly and external circuitry. The induction motor's lack of a physical connection to the rotor eliminates potentially hazardous contact sparks as well as the high production and maintenance costs associated with brushes, slip rings and commutators. The simplicity of rotor construction makes induction motors robust, reliable and inexpensive.

2.2.1 Torque Production in an Induction Motor In a motor, torque is generated from the interaction between current and flux or, more precisely, torque is the cross product of current and flux as illustrated in figure 2.5. The force that produces this torque is known as the Lorentz force and is defined as:

$$\vec{F} = \oint I d\vec{\ell} \times \vec{B} \quad (2.4)$$

Figure 2.5: Vector Relationship of Torque, Flux and Current

where $\vec{F}$ is force, I is current, $d\vec{\ell}$ is the length and direction of the conductor and $\vec{B}$ is the magnetic flux density. For a straight conductor, equation 2.4 reduces to:

$$\vec{F} = I \vec{\ell} \times \vec{B} \quad (2.5)$$

It is clear from equation 2.5 that to obtain maximum torque for given levels of flux and current, the direction of the current should be kept exactly 90° to the direction of the flux. In a DC motor, the perpendicular relationship between flux and armature current is accomplished by means of a commutator. Figure 2.6 shows the mechanical structure of a wound field DC motor. The separate armature and field windings in a separately excited DC motor yield the simple torque relationship:

$$T = K I_a \Psi \quad (2.6)$$

Figure 2.6: Cross-Sectional View of a DC Machine

Equation 2.6 defines torque T for a DC motor in terms of K , a motor parameter constant of proportionality; I_a , the armature current; and Ψ , the flux per pole (Wb). K is defined as $K = \frac{PN}{\pi a}$ where P is the number of stator poles, N is the number of turns in the armature winding and a is the number of parallel paths for the armature current. Flux Ψ is almost linearly proportional to DC field current (until saturation). The beauty of the DC machine lies in the simplicity of its torque control governing equation. Precise control of the DC field current controls the flux. Precise control of the armature current controls the torque. It is important to note that torque control is solely a function of armature current once a field has been established. Torque is not dependent on the speed of rotation of the armature [Chai, 4.25].

Figure 2.7 shows the structure of a synchronous motor. In a synchronous

Figure 2.7: Cross-Sectional View of a Synchronous Machine

motor, the rotor flux vector is set up by a rotor field winding or a permanent magnet on the rotor. The stator establishes a rotating flux vector based on the sum of the stator currents. The rotor rotates in synchronization with the stator flux vector. Torque control is not normally used in synchronous motor control because rotor speed is constrained by the speed of rotation of the stator flux field. The maximum torque available (at the speed of rotation of the stator flux field) is determined by the applied voltage and the power angle. The synchronous machine illustrates the concept of torque development between the magnetic rotor flux vector and the rotating stator flux vector, where the only unknown quantity in evaluating the amount of torque being delivered to the shaft is the torque angle [Guru and Huseyin, 298-305].

Torque generation in an induction motor operates on the same Lorenz force

principle that governs both DC and synchronous motors. With an induction motor however, it is much more difficult to generate the proper interaction between flux and current as the control action is accomplished entirely through the stator winding current. The stator current supplies the rotor current by induction, but it also supplies the stator flux needed to produce torque. The direct implementation of vector control requires the determination of both the modulus and direction of the representative rotor flux vector. Despite this requirement, there is no practical way of directly measuring this rotor flux vector's position relative to the stator flux vector [Bellini and Figalli]. To control torque - which, analogous to the DC and synchronous machines is given by the cross product between the stator current and rotor flux vectors - precise control of these two values must be attained. To further complicate the issue, this operation must be accomplished in a dynamic way through a rotating stator current vector. The control laws which make this possible will be developed in section 3.

That rotor current is induced and that it interacts with stator flux to produce torque have been established. Figure 2.8 shows a very simple case (a rotor which has only one pair of rotor bars) to schematically clarify the torque generation process. As shown, there is a rotating stator flux field with poles labeled N and S. This rotating stator flux field extends through the stationary rotor. A voltage is induced in the rotor bars as they are cut by the moving lines of flux, resulting in a current in the rotor bars limited only by the impedance of the rotor. This induced

Figure 2.8: Torque Development in an Induction Motor with a Single-Coil Rotor current sets up the rotor flux field orthogonal to the stator flux field. The two vector components that sum to form the rotor flux are labeled "Flux from Stator I" and "Flux from Rotor I". As shown in equation 2.5, this results in the force depicted. In a real rotor, many pairs of rotor bars react to the rotating flux. It is the vector sum of the currents in the rotor bars which causes the resultant rotor flux vector [Ryff and Platnick, 271].

2.3 Induction Motor Equivalent Circuit

In this section, a transformer will be used as an analogous example to derive the steady state induction motor equivalent circuit. As shown in figure 2.9, an induction motor is in many ways similar to a transformer. A simple transformer has no electrical connection between the primary and secondary windings; nor is

Figure 2.9: Rotor Response Analogous to a Transformer

there a direct electrical connection between the stator and rotor windings in an induction motor. The current in the rotor I_R can be found by dividing the induced rotor voltage by the rotor impedance. Given the slip s (as defined in equation 2.3), the induced blocked rotor voltage V_{br} and the rotor impedance - made up of the leakage reactance sX_r (X_r is the blocked rotor reactance) and the rotor resistance R_r - it is found that:

$$I_r = \frac{sV_{Br}}{\sqrt{R_r^2 + (sX_r)^2}} \quad (2.7)$$

Note that I_r is a function of the product of slip and the induced blocked rotor voltage. By dividing equation 2.7's numerator and denominator by the slip, it becomes:

$$I_r = \frac{V_{br}}{\sqrt{(\frac{R_r}{s})^2 + X_r^2}} \quad (2.8)$$

I_r is now a function of induced stator voltage. By substituting $R_r + R_r(\frac{1-s}{s})$ for $\frac{R_r}{s}$, the rotor impedance can be expressed as:

$$Z_r = R_r + X_r + R_r(\frac{1-s}{s}) \quad (2.9)$$

which is shown in the rotor equivalent circuit of figure 2.10 [Ryff and Platnick, 273].

Figure 2.10: Rotor Equivalent Circuit

Within the transformer analogy, the stator windings can be considered the primary winding of the transformer. One significant difference between the transformer model and the induction motor model is that an induction motor typically requires 30% to 50% of the stator current for the magnetizing current while a transformer typically requires 5% or less. Representing core losses as a resistor R_c , the stator equivalent circuit can be modeled as shown in figure 2.11.

Using the transformer technique of reflecting the secondary circuit into the primary, one phase of the induction motor model can be reduced to the equivalent circuit shown in figure 2.12. The motor parameters R_r and L_r are reflected through the transformer to form this equivalent circuit.

It is important to note that this equivalent circuit model is only valid when evaluated with the following simplifying assumptions:

1. a sinusoidal distribution of MMFs;

Figure 2.11: Stator Model

Figure 2.12: Phase Model Equivalent Circuit

2. a linear magnetic circuit;
3. a uniform air gap in the induction machine; and
4. that core losses are accurately represented by R_c .

Given these assumptions, the transfer function can be calculated in the Laplace domain using the Laplace operator S . Referring to figure 2.12, the following impedances are defined:

$$A(S) = \frac{SL_m R_c}{SL_m + R_c} \quad (2.10)$$

$$B(S) = \frac{R_r}{s} + SL_r \quad (2.11)$$

$$C(S) = R_s + SL_s \quad (2.12)$$

Using a voltage divider, the voltage $V_A(S)$ measured across R_c is found to be:

$$V_A(S) = V_{in} \left[\frac{\frac{A(S)B(S)}{A(S)+B(S)}}{C(S) + \frac{A(S)B(S)}{A(S)+B(S)}} \right] \quad (2.13)$$

The current through R_c and L_m is found to be:

$$I_A(S) = \frac{V_A(S)}{A(S)} \quad (2.14)$$

and the stator current is found to be:

$$I_s(S) = \frac{V_{in}}{C(S) + \frac{A(S)B(S)}{A(S)+B(S)}} \quad (2.15)$$

Using $L_m = 0.07H$, $L_s = L_r = 0.05L_m$, $V_{in} = 100volts$, $s = 1$ (locked rotor) and $S = j\omega$, the stator current and phase are shown in figure 2.13. As expected,

Figure 2.13: Stator Current and Phase vs Frequency for Locked Rotor

the stator current starts high and drops off as frequency increases. In addition, the stator current's phase approaches -90° . The current and its phase through the magnetizing branch comprised of L_m and R_c are shown in figure 2.14. As expected, the current through the magnetizing branch decreases with frequency as the impedance of L_m increases.

It is difficult to draw conclusions from this model beyond the superficial observations made. It was found to be useful in validating data measurements with the induction motor running in steady-state operation and was used to validate motor parameters.

2.4 Motor Parameter Identification

The goal for this section was to test the induction motor to identify values for the L-R circuit shown in figure 2.12. This steady-state model represents a single

Figure 2.14: Current and Phase through the Magnetizing Branch

phase of the induction motor, reflected to the stator. To identify this motor's parameters, the voltage-current relationship was measured at various speeds and loads. The values for the model were then adjusted to more accurately reflect the data observed. It was assumed that R_s and R_c do not vary and that the rotor and stator leakage inductances (L_{rl} , L_{sl}) are approximately 5% of the value of the mutual inductance L_m .

2.4.1 Synchronous Test The first measurements were made with the motor operating at synchronous speed. Synchronous operation was modeled first because in this mode slip is 0, and R_r and L_{rl} drop out of the model. Synchronous operation was achieved by setting slip speed to 0 in the software control program. This forced the stator flux to rotate in synchronization with the rotor. An advantage of synchronous operation is that a motor model is not used, so these measurements

can be made without knowledge of any motor parameter except the number of poles.

The DC motors were used to spin the induction motor at the desired speed. The flux was manually adjusted while the motor was spinning. As the flux was adjusted, a slight variation in motor speed was observed, meaning the rotor current was not exactly zero and that the induction motor was producing a small torque. This difference between the rotor speed and the electrical speed was traced to loop-speed resolution errors. The technique was improved by adding a small constant slip to the rotor flux angle until no variation in motor speed was observed with variations in flux. In fact, no variation in motor speed was observed when power to the induction motor was turned on and off, indicating that synchronous speed had been attained. Measurements with the improved technique varied by only a few percent from the original measurements. Measurements and calculations are contained in the Data and Calculations appendix.

When the measured voltage-current relationship was iteratively compared to and adjusted to fit the motor model, the best match was obtained when:

$$L_{st} = 0H$$

$$R_s = 1\Omega$$

$$R_c = 300\Omega$$

and L_m was found to be a function of magnetizing flux:

$$L_m = 0.22 - 0.2\Psi$$

As shown in the Mathcad work sheets in the appendices, there is a large phase error in some of the measurements. This can be due to several factors: there is significant ripple and noise on the voltage waveform and it is difficult to measure the phase more accurately than $\pm 20\%$, especially at lower flux levels; L_{sl} does not equal 0; R_c changes with stator flux density and temperature; the idealized model does not exactly match the induction motor; and the rotor current isn't exactly zero since the controller cannot perfectly synchronize the electric field with the moving rotor. The most significant is the voltage measurement error.

2.4.2 Complete Model Test Further measurements of the voltage-current relationship were made with the motor running at a constant speed with a constant load. These measurements and related calculations are shown in the Mathcad work sheets in the appendices. They were used to iteratively determine the value of the rotor leakage inductance L_{rl} and the rotor resistance R_r . The core loss, copper loss and mutual inductance were assumed to be the same as previously determined. It is, of course, not necessary to run these tests using vector control. An open-loop PWM scheme that delivers constant volts per hertz suffices. Again, there is some phase error at lower fluxes, due mostly to noise on the voltage signal.

The motor parameters which best fit the data collected are:

1. $L_m = (0.22 - 0.2\Psi) H$
2. $L_{sl} = L_{rl} = 0H$

Figure 2.15: Simplified Motor Model

3. $R_s = 1\Omega$
4. $R_c = 300\Omega$
5. $R_r = 0.9$ to 1.1Ω depending on rotor temperature

The simplified model with the measured parameters is shown in figure 2.15. To validate this model's accuracy, it was used to predict the line to neutral current while the motor ran with $\omega_r = 1000rpm$, $T = 3Nm$ and $\Psi = 0.4Wb$. This calculated voltage was put out the d/a and then compared to the actual line-to-neutral stator voltage. The results for this test are shown in figure 2.16. A second test was performed with $\omega_r = 2000rpm$, $T = 3Nm$ and $\Psi = 0.4Wb$, and the results are displayed in figure 2.17.

Figure 2.16: Model Prediction vs Actual Line to Neutral Voltage (1000rpm)

Figure 2.17: Model Prediction vs Actual Line to Neutral Voltage (2000rpm)

CHAPTER 3

Vector Control

3.1 A Physical Explanation of Vector Control

The objective of vector control is to simplify the induction motor control equations so that they resemble the equations for a separately excited DC motor. As shown in equation 2.6 and explained in section 2.2.1, torque in a separately excited DC machine is linearly proportional to armature current. In addition, the DC machine's flux is also a simple linear relationship established by the field current. In order to double the torque, the controller has simply to double the armature current. Since both motors operate on the same fundamental principles, it seems reasonable to assume that there should be a simple set of control equations that are analogous to the DC machine equations.

As explained in section 2.2.1, an induction motor's torque is determined by the cross product of rotor flux and rotor current. The torque has a linear relationship with rotor current. In addition, flux is linearly proportional to the current that causes it. The difficulty lies in that there is no direct control of either torque-producing quantity - rotor flux or rotor current - though it has been shown that they exist. Vector control provides a means of controlling the rotor flux and the

rotor current by performing the necessary conversions. Once this is done, control becomes as simple and straightforward as in a DC machine.

3.2 Vector Control Method Selection

The selection of a method of vector control is described in this section using terminology not yet clearly defined in this document. It is, however, necessary to establish which path would be followed in the derivation of the control law. Terminology and concepts referred to in this section are explained in section 3.3.

There are many ways to implement vector control of an induction motor. Beginning with the fundamental governing equations, the derivation path which leads to the final control law can take several turns. The chosen control variables determine which motor variables need be observed. There are essentially three values that can be easily observed on an induction motor: motor speed, voltage and current. While it is properly the magnetic flux that should be observed, there is no practical or robust method to directly measure flux. It is therefore necessary to infer the magnetic flux from the easily measured values of voltage and current. Five primary methods of performing vector control were evaluated for this project. They are:

3.2.1 Open-Loop Current-Mode Flux Observer This method uses the command current I_{sd} and I_{sq} to drive the observer model to arrive at a value for flux and torque. It assumes the ideal operation of the current amplifier. As shown

in section 4, the performance of the current amplifier is less than perfect. In addition, no compensation is built into the observer to adapt to variations in motor parameters or non-linearities. This method is applicable for non-demanding control applications where a high level of performance is not required.

3.2.2 Voltage-Mode Flux Observer During the development of this project, a voltage-mode observer was tested. It was found to operate reasonably well but was not as robust as the current-mode observer. It is believed that this was due to the inherent difficulty in measuring the voltage signals. Since the voltage to the induction motor is a PWM waveform, it is necessary to filter to arrive at a sinusoidal signal which accurately represents the integral of the PWM waveform. The more filtering that is applied, the better the signal looks and the more phase shift is introduced. Given a fundamental frequency in excess of $500Hz$ and a switching frequency of $8kHz$, barely over a decade apart, it is clear that filtering this signal will involve trade-offs. The real benefit to using the voltage mode observer is that it is not dependent on position feedback or rotor resistance. The weakness that eliminated it from use (beyond initial testing) in this project was its inability to operate at slow speeds.

3.2.3 Current-Mode Flux Observer Current measurement and rotor position feedback provided a reasonably accurate flux observer, though errors of approximately 50% were observed. Despite this inaccuracy, it was possible to achieve reasonably good velocity and position control with the induction motor. Torque control was

only as accurate as this model allowed.

3.2.4 Current-Mode Flux Observer with Feedback Loops on I_{sd} and I_{sq} and

Control of PWM This method of observer was not implemented due to the problems encountered in having the DSP perform the task of current amplifier. Noise, inherent in all a/d measurements, becomes significant when the difference between two noisy signals is used. With this method, it is necessary to modify the PWM command directly based on the variations in I_{sd} and I_{sq} . As shown in figure 5.1, I_{sd} and I_{sq} are calculated using ρ which is also an observed and calculated quantity. This would tie PWM operation to the observer loop speed. Given the noise, direct PWM control and calculation requirements for this mode of operation, it was decided that this method would not provide a simple and robust observer.

3.2.5 Current-Mode Flux Observer with Feedback Loops on I_{sd} and I_{sq} The

measured currents were used to drive the observer and were compared with their commanded values. The command currents were modified to compensate for any differences. It was observed that the model accuracy improved significantly with the addition of these two control loops. The accuracy of the observed torque was verified by use of a torque sensor.

This method of vector control, chosen for this project, operates in the rotor frame of reference and employs: a current-driven PWM amplifier; a measured current-mode observer with position feedback; and feedback loops on I_{sd} and I_{sq} .

3.3 Formal Derivation of the Vector Control Equations

3.3.1 Definitions

Definition 1 *From Faraday's law, the voltage across an inductor can be expressed as $V_L = \frac{d\Psi}{dt}$. Inductance is defined as $L \equiv \frac{\delta\Psi}{\delta i}$ and speedance is defined as $K_g \equiv \frac{\delta\Psi}{\delta x}$. Since flux linkage is generally a function of current and position, V_L can be expanded by the chain rule to be $V_L = \frac{\delta\Psi}{\delta i} \frac{di}{dt} + \frac{\delta\Psi}{\delta x} \frac{dx}{dt}$. Combining, $V_L = L \frac{di}{dt} + K_g \frac{dx}{dt}$ which when no moving parts are included becomes:*

$$V_L = L \frac{di}{dt} \quad (3.1)$$

[Chai, 2.23]

Definition 2 *Rotor flux Ψ_r is defined as the flux seen by the rotor. It is the vector summation of the flux caused by the stator and rotor currents. The rotor flux is almost the same as the stator flux and only differs by the amount of rotor flux leakage.*

Definition 3 *Stator flux Ψ_s is defined as the flux seen by the stator. It is the vector summation of the flux caused by the stator and rotor currents. The stator flux is almost the same as the rotor flux and only differs by the amount of the stator leakage flux.*

Given the above definitions, rotor flux is defined as:

$$\Psi_r = L_m I_s + L_r I_r \quad (3.2)$$

where:

$$L_r = L_m + L_{r\ell} \quad (3.3)$$

and $L_{r\ell}$ is the rotor leakage inductance.

In the same manner,

$$\Psi_s = L_m I_r + L_s I_s \quad (3.4)$$

where:

$$L_s = L_m + L_{s\ell} \quad (3.5)$$

and $L_{s\ell}$ is the stator leakage inductance.

Rotor flux and stator flux can be rewritten as:

$$\Psi_s = L_m (I_r + I_s) + L_{s\ell} I_s \quad (3.6)$$

$$\Psi_r = L_m (I_r + I_s) + L_{r\ell} I_r \quad (3.7)$$

which are valid in any frame of reference [Hsu].

3.3.2 Reference Frames and Rotational Matrices Establishing the frame of reference is essential for each expression. The stator frame of reference is identified by a superscript s and the rotor frame of reference is identified by a superscript r . For example, I_{sx}^r is the stator current in the x-direction expressed in the rotor reference frame. The angle between the rotor frame of reference and the stator frame of reference is θ as shown in figure 3.1. The rotor flux frame of reference is shown in figure 3.2. It is defined to be a frame of reference which has its x-axis

Figure 3.1: Stator and Rotor Reference Frames

aligned with the rotor flux. By definition, there can be no y-component of rotor flux.

Figure 3.2: Stator vs Rotor Flux Frame

A rotational matrix, used to convert between frames of reference, is defined as:

$$R(\theta) = \begin{bmatrix} \cos \theta & -\sin \theta \\ \sin \theta & \cos \theta \end{bmatrix} \quad (3.8)$$

Some properties of this rotational matrix are defined as:

$$R(\theta)^{-1} = R(-\theta) \quad (3.9)$$

$$R(-\theta)\dot{R}(\theta) = \begin{bmatrix} 0 & -\dot{\theta} \\ \dot{\theta} & 0 \end{bmatrix} \equiv S(\dot{\theta}) \quad (3.10)$$

$$R(\theta)\dot{R}(-\theta)R(-\theta) = S(\dot{\theta}) \quad (3.11)$$

$$R(\theta)S(\dot{\theta})R(-\theta) = S(\dot{\theta}) \quad (3.12)$$

This rotational matrix and the given properties are used to convert different quantities between frames of reference in the following section.

The purpose of the stator windings is to produce a rotating magnetic flux field. This can be accomplished by use of three wye-connected phase windings as described in section 2.1.1, or by use of two orthogonal windings. Since the resultant vector field is the same whether a two or three-phase source is used, it greatly simplifies the analysis to use a two-phase stator for analysis. The stator currents will be referred to as:

$$I_s^s = \begin{bmatrix} I_{sx}^s \\ I_{sy}^s \end{bmatrix} \quad (3.13)$$

This is identical to equation 2.1 and is repeated here for convenience, with the addition of superscripts to designate the frame of reference. The same format will be used to express the stator voltage [Hsu].

3.3.3 Rotor and Stator Voltage Equations The stator voltage equation is:

$$V_s^s = R_s I_s^s + \frac{d}{dt} \Psi_s^s \quad (3.14)$$

or, restated in matrix notation:

$$\begin{bmatrix} V_{sx}^s \\ V_{sy}^s \end{bmatrix} = R_s \begin{bmatrix} I_{sx}^s \\ I_{sy}^s \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \Psi_{sx}^s \\ \Psi_{sy}^s \end{bmatrix} \quad (3.15)$$

In a similar manner, the rotor voltage equation, which is the voltage measured across the shorted rotor bars, is:

$$0 = R_r I_r^r + \frac{d}{dt} \Psi_r^r \quad (3.16)$$

or, restated in matrix notation:

$$\begin{bmatrix} 0 \\ 0 \end{bmatrix} = R_r \begin{bmatrix} I_{rx}^r \\ I_{ry}^r \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \Psi_{rx}^r \\ \Psi_{ry}^r \end{bmatrix} \quad (3.17)$$

Since $\Psi_r^r = R(-\theta)\Psi_r^s$, equation 3.16 can be rewritten as $0 = R_r I_r^r + \frac{d}{dt} (R(-\theta)\Psi_r^s)$ which is still expressed in the rotor reference frame. Differentiating, the rotor voltage equation becomes $0 = R_r I_r^r + \dot{R}(-\theta)\Psi_r^s + R(-\theta)\dot{\Psi}_r^s$. Applying the rotational matrix $R(\theta)$ for conversion to the stator reference frame, the rotor voltage equation becomes $0 = R(\theta)R_r I_r^r + R(\theta)\dot{R}(-\theta)\Psi_r^s + R(\theta)R(-\theta)\dot{\Psi}_r^s$ which is now expressed in the stator reference frame. Using the properties outlined in equations 3.8 through 3.12 to simplify, the rotor voltage equation expressed in the stator reference frame is:

$$0 = R_r I_r^s - S(\dot{\theta})\Psi_r^s + \dot{\Psi}_r^s \quad (3.18)$$

The rotor voltage equation expressed in the stator reference frame in matrix

form is:

$$\begin{bmatrix} 0 \\ 0 \end{bmatrix} = R_r \begin{bmatrix} I_{rx}^s \\ I_{ry}^s \end{bmatrix} - \begin{bmatrix} 0 & -\dot{\theta} \\ \dot{\theta} & 0 \end{bmatrix} \begin{bmatrix} \Psi_{rx}^s \\ \Psi_{ry}^s \end{bmatrix} + \begin{bmatrix} \dot{\Psi}_{rx}^s \\ \dot{\Psi}_{ry}^s \end{bmatrix} \quad (3.19)$$

Since $\Psi_r^s = R(-\rho)\Psi_r^{r\Psi}$, the rotor voltage equation can be rewritten as $0 = R_r I_r^s - S(\dot{\theta})\Psi_r^s + \frac{d}{dt}(R(-\rho)\Psi_r^{r\Psi})$ which is still expressed in the stator frame. Simplifying, the stator voltage equation becomes $0 = R_r R(\rho)I_r^s - S(\dot{\theta})R(\rho)\Psi_r^s + R(\rho)\dot{R}(-\rho)\Psi_r^{r\Psi} + R(\rho)R(-\rho)\dot{\Psi}_r^{r\Psi}$. Using the properties outlined in equations 3.8 through 3.12, the rotor voltage equation expressed in the rotor flux frame becomes:

$$0 = R_r I_r^{r\Psi} + S(\dot{\rho} - \dot{\theta})\Psi_r^{r\Psi} + \dot{\Psi}_r^{r\Psi} \quad (3.20)$$

Restated in matrix format, the rotor voltage equation, expressed in the rotor flux frame is:

$$\begin{bmatrix} 0 \\ 0 \end{bmatrix} = R_r \begin{bmatrix} I_{rx}^{r\Psi} \\ I_{ry}^{r\Psi} \end{bmatrix} + \begin{bmatrix} 0 & \dot{\theta} - \dot{\rho} \\ \dot{\rho} - \dot{\theta} & 0 \end{bmatrix} \begin{bmatrix} \Psi_{rx}^{r\Psi} \\ \Psi_{ry}^{r\Psi} \end{bmatrix} + \begin{bmatrix} \dot{\Psi}_{rx}^{r\Psi} \\ \dot{\Psi}_{ry}^{r\Psi} \end{bmatrix} \quad (3.21)$$

There is no y-component of rotor flux in the rotor flux frame, as the rotor flux frame is defined as having the x-axis coincident with the rotor flux. Therefore $\Psi_{ry}^{r\Psi} = 0$ and equation 3.21 can be simplified. Since there is only an x-component of rotor flux, it can be expressed as $|\Psi_r|$ and it is understood that it is in the x-direction. Equation 3.21 can be restated as two simple equations. The rotor voltage equations expressed in the rotor flux frame are:

$$0 = R_r I_{rx}^{r\Psi} + |\dot{\Psi}_r| \quad (3.22)$$

$$0 = R_r I_{ry}^{r\Psi} + (\dot{\rho} - \dot{\theta})|\Psi_r| \quad (3.23)$$

Equation 3.7, stated in the rotor flux frame and solved for I_r^Ψ , becomes:

$$\begin{bmatrix} I_{rx}^\Psi \\ I_{ry}^\Psi \end{bmatrix} = \frac{1}{L_r} \begin{bmatrix} \Psi_{rx}^\Psi \\ 0 \end{bmatrix} - \frac{L_m}{L_r} \begin{bmatrix} I_{sx}^\Psi \\ I_{sy}^\Psi \end{bmatrix} \quad (3.24)$$

Substituting I_{rx}^Ψ and I_{ry}^Ψ from equation 3.24 into equations 3.22 and 3.23 respectively, the rotor voltage equations become:

$$0 = \frac{R_r}{L_r} |\Psi_r| - \frac{R_r L_m}{L_r} \dot{I}_{sx}^\Psi + |\dot{\Psi}_r| \quad (3.25)$$

$$0 = -\frac{R_r L_m}{L_r} I_{sy}^\Psi + (\dot{\rho} - \dot{\theta}) |\Psi_r| \quad (3.26)$$

which is the form used in the vector control block diagram shown in figure 5.1 [Hsu].

3.3.4 Steady-State Operation Under steady state operation, an equivalent circuit can be developed. Using equations 3.1, 3.2, 3.3, 3.4 and 3.5, the stator and rotor voltage equations (3.14 and 3.16) can be restated as:

$$V_s^s = R_s I_s^s + L_s \dot{I}_s^s + L_m \dot{I}_r^s \quad (3.27)$$

$$V_r^r = R_r I_r^r + L_r \dot{I}_r^r + L_m \dot{I}_s^r \quad (3.28)$$

Using the substitution:

$$\dot{I}_r^r = \frac{d}{dt} (R(-\theta) I_r^s) = \dot{R}(-\theta) I_r^s + R(-\theta) \dot{I}_r^s \quad (3.29)$$

$$\dot{I}_s^r = \frac{d}{dt} (R(-\theta) I_s^s) = \dot{R}(-\theta) I_s^s + R(-\theta) \dot{I}_s^s \quad (3.30)$$

and by setting $V_r^r = 0$, equation 3.28 becomes:

$$0 = R_r I_r^r + L_r (\dot{R}(-\theta) I_r^s + R(-\theta) \dot{I}_r^s) + L_m (\dot{R}(-\theta) I_s^s + R(-\theta) \dot{I}_s^s) \quad (3.31)$$

which is still expressed in the rotor reference frame. By applying the rotational matrix $R(\theta)$ and simplifying, equation 3.28 becomes:

$$0 = R_r I_r^s - L_r S(\dot{\theta}) I_r^s + L_r \dot{I}_r^s - L_m S(\dot{\theta}) I_s^s + L_m \dot{I}_s^s \quad (3.32)$$

Using equation 3.10, and $\dot{\theta} = \omega_r$, equations 3.14 and 3.16, written in matrix form become:

$$\begin{bmatrix} V_{sd} \\ V_{sq} \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} R_s & 0 & 0 & 0 \\ 0 & R_s & 0 & 0 \\ 0 & L_m \omega_r & R_r & L_r \omega_r \\ -L_m \omega_r & 0 & -L_r \omega_r & R_r \end{bmatrix} \begin{bmatrix} I_{sd}^s \\ I_{sq}^s \\ I_{rd}^s \\ I_{rq}^s \end{bmatrix} + \begin{bmatrix} L_s & 0 & L_m & 0 \\ 0 & L_s & 0 & L_m \\ L_m & 0 & L_r & 0 \\ 0 & L_m & 0 & L_r \end{bmatrix} \begin{bmatrix} \dot{I}_{sd} \\ \dot{I}_{sq} \\ \dot{I}_{rd}^s \\ \dot{I}_{rq}^s \end{bmatrix} \quad (3.33)$$

Since rotor speed ω_r is not constant, standard linear circuit analysis cannot be used to solve equation 3.33. Note: ω_e is the applied electrical frequency. If steady-state operation ($\omega_e = \text{constant}$) and applied inputs are sinusoidal, then a solution can be found using the Fourier transform:

$$\mathfrak{F}(X(t)) = X(\omega) \quad (3.34)$$

$$\mathfrak{F}(\dot{X}(t)) = j\omega X(\omega) \quad (3.35)$$

Consider the d -components of equation 3.33 which when restated in the frequency domain become:

$$V_{sd} = R_s I_{sd}(\omega_e) + j\omega_e L_s I_{sd}^s(\omega_e) + j\omega_e L_m I_{rd}^s(\omega_e) \quad (3.36)$$

$$0 = L_m \omega_r I_{sq}(\omega_e) + R_r I_{rd}^s(\omega_e) + L_r \omega_r I_{rq}^s + j\omega_e L_m I_{sd} + j\omega_e L_r I_{rd}^s \quad (3.37)$$

Since:

$$I_{sq}^s(\omega_e) = -jI_{sd}^s(\omega_e)$$

$$I_{rq}^s(\omega_e) = -jI_{rd}^s(\omega_e)$$

equations 3.36 and 3.37 can be simplified:

$$V_{sd} = (R_s + j\omega_e L_s) I_{sd} + j\omega_e L_m I_{rd}^s \quad (3.38)$$

$$0 = jL_m \omega_e \left(\frac{\omega_e - \omega_r}{\omega_e} \right) I_{sd} + \left(R_r + jL_r \omega_e \left(\frac{\omega_e - \omega_r}{\omega_e} \right) \right) I_{rd}^s \quad (3.39)$$

By defining slip:

$$s \equiv \frac{\omega_e - \omega_r}{\omega_e} \quad (3.40)$$

equation 3.39 becomes:

$$0 = jL_m I_{sd} + \left(\frac{R_r}{s} + jL_r \omega_e \right) I_{rd}^s \quad (3.41)$$

By using equations 3.38 and 3.41, the standard steady-state model for the induction motor is defined and is the same as derived in section 2.3 (assuming $R_c = 0$). A similar derivation can be performed with analogous results starting with the q components of equation 3.33.

3.3.5 Torque Equations As discussed in section 2.2.1, torque can be expressed in a number of equivalent ways. Two expressions of torque are:

$$T = -\Psi_r^{r\Psi} \times I_r^{r\Psi} \quad (3.42)$$

$$T = \Psi_s^{r\Psi} \times I_s^{r\Psi} \quad (3.43)$$

which are valid in any frame of reference. Note in equation 3.42 that since torque is the cross product of rotor flux and rotor current, there is no torque if they are in the same direction. Maximum torque is achieved when the rotor flux and rotor current are orthogonal. Unfortunately, there is no direct control of I_r . By rearranging and solving for I_r in equation 3.6, the torque from equation 3.42 becomes:

$$T = -\Psi_r^{r\Psi} \times \left[\frac{1}{L_r} (\Psi_r^{r\Psi} - L_m I_s^{r\Psi}) \right] \quad (3.44)$$

which when simplified gives:

$$T = \frac{L_m}{L_r} \Psi_r^{r\Psi} \times I_s^{r\Psi} \quad (3.45)$$

$$T = \frac{L_m}{L_r} \begin{bmatrix} |\Psi_r| \\ 0 \end{bmatrix} \times \begin{bmatrix} I_{sx}^{r\Psi} \\ I_{sy}^{r\Psi} \end{bmatrix} \quad (3.46)$$

$$T = \frac{L_m}{L_r} |\Psi_r| \cdot I_{sx}^{r\Psi} \quad (3.47)$$

It is important to remember that the definition of rotor flux $|\Psi_r|$ derives from the vector summation of the flux due to the rotor current and the flux due to the stator current. Finally, equation 3.47, for an induction motor, is starting to resemble the torque equation for a separately excited DC motor (equation 2.6) as discussed in section 2.2.1. Torque is now expressed in terms of flux and current. Restating the rotor flux equation (equation 3.7) in the rotor flux frame:

$$|\Psi_r| = L_m (I_{rx}^{r\Psi} + I_{sx}^{r\Psi}) + L_{r\ell} I_{rx}^{r\Psi} \quad (3.48)$$

By inserting $|\Psi_r|$ into equation 3.46, torque becomes:

$$T = L_m I_{rx}^{r\Psi} I_{sx}^{r\Psi} \quad (3.49)$$

Restated in the vernacular of related literature,

$$T = L_m I_{sd} I_{sq} \quad (3.50)$$

[Hsu]

CHAPTER 4

Hardware Description

This chapter briefly describes the hardware used to achieve vector control of an induction motor. Performance results demonstrating the operation of specific pieces of hardware but not pertaining directly to system performance have been included here. System performance results are shown in chapter 6.

The system used to develop and test the vector-control algorithms is comprised of seven main components:

1. Host computer
2. DSP board
3. DC bus and power electronics
4. Interface board and transducers (which includes the current amplifier and analog input circuits)
5. Induction motor and encoder
6. DC motors (2) and torque transducer
7. Load box

These items are depicted in figure 4.1.

Figure 4.1: Hardware System Drawing

4.0.6 Host Computer The host computer is an IBM-compatible 486 personal computer. It also contains the DSP board. Its primary purpose is to provide a platform for writing, compiling, linking and downloading the control code to the DSP board. It also provides real-time control of code variables through semaphoring - the shared use of common RAM to transfer information - and provides a means of monitoring the actions of selected control variables during code operation.

4.0.7 DSP Board The DSP board contains the Texas Instruments TMS320C30 floating-point microprocessor. The DSP development board was manufactured by Sonitech International Inc. It resides in an expansion slot inside the PC and comes

Figure 4.2: DC Bus and IGBTs

equipped with 16 kB of RAM. The DSP board communicates with the interface board through the external I/O bus. The external I/O bus is comprised of data and address lines as well as relevant I/O control lines.

4.0.8 DC Bus and Power Electronics The low inductance of the DC bus is achieved by use of two flat conductors separated by an epoxy-fiberglass insulator. The four capacitors and three IGBT power modules are bolted directly to the bus as shown in figure 4.1. A 208-volt, 3-phase variac in conjunction with a bridge rectifier supplies DC to the bus. The IGBTs are manufactured by Powerex Inc. in pairs and contain related drive circuitry. The driver boards receive $\pm 15V$ which supply the DC-to-DC converters that in turn supply each IGBT with an isolated power supply. The PWM drive signals are isolated through use an opto-isolator

Figure 4.3: Phase-to-Phase Voltage Rise Time

Figure 4.4: Switching Frequency at 20kHz

Figure 4.5: PWM Voltage Wave Form (fundamental = 185Hz)

with high $\frac{dv}{dt}$ immunity.

Figure 4.3 shows one phase-to-phase voltage transition. It is interesting to note the overshoot and high rate of change achievable with IGBTs. Figure 4.4 shows the switching waveform at 20kHz. Small transients riding on the square wave indicate where another switching action has occurred on another phase. Notice that these transients occur exactly 5μsec from the vertical edges of the square wave, which correlates perfectly with the programmed dead time of 5μsec. Figure 4.5 shows the PWM voltage waveform that creates the sinusoidal currents. In this case, the fundamental frequency is approximately 185Hz.

4.0.9 Interface Board and Transducers (Xdcers) The interface board is a prototype, wire-wrapped circuit board which has plated-through holes and a single-sided ground plane. All integrated circuits are mounted to wire-wrap sockets. Individual

Figure 4.6: Current and Voltage Transducers

components (such as resistors and capacitors) are soldered on component carriers and inserted into wire-wrap sockets. Normal wire-wrap wire (#30AWG) connects all signals and power connections to components. Heavier wire (#24AWG) is used to run the power supply voltages to different regions of the board. Noise-sensitive circuits have their ground pins soldered to the ground plane to achieve better noise immunity. Figure 4.3 shows one voltage-switching transition. It is interesting to note that $\frac{dv}{dt} \approx 3500 \frac{\text{volt}}{\mu \text{sec}}$ which is a good indication of where the noise is coming from.

There are three current and three voltage transducers mounted on a separate board. These transducers are all manufactured by LEM Corporation. The bandwidth on these transducers is approximately 200kHz . They provide the following

two signals to the DSP:

1. Voltages: $(V_{sA} - V_{sB})$, $(V_{sB} - V_{sC})$ and the DC bus where the scale is $500volts \Rightarrow 5volts$
2. Currents: I_{sA} , I_{sB} , I_{sC} and where the scale is $30amps \Rightarrow 4volts$

Current Amplifier The current-amplifier control circuitry is located on the I/O board. It uses a voltage reference signal from the d/a that represents the desired current, compares it to the actual current and generates an error signal. This error signal is compared to a triangle wave, and the resultant comparator output is the PWM signal. This PWM signal is sent to the IGBT driver circuits via differential line drivers for noise immunity. Such a proportional feedback design requires that there be a steady-state error between the reference and the actual current signals. The performance of the current amplifier (and power electronics) with the motor operating under moderate load at $1800rpm$ is shown in figure 4.7. Both the reference and the actual current waveforms are perfectly sinusoidal and very little difference occurs between the two signals.

Figure 4.8 graphs the current amplifier performance with the motor free of load and running at $7000rpm$ ($230Hz$), with a reduced magnetizing current. The performance of the current amplifier has deteriorated significantly. The phase shift and drop in amplitude is caused by a number of factors. The two most significant contributors are:

Figure 4.7: Current: Reference vs Actual

Figure 4.8: Current Amplifier Performance Reference vs Actual

1. The motor is lightly loaded, making it more of an inductive load (R'_r from figure 2.12 goes to ∞).
2. There is no bus voltage available to force the current to match the reference.

As the motor speed is increased, field weakening is applied to reduce the back emf, but there is a limit to the amount of field weakening that can be applied under control. When running at over four times the rated motor speed, the back emf has increased to a level almost equal to the bus voltage so that overmodulation occurs and the PWM ceases switching at the peak of the sine wave.

One test evaluated the performance of the current amplifier run through the DSP with a bus-write PWM instead of the aforementioned analog circuitry. The measured current signal was compared to the commanded current signal and the PWM was adjusted to correct for error. While this method worked, it was found too be sensitive to an elevated proportional gain. The resultant current waveform was also found to be noisy. It was concluded that there were three major reasons why the performance of the direct bus-write PWM was inferior to that of the analog version:

1. There was a $125\mu\text{sec}$ delay introduced by the loop speed of the controller.
2. There is noise associated with a/d conversions in this environment. While the noise was found to be almost insignificant compared to the current signal,

it comprised a significant portion of the difference between the actual and commanded currents. It is this difference signal that was used to generate the bus-write PWM signal.

3. The analog circuitry has inherent signal averaging. If it were possible to sample at a faster rate and average the measured a/d signal, the signal-to-noise ratio could be improved along with the performance of the direct bus-write PWM.

Analog Input Circuits The analog input circuits are located on the interface board which is set up to handle $\pm 5V$ analog signals. The current input signals from the LEM current transducers operate at $\pm 4V$. Analog offsets can be nulled in software at both the a/d and d/a stages. Failure to accurately null the analog offsets results in substantial torque ripple.

The interface board is made up of custom-built circuitry which handles the I/O for the DSP board. Communications with the DSP board are via a short ribbon-cable bus. The current and voltage transducers are located on two separate circuit boards which connect to the interface board via analog cables.

4.0.10 The Induction Motor and Encoder The induction motor is a two-horsepower, three-phase, 208/460-volt AC induction motor manufactured by Baldor. Connected for 208-volt operation, the motor has the following operating specifications:

- Model number: M3587T

Figure 4.9: Induction Motor and Encoder

- Type: TEFC converted to TEBC (shaft-mounted cooling fan removed)
- Full-load torque: $8.135Nm$ ($6ft \cdot lbf$)
- Full-load amps: $6.2amps\ rms$
- Locked-rotor torque: $26.303Nm$ ($19.4ft \cdot lbf$)
- Locked-rotor amps: $21amps\ rms$
- Breakdown torque: $31.184Nm$ ($23ft \cdot lbf$)
- Efficiency: 82.5%
- Motor parameters: $R_s = 1.11$, $R_r = 0.8625$, $X_{sl} = 1.2475$, $X_{rl} = 1.0525$,
 $X_m = 36.95$
- Specs: mass = $25.4kg$, speed = $1725rpm$, poles = 4.

Figure 4.10: Two DC Motors

The encoder is an optical encoder manufactured by Heidenhain. It has 2500 counts per revolution and a maximum speed of operation of 7000rpm.

4.0.11 Torque Sensor and DC motors The torque sensor is an in-line torque sensor, model number 1104-100, manufactured by Lebow Industries Ltd. It has a full-scale range of 10 newton-meters and delivers 1mV per newton-meter. This signal is amplified and scaled to yield 1V/Nm which is appropriate for reading by the oscilloscope.

The DC motors were manufactured by GE approximately half a century ago and there is no information available on them. They are separately excited DC motors with a nameplate rating of 3/4 horsepower. They are driven by a variable DC supply when testing the induction motor as a generator and are also used as generators to apply a load to the induction motor during motor controls testing.

During use as generators, a load box is connected to the output of the armatures to dissipate the generated power.

4.0.12 The Load Box The load box (seen in the background of figure 4.2) is made up of 10 80Ω , $110W$ power resistors and has two cooling fans to dissipate heat. Jumpers on the top reconfigure the load box for the various loads required.

CHAPTER 5

Control Code

5.1 Vector Control Block Diagram

The vector control block diagram (figure 5.1) depicts the derived model for an induction motor and the vector control equations. The software (written in C) represented by this block diagram contains approximately 1200 lines of code with comments. Since the TMS320C30 is a floating-point DSP and is very quick at performing calculations, no look-up tables were used to create the required sinusoids. All mathematical computations were performed as shown. In the instances where division was performed, the denominator was prevented from becoming too small, preventing division by zero. Since all signals used in the computations were subject to noise and offsets, many had to be filtered. In addition, pure integration was replaced by a high-pass filter to avoid integrating noise and offsets. This technique yields acceptable results and avoids the problems associated with pure integration.

The prototype board hosts eight d/a channels. Four are used to run this controller. The other four are used extensively for debugging and optimizing the code. Any one of the software variables shown in figure 5.1 could be selected for observation in real time on an oscilloscope. This window into the code proved to

be invaluable while evaluating both the prototype hardware and new software.

The variables listed in figure 5.1 are the variable names that are used in the software in the software observer and Greek symbols in the sections which represent the induction motor and its equations. The symbols used in the governing equations are used in the motor model section. Every attempt was made to choose meaningful names to make the code and block diagrams easier to understand. Since the software is written with an ASCII editor, no Greek symbols can be used. See section 1.2 for the definitions of the symbols used in this document and their software equivalents.

5.2 Flow Charts

The following figures depict the basic structure of the software code. The main control-loop speed was set at $8kHz$. The software is not listed as it is deemed to be proprietary by the sponsor. After initialization has been run, the control code enters the control loop. The loop-speed timer is set and the various subroutines are called in the order shown.

Figure 5.1: Vector Control Block Diagram

Flow Chart: Main

Flow Chart: Initialize

Flow Chart: Set Parameters

Flow Chart: Speed

Flow Chart: Get Analogs

Flow Chart: Mode

Flow Chart: Is Command Out

Flow Chart: Motor Power Calculation

5.3 Control

The vector control code operates as if it were a black-box controller, with an input of torque command and an output of motor torque, which can be used to achieve position, velocity or torque control of the induction motor. The type of controller to be used is determined by the application. The following are a few examples of the controllers that were implemented with the vector control code.

5.4 Position Controller

Position control is achieved using the classic PID controller shown in figure 6.1. This PID controller uses the vector control block with input of torque command and output of motor torque to achieve the desired position. The terms for proportional gain, integral gain and derivative (velocity) gain were derived experimentally while performing the weight-lifting tests described in section 6.1.

Figure 5.2: PID Position Controller Block Diagram

5.5 Speed Controller

The speed controller is a classic proportional-integral (PI) controller. The speed command received by the controller may be a step command but the PI controller's response to a large step input is a very large torque command. This large torque command results in very large currents which inflict a lot of stress on many of the mechanical and electrical components in the system. In addition, the intended partner for this controller is an engine-drive induction generator to generate the DC bus. In this application, the engine is unable to respond to a step command for power. To alleviate these problems, the speed command is filtered. For small step changes, the command signal is put through a low-pass filter. For large step changes, the $\frac{d}{dt}(\text{SpeedCmd})$ is limited to a maximum ramp level. When the actual speed approaches the command speed, the ramp limit cuts out and the filter cuts in. It is this filtered speed command that the PI controller works with,

assuring stable and smooth operation.

Figure 5.3: Speed Command Filter

Figure 5.4: PI Speed Controller

5.6 Generation Control

Figure 5.5 shows the configuration of the generation-mode controller. This controller's output is divided by the shaft speed ω_r to allow for a proportional gain term which will demand the same magnitude of power output change no matter what voltage it is controlling. This was done since torque is the output from the controller and power is proportional to the product of shaft speed and torque. This ensures that a step change in voltage command would place the same delta-

Figure 5.5: PI Generation Control

demand for power on the DC motors as it would for the same step change at a different speed of operation. This controller demonstrates the ability to perform voltage control, but the final control algorithm will need to be defined in terms of the power that can be produced by the engine drive for each speed of rotation.

This controller needs its own flux-optimization control algorithm (FOCA) to run the induction generator at optimum efficiency, and in addition, it must be able to quickly set the flux to zero for a lost-load condition. The controller also must take into account the back emf and perform flux weakening if the commanded voltage is below that value. These tasks are beyond the scope of this project.

CHAPTER 6

Motor Performance

6.1 Position Control

Position control is tested with a mass attached by cable to a pulley as shown in figure 6.1. The radius of the pulley is $106mm$ ($4.154''$). Near zero speed, the actual position of the rotor becomes essential for precise position control. A high-resolution encoder with 2500 lines (10,000 counts) per revolution ensures precise position feedback. A classic PD controller is applied to position control. A trapezoidal position routine was chosen to demonstrate the induction motor's position control. The mass used in figure 6.1 was $11.34kg$ ($25lbf$). The torque generated to hold $11.34kg$ was $11.73Nm$ ($8.65ft \cdot lbf$). This torque level is 1.44 times the rated torque of $8.135Nm$ ($6ft \cdot lbf$) for this motor yet position control is achieved in a precise and robust manner. The results of this test are shown in figure 6.2. Please note that the reference signal is offset by one graticule as it is not otherwise possible to distinguish between the two traces. Notice that there is no softening of the corners in response to the reference signal despite the implied infinite torque that would be required to go from zero to a finite speed in zero time. Despite this demanding reference signal, any deterioration in the position control response is

Figure 6.1: Position Control Test Setup

lost in the noise of the digital storage oscilloscope.

The trapezoidal position test was repeated with a mass of 22.68kg (50lb). The torque generated to hold this mass was 23.47Nm . The results of this test are shown in figure 6.3. Again, the reference signal is offset one graticule. The following ratios pertaining to this test are worth noting:

$$\frac{\text{StationaryTorque}}{\text{RunningTorque}} = 2.9$$

$$\frac{\text{StationaryTorque}}{\text{LockedRotorTorque}} = 0.89$$

$$\frac{\text{StationaryTorque}}{\text{BreakDownTorque}} = 0.75$$

Figure 6.2: Graph of Position Control: Reference vs Actual with 25lbf

Figure 6.3: Graph of Position Control: Reference vs Actual with 50lbf

With this greater load, too, it is difficult to discern any significant degradation in the position control of the motor. There is, however, a slight overshoot and undershoot at the corners of the trapezoidal position profile, as would be expected. In addition, there is some evidence of torque ripple which shows up as fluctuations in the actual position signal at this higher torque level. This can be most clearly seen while the mass is being raised or lowered. By way of interest, the trapezoidal routine demands that the pulley turn exactly π radians in three seconds. This translates into a linear velocity of $0.11 \frac{m}{sec}$.

One further test was performed for position control. A mass of $36.3kg$ ($80lbf$) was raised approximately $40cm$, held for one minute, then lowered. The torque generated to hold $36.3kg$ was $37.55Nm$ ($27.7ft \cdot lbf$). As shown by the following ratios, this level of torque is 20% higher than claimed by the manufacturer for any range of operation.

$$\frac{StationaryTorque}{RunningTorque} = 4.6$$

$$\frac{StationaryTorque}{LockedRotorTorque} = 1.4$$

$$\frac{StationaryTorque}{BreakDownTorque} = 1.2$$

6.2 Torque Control

The torque response was tested in a manner which most closely resembled its intended use - responding to a torque demand while running at varied angular velocities. The test setup is shown in figure 4.1. For this particular test, the DC

motors were connected as generators and the load box was used to dissipate their generated power. While running in the torque-control mode of operation with a steady-state torque command of $1Nm$, a step command to $6Nm$ was entered. The response is shown in figure 6.4. The rise time is approximately $150msec$. The intended application has a diesel engine to drive an induction generator which powers the DC bus. It is anticipated that this controller's energy demand rate of change would exceed the rate at which the diesel engine could respond so no further attempts at improving the torque response were made. In the event that a faster torque response is required, it is simply a matter of changing the control code and performing additional testing to find out how fast a torque response can be achieved. To complete the picture, a negative step reference signal was applied from a steady state level of $6Nm$ to $1Nm$, and similar results were observed, illustrated in figure 6.5.

Figure 6.4: Dynamic Torque Response to a Step Reference

Figure 6.5: Dynamic Torque Response to Negative Step Reference

CHAPTER 7

Efficiency

7.1 Motor Losses

As outlined in the literature review (section 9.2), motor losses are comprised of five main components:

1. Stator copper loss, which is primarily a function of torque but is also affected by stator voltage, frequency and temperature;
 2. Stator iron loss, which is primarily a function of stator voltage and frequency;
 3. Rotor conductor loss, which is primarily a function of torque, but is also affected by slip and temperature;
 4. Friction and windage loss, which are functions of speed;
 5. Stray load losses consisting of additional iron and copper losses due to the quasi-sinusoidal field distribution in the air gap, and additional losses caused by stator slot effects and skin effects in the conductors which are primarily a function of torque, but are also affected by stator voltage, temperature and frequency [Kusko and Galler].
-

Only some of these loss components are controllable. Each loss component can be expressed in terms of its control variables so that its contribution to the overall loss can be minimized. Based on the equivalent circuit model for the induction motor shown in figure 7.1, an optimization routine for this controller was derived as follows.

7.2 Derivation of the Optimization Algorithm

Figure 7.1 shows a single phase of the standard "T" model which will be used as a tool for evaluating the operating losses. Power loss is a function of the magnitude of the current through each of the resistors shown. R_s represents the stator copper losses, R_r represents the rotor copper losses and R_c is an approximation to represent the cores losses which are comprised of eddy current and hysteresis losses. Following the convention found in literature, I_{sd} is the direct current component which represents the magnetizing current and I_{sq} represents torque-producing current through the rotor. The power loss function can be expressed as:

$$P_l = |I_{sd}|^2 R_s + |I_{sq}|^2 (R_s + R_r) + |I_{sd}|^2 \frac{|Z_m|^2}{R_c} \quad (7.1)$$

Referring to figure 7.1, Z_m is defined as the parallel combination of R_c and $j\omega_e L_m$, Z_r as the rotor impedance and Z_s as the stator impedance. Since $|j\omega_e L_m| > R_c$ for normal frequencies of operation:

$$|Z_m| \approx \omega_e L_m \quad (7.2)$$

Figure 7.1: "T" Equivalent Motor Model

Restating equation 3.50 as:

$$I_{sq}^2 = \frac{T^2}{L_m^2 I_{sd}^2} \quad (7.3)$$

it can be substituted into equation 7.1.

Taking the derivative with respect to time and setting it equal to zero to find the minimum I_{sd} yields:

$$0 = 2I_{sd}R_s - \frac{2T^2}{L_m^2 I_{sd}^3} (R_s + R) + 2I_{sd} \frac{\omega_e^2 L_m^2}{R_c} \quad (7.4)$$

Solving equation 7.3 for T^2 and substituting it into equation 7.4 then simplifying yields:

$$I_{sd} = I_{sq} \sqrt{\frac{R_s + R_r}{R_s + \frac{\omega_e^2 L_m^2}{R_c}}} \quad (7.5)$$

which predicts the optimum setting for I_{sd} in terms of the desired torque (I_{sq}), motor parameters and the fundamental electrical frequency.

At low speeds, core losses are small and copper losses dominate. As speed increases, so does the voltage across R_c , thereby increasing the core losses as the square of the voltage. At high speeds, core losses dominate.

The goal is to establish a flux level which will minimize power losses while maintaining a constant mechanical power output without sacrificing motor responsiveness. Referring to the vector control block diagram of figure 5.1, it is clear that there are two controlled inputs, I_{sd} and I_{sq} . I_{sd} controls flux and I_{sq} controls torque. Torque is the desired output from the motor for all the control schemes presented. Therefore, the controller should select the appropriate level of flux based on rotor speed and requested torque. Equation 7.5 accomplishes this. Empirically, the flux is higher at low speeds when copper losses dominate. As ω_e increases, the flux is reduced to account for the higher core losses.

7.3 Motor Power vs Flux Results

The purpose of calculating the power used by the induction motor is to find a relationship between flux and power for a given torque at a set speed. Since the motor can theoretically produce any desired torque at any desired flux, we wish to find the optimum flux at which the least power is delivered to the motor for a given torque at a set speed. The controller was used to calculate in real time the power delivered to the motor. A family of power curves was determined by running the motor at a constant speed and load while manually adjusting the flux. The minima of these curves is the desired operating point. Figure 7.2 shows the family of power curves at zero speed while producing torque of 4.69, 7.04, 9.39, 11.7, 14 and 16.4Nm (weights of 10, 15, 20, 25, 30 and 35lbf). Figure 7.3 shows the family

Figure 7.2: Power to Motor vs Magnetizing Flux at Zero Speed

of power curves at $\omega_r = 1000rpm$ while producing torque of 1, 2, 3, 4 and 5Nm.

Figure 7.4 shows the family of power curves with $\omega_r = 2000rpm$ and $T = 1, 2, 3$ and 4Nm. Figure 7.5 shows the family of curves with $\omega_r = 3000rpm$ and $T = 1, 2$ and 3Nm.

7.4 Flux-Optimization-Control Algorithm Evaluation

The FOCA shown in equation 7.5 was compared to data taken from the motor during normal operation to evaluate its ability to select the correct value for flux. The optimum level of flux was determined by manually adjusting the commanded flux so that the power into the motor was reduced to its minimum level (as measured by the DSP power routine shown in section 5.2) while maintaining torque

Figure 7.3: Power to Motor vs Magnetizing Flux at 1000rpm

Figure 7.4: Power to Motor vs Magnetizing Current at 2000rpm

Figure 7.5: Power to Motor vs Magnetizing Flux at 3000rpm

and speed output. The results of this testing are shown in table 7.6.

$2\omega_r$	rpm	ω_e	ω_s	T	I_{sq}	L_m	I_{sd}	FOCA
0	0	8.0	8.0	4.7	6.3	0.10	6.0	8.4
0	0	11.3	11.3	9.4	10	0.08	8.75	13.3
209.4	1000	215.5	5.6	3.0	5.8	0.12	4.1	3.9
209.4	1000	216.8	7.4	6.0	8.3	0.10	6.0	6.3
418.9	2000	433.5	14.6	4.0	9.7	0.14	2.2	3.1
418.9	2000	427.3	8.4	2.0	6.6	0.14	2.0	2.1
628.3	3000	634.6	6.3	3.0	9.8	0.14	1.8	2.2

(7.6)

7.5 Flux-Optimization-Control Algorithm Critique

It can be seen from the results presented in table 7.6 that for normal operation at speeds near its intended operating range, the FOCA works reasonably well. In addition, since it is based solely on motor parameters, it should be transportable to other induction motor platforms. However, the FOCA appears to break down at zero speed (which makes sense given the assumptions made in equation 7.2) as torque is increased. A more accurate FOCA was found fitting a curve to the data. The algorithm derived by curve fitting has the following form:

$$I_{sd} = \frac{0.025 |T| - \frac{\omega_r}{100000} + 0.05}{L_m} \quad (7.7)$$

which performs better than the motor-parameter-derived FOCA. Specifically, the FOCA in equation 7.7 performs very well at zero speed.

More research is required to derive a more accurate motor-parameter-based FOCA. It may be that the simplifying assumptions which lead to the motor equivalent circuit render every FOCA derived from the circuit inaccurate. Some literature explores solutions based on motor specific data, but these methods will not apply when a motor type is changed and only the motor parameters are known. It is a worthy goal to try to achieve a motor-parameter-based FOCA so that this controller can control many different motors efficiently. It is important to note that there is no closed form solution when the core losses are included in the analysis [Kusko and Galler].

CHAPTER 8

Future Work

8.1 Generation Control

Once the torque response of the engine has been identified, a control algorithm for the induction generator will need to be developed to match. In addition, the traction induction motors which will be running on this DC bus may need to have constraints placed on their demands for power to match the engine/induction generator's ability to deliver power.

8.2 Motor-Parameter-Based FOCA

It is believed that additional work needs to be done to develop a motor-parameter-based FOCA. The empirically derived solution works well for this induction motor, but the controller is not portable with this algorithm. While there is some information concerning flux optimization in literature, nothing was able to be directly implemented.

8.3 DC Offsets

DC offsets have been observed to cause torque ripple during motor operation. Various methods for automating the correction for DC offsets in the a/d and d/a circuits were tested. While there was a marked reduction in torque ripple after implementing these algorithms, the automated routines could not match the results that were manually attained by use of torque sensor feedback. Development of a better automated DC offset adjustment should improve the operation of this controller and mitigate torque ripple.

CHAPTER 9

Literature Review

The literature review section is divided into three sections: Vector Control; Efficiency; and Related Topics of Interest. The titles of papers, followed by citations, are used as headings. A short review of each paper may be followed by comments relevant to the project.

Not every paper cited in this chapter applies specifically to this project. Some relate to the final application of the project. Others were included because they address issues which could be evaluated in the future and may form the basis for ongoing research and improvements.

9.1 Vector Control

This section of the literature review cites various papers which describe different applications of vector control.

9.1.1 An FFT-Based Novel Approach to Noninvasive Speed Measurement in Induction Motor Drives [Ferrah, Bradley and Asher] Use of a rotor-position feedback transducer to achieve vector control disrupts the robust and mechanically simple qualities of an induction motor. It is therefore desirable to achieve control without a tachometer (tach-less control) and many attempts have been made to

this end. The following list outlines the attempts and briefly describes the method employed:

1. 1975 - Identification of the rotor position by use of motor current and voltage signals;
2. 1979 - Rotor slot harmonics were used to achieve slip frequency control; success was achieved at speeds greater than $300rpm$;
3. 1985 - Tach-less control was performed based on the phase relationship between current and voltage;
4. 1987 - Tach-less control was achieved using rotor slot harmonics in the speed range of 20%–100%. Lower end speed control was achieved with injection of a signal of constant frequency from the inverter to enhance the slot harmonic signal;
5. 1990 - Rotor slot harmonics were reconstructed from the DC link using switched capacitor filters to detect the rotor speed;
6. Recently - Schemes have been presented that use the digitized current and voltage waveforms to determine rotor speed. These schemes rely heavily on a motor-parameter-based model which changes with frequency and temperature.

One unifying characteristic is found in each of these methods: slow speed operation could not be achieved reliably. The authors [Ferrah, Bradley and Asher]

state that in general, these methods break down at slow speeds due to the “degradation of analog system performance at low frequencies.”

In contrast, the method they explore employs only a single input - the digitized stator current. With this input, the authors claim to be able to recreate rotor speed, supply frequency, slip and slip frequency. Unlike the aforementioned analog schemes, this method autotunes to the supply frequency and does not require a fundamental signal input from the inverter.

Current signals are digitized and processed with anti-aliasing filtering. The fundamental current signal is then identified. Assuming a maximum slip of 5%, a FFT algorithm is used to compute in real time the rotor slot harmonics. From the rotor slot harmonics, the speed of the rotor is derived. This method is claimed by the authors to operate accurately down to $2Hz$ (the lower limit of their inverter). In addition, it is claimed to be insensitive to motor parameter variations.

This method may be an effective alternative to add redundancy to the design implemented in this project. Should the rotor speed signal fail, this algorithm would allow continued operation with only slight degradation in machine performance.

9.1.2 A Physically Insightful Approach to the Design and Accuracy Assessment of Flux Observers for Field Oriented Induction Machine Drives [Jansen and Lorenz]

The vector control equations for direct rotor-field flux control in the stator frame of reference are derived. Various observers are presented:

1. Current-model, open-loop flux observer;
2. Voltage-model, open-loop flux observer;
3. Cancellation-method, open-loop flux observer;
4. Closed-loop, Gopinath-style rotor flux observer;

Modifications are made to these observers to implement velocity-invariant dynamics due to the non-linear controller for the non-linear flux model. Since the current model has the best low-speed performance and the voltage model has the best high-speed performance, the optimum observer is a combination of the two. This resultant observer is a closed-loop, velocity-invariant flux observer which is ideally suited for the entire dynamic range from low speed to field-weakening speed. Since this thesis project chose to use an observer operating in the rotor flux frame of reference, this paper is not directly applicable.

9.1.3 Simulation of Induction Machine Vector Control and Parameter Identification

[Wade, Dunnigan and Williams] This paper evaluates the suitability of commercially available software to perform simulation of vector control of an induction motor. Matlab, Matrix_x, user-written code and Simulink are compared. The authors' selected method was Simulink. The basic building blocks for simulation are reviewed and sample simulation blocks are shown. Sample equations for the mechanical induction machine and the twin-axis (d-q) observer in the stationary reference frame are shown. An example of all of the blocks together on a Simulink

screen is also presented. Differences between simulation results and actual applications, and their sources are discussed.

Simulations involving parameter identification are examined. Of particular interest is a method for determining R_r by use of the extended Kalman filter. Simulation of operation during voltage saturation conditions yields interesting results. Use of an algorithm to extend the effective saturation voltage is evaluated and found to increase the dynamic range by 15.5%. PWM supply is also simulated and compared to a linear source. Included in appendices are code listings of some of the model blocks used (including the induction machine and the extended saturation algorithm m-file listings). The authors conclude that simulation can provide information for testing new control theories and methods to gain insight to their operation prior to their implementation in hardware.

9.1.4 Vector Control of an Induction Motor [Andrade, Hughes and Corda] The concepts of induction-motor control are clearly and succinctly presented with a focus on the spatial distribution of quantities and their angular relationship. The many drawings that illustrate each of the important relationships between current, flux and torque give the control law an intuitive feel. Two observations concerning steady state operation are drawn:

1. The spatial angle between stator and rotor MMFs for a given set of stator currents is defined only by the slip frequency and the rotor time constant;

2. The resultant rotor-flux-linkage space-phasor is in space quadrature with the rotor MMF and is displaced in space by an angle with respect to the stator MMF. This angle depends only on the slip frequency and the rotor time constant.

It is observed that an induction motor will adjust to the load placed upon it during normal operation. From these observations, the author concludes that vector control will only affect performance during transient operation.

The claim that vector control cannot improve the performance of an induction motor during steady-state operation is based on two assumptions: the motor is current-fed and in a constant flux mode of operation. Since this project's controller is designed to modify the level of flux to achieve efficient motor operation within a large dynamic range, part of this paper is not applicable.

9.2 Efficiency

The efficiency section of the literature review cites relevant papers which address minimizing losses in motor operation.

9.2.1 A New Synthesis Method for Efficiency Optimized Speed Control System of Vector-Controlled Induction Machine [Murata, Tsuchiya and Takeda] The two controllable losses of an induction machine are copper losses and core losses. These losses are described by the stator current proportional to torque, and the current component in the direction of the rotor-linkage flux. Maximum efficiency is

achieved by controlling the ratio of these two currents, which is shown to be a function of motor parameters.

The derived equations show it is the core losses that dominate under light load conditions and the copper losses that dominate under heavy-load conditions. Good test results for this controller were attained in simulation.

9.2.2 A Novel Vector Control Hysteresis Current Controller for Induction Motor Drives [Chang, Lo and Pan] Switching losses and their effect on efficiency are addressed by this paper. A novel approach to adjusting the switching frequency based on the zero voltage vector and its derivative is presented. Control is accomplished in the d-q reference frame. Simulation and test results show switching frequency and therefore switching losses can be reduced with this approach.

9.2.3 A Systematic Approach to Selecting Optimal Flux References in Induction Motors [Bodson and Chiasson] This paper derives a flux optimization algorithm that can be implemented in real time. It also addresses both motor and generator cases. The algorithm is based on "well-known formulas" for synchronous machines. The solution for optimum flux is based on motor parameters and can therefore be implemented in conjunction with an adaptive parameter identification scheme. The optimum flux is found by solving a fourth-order algebraic equation. The Newton-Raphson root-finding method is recommended as adequate. A method for initializing this root-finding algorithm to ensure against finding extraneous roots is given. No claims are made concerning any actual application.

9.2.4 Characteristic Induction Motor Slip Values for Variable Voltage Part Load

Performance Optimization [Jian, Schmitz and Novotny] Optimization of induction motor efficiency is evaluated in terms of achieving optimum slip. Using the standard induction motor equivalent circuit, equations to calculate optimum slip are derived. Optimization is shown to be achieved, independent of mechanical output power, by reducing the voltage approximately as the square root of the load torque to achieve optimum slip. A shaded graphical analysis showing torque, per unit voltage and power factor clarifies where gains can be made. Observed values for X_m and R_c are shown as a function of voltage. It is interesting to note that the authors observed that R_c is not linear with an increase in voltage. It is further demonstrated that minimum current results in much less than optimum efficiency because it requires much less than optimum slip. Since voltage is not a directly controlled variable in this project's controller, this paper's methods would require adaptation. In addition, this method is meant for use at a single speed with a controller to achieve efficiency during light-load operation, which does not parallel the intended use of this project's motor controller.

9.2.5 Control Means for Minimization of Losses in AC and DC Motor Drives

[Kusko and Galler] There are five power loss components within an induction motor:

1. Stator copper loss, which is primarily a function of torque but is also affected by stator voltage, frequency and temperature;

2. Stator iron loss, which is primarily a function of stator voltage and frequency;
3. Rotor conductor loss, which is primarily a function of torque, but is also affected by slip and temperature;
4. Friction and windage loss, which are functions of speed;
5. Stray load losses consisting of additional iron and copper losses due to the quasi-sinusoidal field distribution in the air gap, and additional losses caused by stator slot effects and skin effects in the conductors, which are primarily functions of torque but are also affected by stator voltage, temperature and frequency.

The conditions for minimizing losses are to be attained given a constant speed and torque output for the motor. The power-loss components, considered in the optimization calculations, are the rotor and stator copper losses and the core losses. Since rotor speed and torque are constant, only the stator voltage and frequency can be controlled to achieve maximum efficiency. Since voltage and frequency are dependent, it is only necessary to solve for one. It is noted that there is no closed-form solution to this optimization problem when the core losses are included. The formula presented for the selection of the optimum frequency is based on a numerical solution:

$$f = \frac{60n_r}{n_s(an_r - b)}$$

where n_s is the synchronous speed; n_r is the rotor speed; and a and b are constants to be derived from experimental data. The theories are tested on a 75kW induction motor.

This paper thoroughly presents the material and supplies a practical solution which could be implemented on a controller with voltage and frequency as its controllable variables. Since the controllable variables in the vector control scheme for this development project are flux and torque, this optimization method would require adaptation. In addition, L_m is considered to be constant in the derivation, whereas L_m is a function of magnetizing current in this project's controller. No efficiency results were presented for speeds other than the rated speed of the motor.

9.2.6 Efficiency Considerations for Low Frequency Operation of Induction Motors

[Nishikata and Novotny] This paper evaluates the benefits and trade-offs of a small induction motor with a gear reduction mechanism versus a larger induction motor using and a vector-controlled inverter. A 5hp motor driving a 3:1 gear reduction is compared to a direct-drive 15hp motor. It is noted that core losses are significantly reduced for slow-speed operation of the motor. Equations to calculate core losses as a function of flux, slip and electrical frequency are given. It is interesting to note that at low speeds, optimum efficiency is achieved at minimum current, which is not the case at higher speeds.

While the slow-speed motor must have three times the horsepower rating to achieve equivalent torque, its inverter rating can be reduced if the motor is rewound

to allow full voltage at 20Hz. A rewound 15hp motor would then draw the same current as a 5hp motor and could be driven by a 5hp inverter. Since it is necessary to operate at a higher flux level at slow speeds to achieve efficiency, a larger inverter is needed for the 15hp motor.

The paper concludes that slow-speed operation of the 15hp motor is not as efficient as the 5hp motor and gear box. Although there are some losses associated with the inverter that drives the 15hp motor, these are comparable to the mechanical losses seen in the gear reduction box. It is shown that the slip loss in the rotor of the 15hp motor causes the huge loss in efficiency.

9.2.7 Efficiency Optimization of Vector-Controlled, Inverter-Fed Induction Motor

Drives [Murphy and Honsinger] A standard efficiency-optimization approach introduces the subject. References are made to the standard equivalent-circuit model. An interesting derivation demonstrates that operation at the maximum torque per ampere of stator current does not always correspond to optimum efficiency. Equations are given in the first part of this analysis to calculate optimum slip based on the standard equivalent-circuit model. It is then pointed out that significant harmonics, which arise from six-step or quasi-square-wave voltage inverters, are not taken into account. A modified induction-motor equivalent circuit is presented to take into account the harmonic currents. It is explained that harmonics $k = 7, 13, 19, 25, \dots (6n + 1)$, where $n = 1, 2, 3, \dots$ have a positive sequence and establish positive torque. Harmonics $k = 5, 11, 17, 23, \dots (6n - 1)$ have a neg-

ative sequence and establish negative torque. These effects are then taken into account and used as a part of the optimization evaluation. The implications of the second part of this analysis are minimal since the harmonics for this project's PWM inverter are so much higher in frequency than those generated by a six-step inverter. This information was not considered in the efficiency optimization implemented for this project.

9.2.8 Optimal Efficiency Analysis of Induction Motors Fed by Variable-Voltage and Variable-Frequency Source [Chen and Yeh] This paper addresses efficiency and some of the non-ideal factors that affect machine performance. The harmonic effect, the skin effect and the saturation effect are all addressed individually. All three effects are included in the simulations. A Thevenin equivalent circuit for the standard induction motor is derived to find relationships useful for obtaining optimal efficiency.

9.3 Related Topics of Interest

Papers of interest related to vector control or motor inverter operation are reviewed in this section.

9.3.1 Adaptive Control with Parameter Estimation for Induction Motor Drives [Bellini and Figalli] This paper draws on previous work to establish a fully linearized model of the induction motor. Vector control is applied in the rotating rotor flux frame. It is pointed out that it is not possible to measure the actual

rotor flux and that control must be based on an indirect method which uses the model and measured motor values such as stator voltage or current and rotor position.

Using the fully linearized model of the induction motor as a reference, an adaptive scheme is applied to estimate the motor parameters that vary during operation. Using this approach, motor parameters are found with "very short estimate times." The validity of this control approach is proven by use of simulation.

9.3.2 Analysis of a Concentrated Winding Induction Machine for Adjustable Speed Drive Applications Part 1 (Motor Analysis) [Toliyat, Lipo and White]

Conventional design of an induction motor aims for perfectly sinusoidal winding distributions on the stator and rotor to minimize space harmonics. When supplied with a balanced sinusoidal voltage, the resultant air gap field is nearly sinusoidal. However, when a PWM inverter supplies the voltage waveforms, the rectangular voltage and current pulses impressed on the motor no longer match the sinusoidally distributed motor windings.

"In a machine with a sinusoidal air-gap flux distribution, only a small part of the active iron core is near saturation at any one time. Only in the small region where the flux density wave is maximum is the iron saturated to the desired extent, while the rest of the iron does not reach full saturation. The iron would be better utilized if the air-gap flux density would have a more nearly rectangular form. The power density of the machine could clearly be improved if the stator and the

rotor flux density together result in a nearly rectangular air-gap flux density wave traveling at constant angular velocity. The peak fundamental component of such a square wave is $\frac{4}{\pi}$ times or 127% as large as that of a comparable sine distribution having the same peak flux density."

Implementation of rectangularly distributed windings resulted in a 15% increase in power output compared to a comparable sinusoidally wound induction motor when both were fed by a PWM inverter. Since this project utilizes a PWM inverter, overall efficiency should improve if these principles are applied to motor design.

9.3.3 Effect of PWM Inverters on AC Motor Bearing Currents and Shaft Voltages

[Erdman et al] Voltages are induced in the rotor with respect to the stator as a result of small magnetic field dissymmetries in the air gap, which are inherent in a practical machine design. While modern induction motors have been designed to minimize rotor voltages during normal operation, rotor voltages are present during transient operation. While the induced rotor voltages are small in magnitude, significant current can flow due to the low impedance of the stator and rotor. The impedance of the bearings varies with type of bearing, speed of operation, vibration, load and type of lubricant. Current flow is nonlinear, and resembles a capacitor breakdown. At first, no current flows, as the lubricant acts as a dielectric. If the voltage exceeds this breakdown threshold, current flows. This current flow causes electrical pitting, which results in premature bearing failure.

The flow of current through the bearings of induction motors and the premature failures it causes have been known for some time. Steps have been taken to mitigate these effects in sinusoidally driven induction motors. The paper examines the effects of PWM waveforms on rotor voltage and bearing currents. The tests show a significant increase in rotor voltage and the resultant bearing current as a result of PWM waveforms. The authors conclude that bearings in PWM-driven induction motors are subject to increased currents and therefore fail more frequently. An electrostatically shielded induction motor is presented as a solution.

While our testing for this project has shown good control and robust operation of an induction motor, the issue of the effects of PWM waveforms on the long-term reliability of an induction motor have not been addressed.

9.3.4 Exponentially Convergent Rotor Resistance Estimation for Induction Motors

[Marino, Peresada and Tomei] Rotor resistance (or rotor time constant) variation is a well documented source of problems in motor control applications. This paper cites many real-time and off-line solutions. Motor parameters, except rotor resistance, are used with an exponentially convergent algorithm to determine rotor resistance. Simulation results are presented. In addition, experimental results are presented, and the method is shown to be tolerant of voltage and current transducer noise and a $0.8m$ sec sampling rate.

9.3.5 Improved Rotor Models for Vector Controllers [Healey, Williamson and Smith]

The fixed-parameter linear model of the squirrel cage induction motor is simple

to work with. Unfortunately, this simplicity comes at the price of accuracy. The detuning of motor parameters yields inefficient operation and adversely affects transient response. One solution relies on an adaptive routine to correct for motor parameters. The paper presents an advanced motor model and clearly guides the determination of requisite model parameters. The model presented increases the standard three-impedance, T-equivalent circuit to a five-impedance circuit. While the simplified model may be adequate for smaller induction machines, larger applications would benefit from a more accurate model.

9.3.6 Increased Dynamic Range for Three Phase PWM Motor Drives [Clower]

The paper presents a novel way to increase the dynamic range of a PWM adjustable-speed drive. During high-speed operation, the back emf from the motor saturates the current amplifiers. Given that the ideal current amplifiers are really voltage amplifiers with current feedback, they saturate when the back emf increases to where it leaves no voltage headroom for operation. One solution is to simply increase the bus voltage to increase the dynamic range. Once the bus voltage has reached its maximum, however, the maximum speed for a given level of flux has been reached. It is interesting to note that only one of the current amplifiers saturates at a time, as the voltage signals are 120° out of phase. If phase A and B are assumed to be fed by ideal current sources and phase C is fed with a voltage source, phase C can shift the neutral point. Once this shift occurs, additional voltage headroom becomes available to the current amplifiers of phases A and B.

While the voltage waveforms do not look sinusoidal with respect to a ground reference, they are sinusoidal with respect to each other. This idea is currently being patented.

There is a predicted increase in the dynamic range of 15% by use of this control scheme. Given the huge dynamic range intended for this project's controller, and the limits imposed on its DC bus voltage, this subject is of interest for future study. Despite the likelihood of complications in implementation, given this project's hardware design, this may prove to be a helpful technique.

9.3.7 Losses and Parasitic Torques in Electric Motors Subjected to PWM Waveforms

[De Buck] A theoretical analysis of the effects of PWM waveforms on an induction motor is presented. Included in this discussion are copper losses and skin effects and their effect on winding temperature. The use of PWM waveforms instead of a sinusoidal supply will result in a 70% reduction in service lifetime. Also covered are the parasitic torque effects that result from PWM operation. Though the paper is dated, it raises some relevant questions about the design of induction motors for use in a PWM environment.

9.3.8 Measurement of Rotor Time Constant Taking into Account Magnetizing

Flux in the Induction Motor [Kwon, Lee and Youn] An accurate value for the rotor time constant is essential for high-performance, servo-motor-like control of an induction motor. To further complicate the issue, the rotor time constant varies with temperature and flux. The proposed method injects a DC level into two of

the windings during steady-state operation, which is then removed. The stator voltage is then measured to arrive at the rotor time constant.

This method appears to be fundamentally sound for determining the rotor time constant. The authors do not address the practical effects of injecting a DC current during steady-state operation. It is believed that there will, at the very least, be a significant torque ripple.

9.3.9 Minimizing DC Energy Storage [Alakula and Persson] It is anticipated that power quality regulators will eventually prevent the use of the traditional rectifier-capacitor front ends on all equipment connected to the power line. Some reasons for this prohibition include the harmonics placed on the line, the larger gauge wires required to handle the current surges and the higher losses associated with currents that are neither proportional nor in phase with the line voltage. To remedy the problem, the author has proposed the use of two identical inverters, one operating as a rectifier on the line, the other to control an adjustable speed drive to run an induction motor. The thrust of this paper is that the DC link capacitance can be reduced substantially by coordinating the switching of the two inverters. Since the electrolytic DC capacitors are a weak link in the long-term reliability of the overall system, it is beneficial to eliminate them. In fact, the author claims that with this method, the DC link capacitance can be reduced to $88\mu F$ for a $10kW$, $380V$ inverter. This level of capacitance can easily be attained by a high-voltage plastic-type capacitor. This should increase the life and reliability of

the inverter.

While it may take some significant redesign work, it is believed that this concept is directly applicable to an induction generator supplying an induction motor. Reducing bus capacitance could yield the described benefits to this development project.

9.3.10 Modeling and Control of Induction Motors with Magnetic Saturation in Field Coordinates [Hsu] Magnetic saturation is a non-linear effect that adversely affects the accuracy and performance of motor controllers. Since magnetic saturation is a function of instantaneous stator and rotor currents, it can change very rapidly and is therefore difficult to identify and predict. While there are predictive models in related literature, there is little information on controlling this phenomenon. The paper presents a control scheme that compensates for saturation effects and renders a reduction in parameter-induced transients and steady-state torque errors. Simulations of the saturated induction motor validate the control scheme. Although this control scheme looks interesting, no practical results were presented.

9.3.11 Rotor Time Constant Identification in Vector Controlled Induction Motor Applied Flux Model Reference Adaptive System (MRAS) [Jemli et al.] The paper addresses the identification of the rotor time constant. Two independent flux observers, one based on the voltage signals and the other based on the current signals, contribute to the rotor flux equation. Using this distinction, the MRAS

method identifies the correct rotor flux. Testing was performed between 200rpm and 3000rpm.

9.3.12 Transient Effects in Application of PWM Inverters to Induction Motors

[Persson] The premature failures of standard squirrel cage induction motors from apparent overvoltage conditions in PWM applications prompted the author to investigate. After evaluating numerous failed motors, the following factors were identified as contributing to premature motor failure:

1. RMS input voltage to the inverter;
2. rise time of the inverter output voltage wave;
3. cable length from the inverter to the motor;
4. motor insulation: wire insulation and testing, stator winding design, impregnation etc.

Working in conjunction with manufacturers' recommendations, the author presents various recommendations for induction motor design for use in PWM applications.

9.4 Conclusion

Related vector control literature was reviewed, a prototype controller was constructed, control code was written and the control theories were tested. It worked.

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A. Data and Calculations

The following pages are data and calculations performed on Mathcad 6.0.

Data from synchronous operation (slip=0) at 2000 RPM

Measured frequency of input current (Hz)	Measured time lag of stator input current (seconds)	Measured peak to peak amplitude of input current (A)	Measured peak to peak amplitude of input voltage (V)	Best fit for motor mutual inductance (H)
$f_e := \begin{bmatrix} 67 \\ 67 \\ 66.46 \\ 66.46 \\ 67 \end{bmatrix}$	$dt := \begin{bmatrix} 3.2 \cdot 10^{-3} \\ 3.2 \cdot 10^{-3} \\ 3.3 \cdot 10^{-3} \\ 3.35 \cdot 10^{-3} \\ 3.34 \cdot 10^{-3} \end{bmatrix}$	$I_a := \begin{bmatrix} 0.245 \cdot 7.5 \\ 0.376 \cdot 7.5 \\ 0.51 \cdot 7.5 \\ 0.71 \cdot 7.5 \\ 1.0 \cdot 7.5 \end{bmatrix}$	$V_a := \begin{bmatrix} 105 \\ 152 \\ 205 \\ 296 \\ 373 \end{bmatrix}$	$L := \begin{bmatrix} 0.14 \\ 0.135 \\ 0.135 \\ 0.135 \\ 0.12 \end{bmatrix}$

Stator resistance and copper losses are assumed constant and unaffected by flux or speed.

$$R_s \approx 1.0 \quad R_c \approx 300$$

Calculated stator voltage for a given current amplitude and frequency:

$$V(S, I_{in}, L_m) := R_s \cdot I_{in} + \frac{R_c \cdot S \cdot L_m}{(R_c + S \cdot L_m)} \cdot I_{in}$$

$$i := 0..4 \quad V_{out_i} := V(2 \cdot \pi \cdot j \cdot f_{e_i}, I_{a_i}, L_i)$$

Calculated percent error:

$$\text{VoltageError}_i := \frac{V_{a_i} - |V_{out_i}|}{V_{a_i}} \cdot 100 \quad \text{PhaseError}_i := \frac{dt_i \cdot f_{e_i} \cdot 360 - \arg(V_{out_i}) \cdot \frac{180}{\pi}}{dt_i \cdot f_{e_i} \cdot 360} \cdot 100$$

Results

Peak to peak stator voltage calculated from model (V):

$ V_{out_i} $
106.634
158.013
212.659
296.054
374.934

Measured peak to peak stator voltage (V):

V_{a_i}
105
152
205
296
373

Percent error:

VoltageError_i
-1.556
-3.956
-3.736
-0.018
-0.519

Stator current phase lag calculated from model (deg):

$\arg(V_{out_i}) \cdot \frac{180}{\pi}$
77.917
78.268
78.345
78.345
79.311

Measured stator current phase lag (deg):

$dt_i \cdot f_{e_i} \cdot 360$
77.184
77.184
78.954
80.151
80.561

Percent error

PhaseError_i
-0.949
-1.405
0.772
2.253
1.551

Data from synchronous operation (slip=0) at 1000 RPM

Measured frequency of input current (Hz)	Measured time lag of stator input current (seconds)	Measured peak to peak amplitude of input current (A)	Measured peak to peak amplitude of input voltage (V)	Best fit for motor mutual inductance (H)
$f_e := \begin{bmatrix} 33 \\ 33.5 \\ 33 \\ 33.2 \end{bmatrix}$	$dt := \begin{bmatrix} 6.05 \cdot 10^{-3} \\ 6.35 \cdot 10^{-3} \\ 6.58 \cdot 10^{-3} \\ 6.7 \cdot 10^{-3} \end{bmatrix}$	$I_a := \begin{bmatrix} 0.25 \cdot 7.5 \\ 0.51 \cdot 7.5 \\ 1.027 \cdot 7.5 \\ 1.55 \cdot 7.5 \end{bmatrix}$	$V_a := \begin{bmatrix} 49.8 \\ 99.6 \\ 192 \\ 215 \end{bmatrix}$	$L := \begin{bmatrix} 0.125 \\ 0.125 \\ 0.12 \\ 0.09 \end{bmatrix}$

Stator resistance and copper losses are assumed constant and unaffected by flux or speed.

$$R_s \approx 1.0 \quad R_c \approx 300$$

Calculated stator voltage for a given current amplitude and frequency:

$$V(S, I_{in}, L_m) := R_s \cdot I_{in} + \frac{R_c \cdot S \cdot L_m}{(R_c + S \cdot L_m)} \cdot I_{in}$$

$$i := 0..3 \quad V_{out}_i := V(2 \cdot \pi \cdot j \cdot f_{e_i}, I_{a_i}, L_i)$$

Calculated percent error:

$$\text{VoltageError}_i := \frac{V_{a_i} - |V_{out}_i|}{V_{a_i}} \cdot 100 \quad \text{PhaseError}_i := \frac{dt_i \cdot f_{e_i} \cdot 360 - \arg(V_{out}_i) \cdot \frac{180}{\pi}}{dt_i \cdot f_{e_i} \cdot 360} \cdot 100$$

Results

Peak to peak stator voltage calculated from model (V):

$ V_{out}_i $
48.613
100.66
191.784
218.857

Measured peak to peak stator voltage (V):

V_{a_i}
49.8
99.6
192
215

Percent error:

VoltageError_i
2.383
-1.065
0.113
-1.794

Stator current phase lag calculated from model (deg):

$\arg(V_{out}_i) \cdot \frac{180}{\pi}$
82.86
82.818
82.965
83.38

Measured stator current phase lag (deg):

$dt_i \cdot f_{e_i} \cdot 360$
71.874
76.581
78.17
80.078

Percent error

PhaseError_i
-15.285
-8.145
-6.133
-4.123

Data from operation at 2000 RPM with a 4 Nm load from DC generators

Measured frequency of input current (Hz)	Measured time lag of stator input current (seconds)	Measured peak to peak amplitude of input current (A)	Measured peak to peak amplitude of input voltage (V)
$f_e := \begin{bmatrix} 70 \\ 69 \\ 68 \\ 68 \\ 67.5 \\ 67.5 \\ 67.5 \end{bmatrix}$	$dt := \begin{bmatrix} 0.79 \cdot 10^{-3} \\ 0.87 \cdot 10^{-3} \\ 0.93 \cdot 10^{-3} \\ 1.0 \cdot 10^{-3} \\ 1.1 \cdot 10^{-3} \\ 1.4 \cdot 10^{-3} \\ 1.62 \cdot 10^{-3} \end{bmatrix}$	$I_a := \begin{bmatrix} 1.74 \cdot 7.5 \\ 1.60 \cdot 7.5 \\ 1.50 \cdot 7.5 \\ 1.44 \cdot 7.5 \\ 1.41 \cdot 7.5 \\ 1.41 \cdot 7.5 \\ 1.47 \cdot 7.5 \end{bmatrix}$	$V_a := \begin{bmatrix} 236 \\ 255 \\ 275 \\ 300 \\ 334 \\ 350 \\ 387 \end{bmatrix}$
Measured slip speed (rad/sec)	Best fit for motor mutual inductance (H)		
$ws := \begin{bmatrix} 23 \\ 17 \\ 14.5 \\ 12 \\ 10 \\ 8 \\ 6.5 \end{bmatrix}$	$L := \begin{bmatrix} 0.135 \\ 0.135 \\ 0.135 \\ 0.135 \\ 0.135 \\ 0.13 \\ 0.12 \end{bmatrix}$		

Stator resistance and copper losses are assumed constant and unaffected by flux or speed.

$$R_s \approx 1.0 \quad R_c \approx 300$$

Calculated rotor resistance reflected into stator:

$$i := 0..6 \quad \text{slip}_i := \frac{ws_i}{f_{e_i} \cdot 2 \cdot \pi} \quad R_i := \frac{0.9}{\text{slip}_i}$$

Calculated stator voltage for a given current amplitude and frequency:

$$V(S, I_{in}, L_m, R_r) := R_s \cdot I_{in} + \frac{1}{\left(\frac{1}{R_c} + \frac{1}{S \cdot L_m} + \frac{1}{R_r} \right)} \cdot I_{in}$$

$$i := 0..6 \quad V_{out_i} := V(2 \cdot \pi \cdot j \cdot f_{e_i}, I_{a_i}, L_i, R_i)$$

Calculated percent error:

$$\text{VoltageError}_i := \frac{V_{a_i} - |V_{out_i}|}{V_{a_i}} \cdot 100$$

$$\text{PhaseError}_i := \frac{dt_i \cdot f_{e_i} \cdot 360 - \arg(V_{out_i}) \cdot \frac{180}{\pi}}{dt_i \cdot f_{e_i} \cdot 360} \cdot 100$$

Results

Peak to peak stator voltage
calculated from model (V):

$ V_{out_i} $
217.466
251.705
262.907
289.144
317.368
357.355
397.631

Measured peak to peak
stator voltage (V):

V_{a_i}
236
255
275
300
334
350
387

Percent error:

VoltageError_i
7.853
1.292
4.398
3.619
4.98
-2.101
-2.747

Stator current phase lag
calculated from model (deg):

$\arg(V_{out_i}) \cdot \frac{180}{\pi}$
14.421
19.081
21.946
25.694
29.629
35.732
42.875

Measured stator current
phase lag (deg):

$dt_i \cdot f_{e_i} \cdot 360$
19.908
21.611
22.766
24.48
26.73
34.02
39.366

Percent error

PhaseError_i
27.563
11.706
3.601
-4.959
-10.845
-5.032
-8.914

Data from operation at 1000 RPM with a 3 Nm load from DC generators

Measured frequency of input current (Hz)	Measured time lag of stator input current (seconds)	Measured peak to peak amplitude of input current (A)	Measured peak to peak amplitude of input voltage (V)
$f_e := \begin{bmatrix} 37.5 \\ 36 \\ 35.5 \\ 35 \\ 34.6 \\ 34.3 \\ 34 \\ 33.7 \\ 33.7 \\ 33.6 \end{bmatrix}$	$dt := \begin{bmatrix} 1.46 \cdot 10^{-3} \\ 1.86 \cdot 10^{-3} \\ 2 \cdot 10^{-3} \\ 2.4 \cdot 10^{-3} \\ 2.72 \cdot 10^{-3} \\ 3.13 \cdot 10^{-3} \\ 3.7 \cdot 10^{-3} \\ 4.4 \cdot 10^{-3} \\ 4.7 \cdot 10^{-3} \\ 5.2 \cdot 10^{-3} \end{bmatrix}$	$I_a := \begin{bmatrix} 1.71 \cdot 7.5 \\ 1.51 \cdot 7.5 \\ 1.34 \cdot 7.5 \\ 1.27 \cdot 7.5 \\ 1.20 \cdot 7.5 \\ 1.16 \cdot 7.5 \\ 1.17 \cdot 7.5 \\ 1.26 \cdot 7.5 \\ 1.41 \cdot 7.5 \\ 1.65 \cdot 7.5 \end{bmatrix}$	$V_a := \begin{bmatrix} 100 \\ 106 \\ 116 \\ 129 \\ 142 \\ 154 \\ 166 \\ 184 \\ 205 \\ 221 \end{bmatrix}$

Measured slip speed (rad/sec)	Best fit for motor mutual inductance (H)
$ws := \begin{bmatrix} 27 \\ 20 \\ 15 \\ 12 \\ 9.5 \\ 7.5 \\ 5.7 \\ 4.5 \\ 3.6 \\ 3.2 \end{bmatrix}$	$L := \begin{bmatrix} 0.125 \\ 0.125 \\ 0.125 \\ 0.125 \\ 0.125 \\ 0.125 \\ 0.12 \\ 0.12 \\ 0.09 \\ 0.09 \end{bmatrix}$

Stator resistance and copper losses are assumed constant and unaffected by flux or speed.

$$R_s \approx 1.0$$

$$R_c \approx 300$$

Calculated rotor resistance reflected into stator:

$$i := 0..9 \quad \text{slip}_i := \frac{ws_i}{f_{e_i} \cdot 2 \cdot \pi} \quad R_i := \frac{0.9}{\text{slip}_i}$$

Calculated stator voltage for a given current amplitude and frequency:

$$V(S, I_{in}, L_m, R_r) := R_s \cdot I_{in} + \frac{1}{\left(\frac{1}{R_c} + \frac{1}{S \cdot L_m} + \frac{1}{R_r}\right)} \cdot I_{in} \quad V_{out_i} := V(2 \cdot \pi \cdot j \cdot f_{e_i}, I_{a_i}, L_i, R_i)$$

Calculated percent error:

$$\text{VoltageError}_i := \frac{V_{a_i} - |V_{out_i}|}{V_{a_i}} \cdot 100 \quad \text{PhaseError}_i := \frac{dt_i \cdot f_{e_i} \cdot 360 - \arg(V_{out_i}) \cdot \frac{180}{\pi}}{dt_i \cdot f_{e_i} \cdot 360} \cdot 100$$

Results

Peak to peak stator voltage
calculated from model (V):

$ V_{out_i} $
107.463
116.047
126.199
137.802
148.916
161.83
177.576
203.614
189.944
224.278

Measured peak to peak
stator voltage (V):

V_{a_i}
100
106
116
129
142
154
166
184
205
221

Percent error:

VoltageError_i
-7.463
-9.478
-8.792
-6.824
-4.87
-5.084
-6.973
-10.66
7.345
-1.483

Stator current phase lag
calculated from model (deg):

$\arg(V_{out_i}) \cdot \frac{180}{\pi}$
12.847
17.358
22.773
27.67
33.341
39.43
47.624
53.406
64.108
66.073

Measured stator current
phase lag (deg):

$dt_i \cdot f_{e_i} \cdot 360$
19.71
24.106
25.56
30.24
33.88
38.649
45.288
53.381
57.02
62.899

Percent error

PhaseError_i
34.821
27.99
10.902
8.499
1.592
-2.021
-5.159
-0.047
-12.429
-5.046