DESIGNING AND IMPLEMENTING A DSP BASED VARIABLE-SPEED DRIVE FOR THEATRE STAGE SYSTEMS

by

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Abstract

Today's power electronics technology and digital control system provide a convenient means to boost the performance of electrical drive systems. While the industry has witnessed a variety of applications of electrical variable-speed drives, this work is focusing on designing and implementing a DSP controlled 3-phase variable-speed drive for a theatre stage system to achieve automatic operations and accurate position control.

A cascade structure with feed-forward reference is selected as the control algorithm for this application. As a key factor in motion control, profiles of acceleration, speed and position (ASP) are also investigated. A trapezoidal speed profile with limited jerks is recommended for the final implementation. In addition, some operational conditions for safe and agreeable operations have been discussed and identified.

Before proceeding with the experiment, the dynamics of the whole system including the controller, the voltage source inverter, the induction machine and the stage has been evaluated by MATLAB/Simulink. The non-linear behavior of induction machines is studied and a linearized induction machine model under variable-frequency conditions is developed to verify the closed loop stability of the whole system.

A test facility is then constructed to validate system performance. A DSP controlled 3-phase inverter is built and tested. Multiple measures are implemented into the IGBT drive for obtaining immunity to high common-mode dv/dt noise. Quantitative methods for sizing the gate resistor are also presented in this paper. Four different motion control algorithms are employed to find the most suitable for this application. Both experimental and simulation results indicate that closed loop position control is critical to achieve required accuracy of position for the stage system.
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Chapter 1 Introduction

1.1 Current System Overview

The Frederic Wood Theatre, located at the north end of University of British Columbia (UBC) Point Grey campus, has been home to UBC's theatrical performances for over fifty years. The construction of the current building finished in 1963.

As the heart of a theatre, the stage serves as a space for actors or performers and a focal point for the members of the audience. As is necessary in a drama, sceneries are required to be changed according to the mood, and rotary stage can serve a performance to the need of scenery change. There is a 27-foot rotary stage in the Frederic Wood Theatre. The rotary stage is driven by a 3-hp dc motor via a steel cable coupled the motor and the stage. When changing the scenery, the stage is operated manually. The position control of the stage is somewhat coarse since it can not realize automatic control and only depends on the operator's experience. An experienced operator knows exactly when to turn off the machine such that the huge stage together with the actors standing on it can stop at the desired position. The scenery setting differs time to time, and so does the number of actors. All those bring complexity and uncertain factors to the operation of the stage. Consequently, a demand of an automatic stage drive and control system comes into being.

Induction motors, owing to their simplicity, reliability and relatively low cost compared to dc motors, have gained the favor of users in many industrial applications. The fast advent of power electronics in the past decades presents users mature and reliable devices, which makes it very easy and cost-effective to feed an induction machine with adjustable-frequency ac voltage sources. Actually a variable-frequency drive plus an induction motor have become a quite standard solution for variable-speed applications, which is also the choice being implemented in this project.

1.2 Thesis Objectives and Outlines

The objective of this thesis is to identify several approaches for controlling the stage system. The design will be implemented in a DSP-based digital controller. An
experimental variable-speed drive is also implemented in order to verify the design and simulation results. Specific objectives include:

- To find safe operation areas for the stage
- To design and compare different motion control algorithms and find a proper one for this application
- To develop an experimental variable-speed drive
- To test the experimental system and evaluate the conceptual design and simulation

The thesis is organized into six chapters. Chapter 1 gives a brief introduction of the current stage system and outlines the objectives of this thesis. In Chapter 2, some basic principles of mechanics are reviewed and a safe operation area for the stage is proposed. Chapter 3 is focused on the analysis of different motion control algorithms and the stability of the whole system. Chapter 4 will be dealing with the implementation of an experimental variable-speed drive. Some designing concerning both hardware and software are presented. Selective experimental results are included in Chapter 5. The last chapter concludes the design and the implementation and proposes some work needed to be done in the future.
Chapter 2  Dynamics of the Stage Mechanic System

2.1  Elementary Principles of Mechanics

In the stage and its drive system, both mechanical and electromagnetic energies exist and there is the exchange between the two types of energies. Since the whole system involves mechanical and electrical engineering, it is necessary to recall some basic concepts and laws related to mechanics. The most general equation to describe rotational motion is

\[ T_M - T_L = J \frac{d\omega}{dt} \]  

(2.1)

where \( T_M \) is the driving torque and \( T_L \) the load torque, \( \omega \) is the angular speed, \( J \) is the overall moment of inertia of the rotating mass about the axis of rotation. As speed is the derivative of the shaft position \( \theta \), we have

\[ T_M - T_L = J \frac{d\omega}{dt} = J \frac{d^2\theta}{dt^2} \]  

(2.2)

where \( \alpha = \frac{d\omega}{dt} = \frac{d^2\theta}{dt^2} \)  

(2.3)

is the angular acceleration.

The rotational system can be considered as a second-order differential equation, with the input as the driving torque and the load torque and the output as speed and position [1]. The following diagram describes such a mechanical system with a lumped mass.

![Rotating object and its block diagram representation](image)

Fig. 2.1 Rotating object and its block diagram representation
2.1.1 Moment of Inertia

The moment of inertia, defined as

\[ J = \int_0^M r^2 dM \]  

(2.4)

is the integral of the moments of the masses relative to the rotation axis. Considering a homogenous cylinder with mass evenly distributed, the moment of inertia can be obtained by

\[ J = M \frac{R^2}{2} \]  

(2.5)

with M being the mass of the object, and R is the radius of the cylinder. An estimation of the moment of inertia of the stage is listed in Appendix A.

2.1.2 Effect of Gearing

As the tangential force per rotor surface is limited by iron saturation and heat losses in the conductors, the direct coupling of a low speed motor with the load may result in an unnecessarily large motor [2]. It is then often preferable to employ transmissions, operating the motor at a high speed and thus increasing its power density. This is the approach in the current stage driving system. Transmission systems affect the moment of inertia of the coupled rotating masses as well as the torque applied to the rotation axis. In our mechanical system, there is a simple transmission system connecting the motor to the stage by a belt pulley and a wire rope twisted between the pulley and stage.

An ideal transmission system is shown in Fig. 2.2, where two pulleys are engaged by a wire rope without backlash or slip. Assuming the left hand wheel is the driving pulley, we have

\[ T_1 - r_1 F_1 = J_1 \frac{d\omega_1}{dt} \]  

(2.6)

where \( F_1 \) is the force exerted by pulley 2. If there is no load torque applied, for pulley 2

\[ r_2 F_2 = J_2 \frac{d\omega_2}{dt} \]  

(2.7)

As \( F_1 \) and \( F_2 \) is a pair of forces against each other, \( F_1 = F_2 \)

(2.8)

Assuming there is no slip between the two pulleys, \( \omega_1 r_1 = \omega_2 r_2 \)

(2.9)
Substituting (2.7), (2.8) and (2.9) into (2.6) results in
\[ T_x = J_x + \lambda^2 = \left[J_x + \left(\frac{r_1}{r_2}\right)^2 J_2\right] \frac{\omega_1}{r_2} \] (2.10)

If we define transmission ratio as \( \lambda = \frac{r_2}{r_1} \), \( \lambda > 1 \),

\[ T_1 = [J_1 + \frac{J_2}{\lambda^2}] \frac{\omega_1}{r_1} = J_{1e} \frac{\omega_1}{r_1} \] (2.11)

\[ = J_{1e} \frac{\omega_1}{r_1} \] (2.12)

where \( J_{1e} \) is the effective moment of inertia reflected at the axis of pulley 1. It contains a component referred from pulley 2. For convenience, \( J_2 \) is used to represent the moment of inertia of pulley 2 referred to the axis of pulley 1 hereafter, i.e.

\[ J_2 = \frac{J_2}{\lambda^2} \] (2.13)

and \[ J_{1e} = J_1 + J_2 \] (2.14)

Expressing force \( F_1 \) and \( F_2 \) in terms of torques results in

\[ F_1 = \frac{T_1}{r_1} \quad \text{and} \quad F_2 = \frac{T_1}{r_2} \] (2.15) (2.16)

where \( T_1 \) is the equivalent torque referred to the axis of pulley 2. Substituting (2.15) and (2.16) into (2.8), we have

\[ \frac{T_1}{r_1} = \frac{T_1'}{r_2} \] (2.17)

and \[ T_1' = \lambda T_1 \] (2.18)
From the above equations, it can be concluded that if $\lambda>1$, the transmission system reduces the speed output by $\lambda$ times, while the torque output is increased by $\lambda$ times. For the following analysis, we assume pulley 1 as the shaft of the induction machine and pulley 2 as the shaft of the stage. Therefore

\[
J_{Me} = J_M + J'_L = J_M + \frac{J_L}{\lambda^2}
\]

\[
J_{Le} = J'_M + J_L = J_M \lambda^2 + J_L
\]

with $J_{Me}$ representing the effective moment of inertia referred to the motor side, $J_{Le}$ being the effective moment of inertia referred to the stage side, $J'_L$ being the moment of inertia of the stage referred to the motor side, and $J'_M$ being the moment of inertia of the motor referred to the stage side.

When simulating the system dynamics, the moment of inertias and torques are all referred to the motor side. A transmission ratio $\lambda=600$, which optimizes the travel time for a half revolution travel, is used in simulations. Although the stage has a comparatively huge moment of inertia with respect to its own shaft, the equivalent moment of inertia referred to the motor side only has the same order that the motor has.

2.1.3 Friction

The stage absorbs the electromagnetic energy from the motor and stores it as its kinetic energy. However, losses may occur when the stage is rotating. Since the rotation speed of the stage is very low, losses like windage loss can be neglected. The only mechanical loss concerned is the friction loss. In the following analysis, the friction is assumed constant as long as the stage is rotating. For simulation purposes the coefficient of friction is assumed as 0.05, a typical value of rotational friction.

2.2 Safe Riding Conditions

Before proceeding to the system design, we have to find a way to decide what the maximum angular speed $\omega_{\text{max}}$ and maximum angular acceleration $\alpha_{\text{max}}$ should be since they associate with the safety for those on-board. We can divide the safety problem into two levels, mechanically safe and physiologically safe. For mechanical safety, the
maximum angular speed $\omega_{\text{max}}$ and maximum angular acceleration $\alpha_{\text{max}}$ should be within the range such that motors and the stage can stand. In addition, slip should not happen between the on-board actors and the stage. For physiological safety, $\omega_{\text{max}}$ and $\alpha_{\text{max}}$ should be within the range that those on-board can stand and have no dizziness or fear caused by the motion of the stage. As physiological problems are out of the scope of this thesis, analysis will be focused on the mechanical level only.

Rotating along the shaft is a typical movement for the stage. The angular speed of the stage changes in the acceleration/deceleration period. Moreover, the angular acceleration/deceleration might change as well. Therefore the motion of the stage is a varying-speed varying-acceleration revolution. If an object is rotating with a varying speed, its acceleration can be divided into two components, a radial/centripetal acceleration that changes the direction of the angular speed, and a tangential acceleration that changes the magnitude of the angular speed. Fig. 3.3 shows the forces and accelerations applied to a person standing on a stage.

![Fig. 2.3 Force analysis for variable speed revolution](image)

Assume the stage is rotating at an angular speed $\omega_L$ and an angular acceleration $\alpha_L$. $F_f$ and $\alpha_f$ represent the force applied to the person and the actual acceleration for
the person respectively. \( \overline{F_f} \) (the same to \( \overline{\alpha_f} \)) can be divided into two components, \( \overline{F_c} \) in the normal direction and \( \overline{F_T} \) in the tangential direction. Recall

\[
F_c = ma_c = m\omega_i^2 r \tag{2.21}
\]

and

\[
F_T = ma_T = m\alpha_L r \tag{2.22}
\]

with \( m \) being the mass of the person and \( r \) being the rotational radius of the person. From (3.1) and (3.2) we have

\[
F_f = \sqrt{F_c^2 + F_T^2} = mr\sqrt{\omega_i^4 + \alpha_L^2} \tag{2.23}
\]

\( \overline{F_f} \), subscripted with \( f \) referring to friction, is actually a friction acting as the force to keep the person moving with the stage simultaneously. The maximum of the friction is given by

\[
F_{f,max} = \mu_s mg \tag{2.24}
\]

where \( \mu_s \) is the coefficient of static friction between the stage and person’s shoes (or his feet if he is barefoot) and \( g \) is the acceleration due to gravity. For safety reasons, the inequality

\[
F_{f,max} \geq F_f \tag{2.25}
\]

must be satisfied to prevent slip from happening. Substitute (3.3) and (3.4) into (3.5) and eliminate \( m \) results in

\[
\mu_s g \geq r\sqrt{\omega_i^4 + \alpha_L^2} \tag{2.26}
\]

or

\[
\sqrt{\omega_i^4 + \alpha_L^2} \leq \frac{\mu_s g}{r} \tag{2.27}
\]
The most serious case, in terms of $r$, happens when people stand at the edge of the stage, thus $\frac{\mu_s g}{r}$ reaches its minimum $\frac{\mu_s g}{R}$ with $R$ the radius of the stage. Fig. 3.4 shows the relationship of maximum angular acceleration versus speed under different friction conditions. Safe operation states are those points embraced by the curves and the $y$ axis.

![Graph showing the relationship of maximum angular acceleration versus speed under different friction conditions.](image)

**Fig. 2.4 Maximum angular acceleration versus speed under different friction conditions**

From the figure it can be concluded that in low to medium speed area, i.e. $\omega_L < 0.6 \text{rad/s}$, acceleration sets the limit to the available safe operation area. However, as speed increases, allowable acceleration decreases dramatically. In designing the motion control algorithm, it is desirable to maintain the operation area of the stage in the lower left corner shown in Fig. 2.4.
Chapter 3  Control Algorithm

3.1 General Motion Control Requirements

Before designing the control system, some objectives should be identified. The core control target for the whole system is the stage position. If we regard the position command as a step input for the stage system, a step response is expected. From control theory's point of view, the position response should be monotonic and it is necessary to have a one-shoot movement for the stage. That means the position control will be either critically damped or over-damped. An under-damped system is not allowed because back-and-forth moves caused by positioning are not suitable in this application, which may make people on the stage feel uncomfortable and may even cause safety problems.

![Fig. 3.1 Desirable position control curve](image)

Second, accuracy of the stage position has to meet the user's need. It is desirable that the positioning error measured at the edge of the stage could be maintained at as low as about 5 mm for each travel.

In addition, every move of the stage should not take too long, which means the response time of the position control has to be considered. Although this demand is not critical for designing the control system, a system with fast response is desirable if this is achieved by not sacrificing the two above mentioned requirements.
Apart from position control, the stage should be operated safely according to the discussion in chapter 2. Jerks, defined as $\frac{d^2\omega}{dt^2}$, caused by uncontinuous changes in either speed or acceleration should also be avoided.

### 3.2 Cascaded Structure of Motion Control

There is general agreement that the most effective control scheme for drives is a cascaded structure, which has a fast inner loop, limiting torque or acceleration, to which an outer speed control loop is superimposed [2]. In order to control position, a position loop is again superimposed on the speed loop. Fig. 3.1 illustrates a typical cascade structure of drive including position, speed and acceleration control.

![Cascaded control structure of drive including position, speed and acceleration control](image)

Fig. 3.2 Cascaded control structure of drive including position, speed and acceleration control

The sequence of the controller, depicted in Fig. 3.2, from the most inner loop to the outer loop is torque-acceleration-speed-position. It is worth mentioning that the whole cascade control structure can be realized only if the torque is feasible as a control target for the machine. In some cases that torque or current of the machine is not controllable, scalar control for example, the inner loops of the uncontrollable quantities have to be eliminated, thus a simplified cascade structure is used. The idea of the simplified cascade control is exemplified in Fig. 3.3.
Although the cascade control structure shows some merits, it is likely to respond slower to the changes of the reference than an equivalent single loop control [2]. In addition, since the position loop can only be configured either critically-damped or under-damped, the system destines to have a comparatively slow performance.

In order to overcome the disadvantage, an extra external reference generator is introduced which feeds the inner loops in parallel the feed-forward reference signals. Indicated in Fig. 3.4, feed-forward signals from a reference model are added to the cascade multiple-loop control structure.

Fig. 3.3 Simplified cascaded control structure of drive including position and speed control

Fig. 3.4 Cascaded control structure with feed-forward signals
Because the reference generator is outside the feedback loops, the stability of the control system stays unaffected [2]. At the same time, slow dynamic response can be improved by introducing the feed-forward signals.

An extra advantage of the feed-forward structure is that the system could be operated according to a preset acceleration-speed-position (ASP for short) profile. Since we could incorporate the safe operation conditions into a pre-defined acceleration-speed-position (ASP) profile, this technique is widely used in drives for elevators, skytrains, and electric cars. Several acceleration-speed-position profiles are discussed hereafter.

3.3 Acceleration-speed-position (ASP) profiles

3.3.1 Time-optimal ASP Profile

Although position control is the main concern of the system, travel time for the stage to move from a previous position to a new position might be optimized such that it will not take the audiences too long waiting for the stage moving. Let’s begin with the time-optimal ASP profile, from which other ASP profiles are developed.

The well known time-optimal solution that could be derived with the calculus of variables, consists of two integrals of maximum acceleration/deceleration (|\( \alpha \)|\( \leq \alpha_{\text{max}} \)) and possibly an intermediate interval with constant maximum speed (|\( \omega \)|\( \leq \omega_{\text{max}} \)) [2]. This is exemplified in Fig. 3.5 for two different travel distances.

Acceleration, speed and position are plotted versus time for two cases depending on whether the maximum speed is reached. The left hand part of Fig. 3.5 shows the case with the maximum speed being reached. Beginning at time zero in steady state condition, \( \theta_0 = \omega_0 = \alpha_0 = 0 \), a new position reference is required. A time optimal transient includes three phases, maximum acceleration, operation at a constant maximum speed and maximum deceleration to the commanded position. If the desired new position is not far from the previous position, maximum speed may be reached when \( \theta < \frac{\theta_{\text{set}}}{2} \), and the transient switches to the maximum deceleration phase immediately after the acceleration, which is shown on the right hand of Fig. 3.5.
In the long travel distance case, i.e. the left part shown in Fig. 3.5, the total acceleration time

\[ t_x = \frac{\omega_{\text{max}}}{\alpha_{\text{max}}} \]  

(3.1)

The traveled distance during 0 to \( t_1 \) is

\[ \theta_1 = \int_0^{t_1} \left( \int_0^{t_1} \alpha dt \right) dt = \frac{1}{2} \alpha_{\text{max}} t_1^2 = \frac{1}{2} \frac{\omega_{\text{max}}^2}{\alpha_{\text{max}}} \]  

(3.2)
Assuming the maximum deceleration and acceleration are equal in magnitude, the travel time during the deceleration period is the same as \( t_1 \), and so is the traveled distance. The travel time for the constant speed region is

\[
t_2 - t_1 = \frac{\theta_{\text{set}} - 2 \theta_1}{\omega_{\text{max}}} = \frac{\theta_{\text{set}}}{\omega_{\text{max}}} - \omega_{\text{max}}
\]

(3.3)

In conclusion, the total travel time is

\[
t_3 = 2t_1 + (t_2 - t_1) = 2\frac{\omega_{\text{max}}}{\alpha_{\text{max}}} + \left( \frac{\theta_{\text{set}}}{\omega_{\text{max}}} - \omega_{\text{max}} \right) = \frac{\theta_{\text{set}}}{\omega_{\text{max}}} + \omega_{\text{max}}
\]

(3.4)

(3.4) gives the answer to the minimum travel time for a limited acceleration/deceleration and limited maximum speed case, where the travel distance is long enough such that maximum speed has been reached. Similarly, total travel time needed in a short travel distance case can be deduced. Table 3.1 gives a brief list for the travel time for different travel distances.

**Table 3.1 Travel time for time-optimal ASP profile**

<table>
<thead>
<tr>
<th>Long Travel Distance</th>
<th>Short Travel Distance</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Travel Time</strong></td>
<td><strong>( \theta_{\text{set}} &gt; \frac{\omega_{\text{max}}^2}{\alpha_{\text{max}}} )</strong></td>
</tr>
<tr>
<td>( t_1 )</td>
<td>( \frac{\omega_{\text{max}}}{\alpha_{\text{max}}} )</td>
</tr>
<tr>
<td>( t_2 )</td>
<td>( \frac{\theta_{\text{set}}}{\omega_{\text{max}}} )</td>
</tr>
<tr>
<td>( t_3 )</td>
<td>( \frac{\theta_{\text{set}}}{\omega_{\text{max}}} + \frac{\omega_{\text{max}}}{\alpha_{\text{max}}} )</td>
</tr>
</tbody>
</table>

In a time-optimal ASP profile, there exist sudden changes of acceleration, or called jerks at \( 0, t_1, t_2 \) and \( t_3 \), shown in Fig. 3.5 where the first row indicates the derivative of acceleration, donated as \( \gamma \), reaches infinity at those moments. In applications involving passengers, jerks should be avoided for the sake of riding safety and comfortability. Thus, a new ASP profile with limited jerk is developed from the time-optimal profile. The new profile is called limited-jerk ASP profile.
3.3.2 Limited Jerk ASP Profile

If the derivative of acceleration $\gamma$ is restricted to a maximum value $\gamma_{\text{max}}$, i.e. $|\gamma| \leq \gamma_{\text{max}}$, the time-optimal ASP profile will then be changed to a limited jerk profile, which is illustrated in Fig. 3.6 for two different travel distances.

When jerk limitation takes effect in the control, we obtain a trapezoidal acceleration characteristic, which becomes triangular for the short-travel-distance case shown in the right hand in Fig. 3.6. During the jerk limitation period, the speed
characteristic follows a function of second order parabola, and the distance function assumes the shape of a third order parabola [3].

Table 3.2 Travel time and traveled distance for limited-jerk ASP profile

<table>
<thead>
<tr>
<th>Time</th>
<th>Travel Time</th>
<th>Traveled Distance</th>
</tr>
</thead>
<tbody>
<tr>
<td>$t_4$</td>
<td>$\frac{\alpha_{\text{max}}}{\gamma_{\text{max}}}$</td>
<td>$\frac{\alpha_{\text{max}}}{\gamma_{\text{max}}}$</td>
</tr>
<tr>
<td>$t_1 - 2t_4$</td>
<td>$\frac{\omega_{\text{max}}}{\alpha_{\text{max}}} - \frac{\alpha_{\text{max}}}{\gamma_{\text{max}}}$</td>
<td>$\frac{\omega_{\text{max}}}{\alpha_{\text{max}}} - \frac{\alpha_{\text{max}}}{\gamma_{\text{max}}}$</td>
</tr>
<tr>
<td>$t_1$</td>
<td>$\frac{\omega_{\text{max}}}{\alpha_{\text{max}}} + \frac{\alpha_{\text{max}}}{\gamma_{\text{max}}}$</td>
<td>$\frac{\omega_{\text{max}}}{\alpha_{\text{max}}} + \frac{\alpha_{\text{max}}}{\gamma_{\text{max}}}$</td>
</tr>
<tr>
<td>$t_3 - t_2 = t_1$</td>
<td>$\frac{\omega_{\text{max}}}{\alpha_{\text{max}}} + \frac{\alpha_{\text{max}}}{\gamma_{\text{max}}}$</td>
<td>$\frac{\omega_{\text{max}}}{\alpha_{\text{max}}} + \frac{\alpha_{\text{max}}}{\gamma_{\text{max}}}$</td>
</tr>
<tr>
<td>$t_2$</td>
<td>$\frac{\theta_{\text{set}}}{\omega_{\text{max}}}$</td>
<td>$\theta_{\text{set}} - \frac{\omega_{\text{max}}}{2} \left[ \frac{\omega_{\text{max}}}{\alpha_{\text{max}}} + \frac{\alpha_{\text{max}}}{\gamma_{\text{max}}} \right]$</td>
</tr>
<tr>
<td>$t_3$</td>
<td>$\frac{\theta_{\text{set}}}{\omega_{\text{max}}} + \frac{\omega_{\text{max}}}{\alpha_{\text{max}}} + \frac{\alpha_{\text{max}}}{\gamma_{\text{max}}}$</td>
<td>$\theta_{\text{set}}$</td>
</tr>
</tbody>
</table>

Travel time and traveled distance at different moments are shown in Table 3.2 for the long travel distance. For the limited jerk profile, the total travel time increases by $\frac{\alpha_{\text{max}}}{\gamma_{\text{max}}}$ compared with the time-optimal condition. If $\gamma_{\text{max}}$ is set to infinity, the limited-jerk profile degrades back to the unlimited-jerk profile, i.e. the time-optimal ASP, and the total travel time decreases to $\frac{\theta_{\text{set}}}{\omega_{\text{max}}} + \frac{\omega_{\text{max}}}{\alpha_{\text{max}}}$ since $\frac{\alpha_{\text{max}}}{\gamma_{\text{max}}}$ = 0. Actually the time-optimal profile can be regarded as a subset of the limited-jerk profile while setting $\gamma_{\text{max}}$ to infinity. The limited-jerk control prolongs the travel time, and the increased time depends on how $\gamma_{\text{max}}$ is set.

A third ASP profile, called the modified limited-jerk ASP profile has been proposed and developed which can effectively reduce jerks generated by the induction
machine during the first few seconds of the start-up transient. The modified limited-jerk ASP profile is attached in Appendix F.

3.4 Motor Drive

3.4.1 Scalar Control vs. Vector Control

The previous sections are focused on motion control. The objective for this section is to choose a proper motor drive technique and design the whole control system.

From the historical perspective, scalar control methods, based on the steady-state equivalent circuit of motors, were first realized to control induction machines. Later, vector control technologies are applied with the fast progress of power electronics and digital controllers. Based on transient equivalent circuit of the motors, vector control techniques can provide high-quality control over the whole speed range while scalar control techniques exhibit comparatively lower performance and even poor performance within low speed ranges. However, scalar control, in terms of implementation, is easier than vector control because it, apart from rotor speed, does not necessarily need to measure state variables [4]. Moreover, the performance can be improved by incorporating extra controls. As the demand for the dynamic performance of the stage system is not very high, applying scalar control techniques in this project might be good enough to meet the requirements while keeping the overall topology simple and being easy for implementation.

Scalar methods control the amplitude and frequency of the control variable, either the stator voltage or current, such that the air-gap flux is kept constant in order to control the electromagnetic torque working around motor’s rated condition. To keep as simple as possible, a constant volts-per-hertz control will be implemented in the motor drive. This is a speed control strategy which is designed to accommodate variable speed commands by using the inverter to apply a voltage of correct magnitude and frequency so as to approximately achieve the commanded speed. The output from volts-per-hertz control is a variable-amplitude variable-frequency voltage signal, which can be realized by applying pulse-width modulation (PWM) to a dc voltage source and thus can avoid using current sensors. However, as the torque and flux are both related and are not decoupled in the stator current, any change in the stator current influences the flux, thus resulting in a
slower torque response. Moreover, the stability problems are the disadvantages of this method.

3.4.2 Electromagnetic Torque under Variable-frequency Operations

Fig. 3.7 shows a steady-state torque-speed characteristic (line 1) for typical induction machines. Assume P1 at line 1 is the previous operating point with the electric torque $T_{e1}$, rotor speed $\omega$ and stator voltage frequency $f_1$. If the motor is required to accelerate, a voltage with higher frequency $f_2$ and higher magnitude, when volts-per-hertz control is used,

![Figure 3.7 Electromagnetic torque change during speed-up process](image)

is applied to the stator. The new torque-speed characteristic is plotted as dotted line in Fig. 3.7, donated as line 2. Since the speed of the rotor can not change instantaneously after the command, the operating point moves from P1 to P2, which is at line 2. From the figure we can find that the new electromagnetic torque is increased to $T_{e2}$, the extra torque will drive the motor to accelerate until a new balanced state is achieved.

On the other hand, feeding a voltage with reduced frequency and magnitude will result in a deceleration, which is shown in Fig. 3.8. Line 1 is the previous torque-speed curve with P1 the previous operating point while dotted line 2 is the new one with P2 the new operating point. We notice that the electromagnetic torque is negative. This operation is known as the braking state for the induction machine. During the braking state, the kinetic energy of the rotating system is fed back to the driver, and the induction machine acts as a generator.
The feedback energy, which will be blocked by the diode-based converter from sending back to the grid, must be dissipated within the drive. Otherwise the regenerated current might charge the dc bulk capacitor to reach a damageable voltage level. In this project a braking branch consisting of a switch-controlled resistor is connected in parallel with the dc capacitor. The resistor will be turned on to dissipate the regenerated energy when unsafe dc voltage is detected.

3.5 Controller Design and Simulation Results

In order to ease designing the controller and verify if the whole system can work well, simulations are run first before stepping to hardware implementation. Because the constant volts-per-hertz control, a speed control, is used, the inner loop of acceleration in Fig. 3.4 is eliminated and a two-level cascade structure with position and speed will be employed in the final implementation. The whole system for simulation, expressed in block diagram in Fig. 3.9, consists of a cascaded motion controller with feed-forward reference signals, a voltage source inverter (VSI), an induction machine and the stage system, which is simplified and represented by a lumped moment of inertia. All simulations are based on a MATLAB/Simulink environment.
A motor test has been carried out to estimate machine parameters which are not given by the manufacturer. Testing method and tested results are attached in Appendix B. In addition, some mechanical parameters of the stage itself and the transmission system have been estimated as well. The simulation is based on those calculated values.

A qd0 model of the induction machine [5] is then employed for simulation. A detailed motor model is given in Appendix C. The rotor is short-circuited and the zero sequence circuit is neglected in the simulation.

A constant volts-per-hertz control is used and third-harmonic injection is integrated into the VSI block. Power electronics devices are represented by ideal switches because the device-level performance is not the main concern of this simulation.

Speed loop and position loop, apart from the mechanical system, are the two most important factors in the whole system. The discussion will begin with the inner loop, the speed loop.

### 3.5.1 Speed Loop Design

Speed, although it is not the final control target for the system, is of importance to realize a safe operation of the stage system. In addition, a well-designed speed control can help its outer loop, the position loop, achieve a high-accuracy control. Here both
open-loop and closed-loop of speed control will be compared. Moreover, two types of closed-loop control for speed will be explored.

In order to generate torque, a slip has to be developed in an induction machine [6]. That means the rotor will only run at the synchronous speed when there is no resistive torque, otherwise the mechanical speed will always be different from the synchronous speed. In addition, the slip is a non-linear function of synchronous speed and electromagnetic torque. By changing the synchronous speed at a given rotor speed, the torque will be changed as well. The non-linearity definitely forces the system to move into a new but hard-to-predict balanced condition.

As always, P or PI control comes to our first consideration. A proportional (P) controller of the speed loop is shown in Fig. 3.10.

![Fig. 3.10 P controller for speed loop](image)

If we assume the gain of volts-per-hertz controller and VSI are unity and neglect the dynamics of the induction machine, we have

\[
(\omega_{\text{ref}} - \omega_r)K_p(1-s) = \omega_r
\]

where \(s\) is the slip. Rearrange (3.5), it yields

\[
\frac{\omega_r}{\omega_{\text{ref}}} = \frac{K_p(1-s)}{1 + K_p(1-s)}
\]

From (3.6) we find the gain of speed of the system included in the speed loop is a function of slip. Slip vs. \(\frac{\omega_r}{\omega_{\text{ref}}}\) is plotted in Fig. 3.13 when \(K_p = 10\) and \(K_p = 20\). From Fig. 3.13 we find that P controller can increase the speed gain in high slip region. However its performance is fair in low slip area, or even poorer than open loop control around the rated slip.
Apart from the P controller, the reduced rotor speed can be compensated by other methods to achieve a near-synchronous rotor speed. A technique called slip compensation is explained in Fig. 3.11. Given the torque-speed characteristic of the machine defined by the solid curve and assuming a load torque $T_L$, the motor will develop a slip corresponding to the length of AB. If the stator frequency is boosted such that the new torque speed curve is obtained, as defined by the dashed line, which makes the operation point move from A to B, then the speed drop caused by the slip is eliminated and the motor will be running at the desired speed.

![Fig. 3.11 Linear slip compensation method](image)

A block diagram of the slip compensation method is then presented in Fig. 3.12.

![Fig. 3.12 Block diagram of slip compensation](image)

If we assume the volts-per-hertz controller and the voltage source interface (VSI) are all linear and they have unity gain in terms of speed, and neglect the dynamic process, it can be seen from Fig. 3.12 that

$$\omega_{ref} + \omega_{ref} - \omega_{r}(1-s) = \omega_{r}$$  \hspace{1cm} (3.7)

where $s$ is the slip. Rearranging (3.7) we have
To compare different controllers, a MATLAB program is developed to compute gains of speed for proportional control and slip compensation. Results are plotted in Fig. 3.13. It is noted that the gain of speed is a function of machine slip, whereas the slip is a non-linear function of machine torque.

\[
\frac{\omega_r}{\omega_{ref}} = \frac{2 - 2s}{2 - s}
\]

(3.8)

Fig. 3.13 Gains of speed, a function of induction machine slip

It can be concluded that slip compensation does reduce the speed error along all the slip range. However, the compensation effect is not remarkable at the rated condition that \(s = 0.03\). A proportional controller can greatly reduce the speed error in high slip area, but it is ineffective, or even worse than open loop control, in the low slip range.

In addition, simulation has been carried out to verify if the two controllers are sensitive to measurement error. In the simulation, a 1% noise was applied to the measured speed signal. As shown in Fig. 3.14(a), the P controller amplifiers measurement errors. The output of the inverter is ruined by the noise even if the proportional ratio \(K_p\)
is set to 5. Therefore a pure proportional controller with a high coefficient is not suitable for speed control in this project. Simulation results, given by Fig. 3.14(b), show that slip compensation is not sensitive to measurement errors.

![P controller and slip compensation](image)

Fig. 3.14 Simulation results of speed when measurement error applied

Other than the P controller, a PI controller might be an alternative. Actually as the movement of the system follows a pre-defined ASP profile, any speed error accumulated will be reflected in the position signal. Since the closed position control will correct the position error, which in term can be regarded as the integration controller to the speed, there is no need to design an extra integration controller for speed.

Based on the above analysis, slip compensation will be used as the closed loop speed control for the rest of the simulation and experiment. The next figure illustrates the speed curves obtained by a MATLAB/Simulink simulation, which employs a time-optimal ASP profile. Speeds are obtained for open loop control as well as slip compensation control.
3.5.2 Position Control

There are some options for position control, including P, PI, PD or PID. Although integration can cancel out the static error, it may introduce position oscillation, which is undesirable during the final positioning period. Differentiation can predict the operation trend, but in real application, it amplifies high frequency component, which sometimes is problematic. Therefore only proportion control is used for position controller.

A high proportion coefficient $K_p$ is desirable in getting a small error system. But as the nonlinear essential of the induction machine, a high gain in position error could
saturate the speed output, sometimes it may even cause the electromagnetic torque collapse. Assume A is the previous operating point with a corresponding torque Te in Fig. 3.17. Due to the sluggish response of the motor during the first few seconds, position controller tries to reduce the position error by sending an increased-speed command to the motor. The more aggressive the controller is, the higher the output frequency is. However, as shown in Fig. 3.17, the operating point could move from A to A', with which a reduced electromagnetic torque is associated. This reversed behavior of the induction machine will be regarded as a more sluggish response from the perspective of the position controller, and the whole system will be trapped into vicious cycles.

![Fig. 3.17 Non-linear torque-speed characteristic of induction machines](image)

The non-linear torque-speed characteristic occurred between the zero speed and the breakdown speed of induction machines forces users to give up using aggressive position controls. By trial and error, the range for Kp is found between 1 to 4 to assure the algorithm work well during all the speed range. It also shows, from the stability verification in 3.5.4, the system with those coefficient settings is far from the stable margin obtained under rated conditions of the induction machine.

### 3.5.3 Comparison of Different Controls

There are totally four combinations for speed and position controls: open loop speed and open loop position control (OSOP), closed loop speed and open loop position control (CSOP), open loop speed and closed loop position control (OSCP), and closed
loop speed and closed loop position control (CSCP). To compare their performances, simulations have been carried out and results for each case are plotted together in Fig. 3.18. In the simulation, position is referred to the stage side in the form of pulses. A 12-bit encoder is assumed to be used, which can output 4096 pulses per revolution. The transmission ratio is set to 600:1 from the motor shaft to stage shaft.

The simulation indicates that closed loop speed control can achieve smaller speed error than open loop speed control. Closed loop position control can also produce a much smaller position error than open loop position control. However, in terms of position error, closed loop speed closed loop position control (CSCP) does not show any advantage over the open loop speed closed loop position control (OSCP).
3.5.4 Control Stability

Although different motion control algorithms have been simulated, it is still not clear if they are stable. In order to study the stability of different control methods, a
linearized model for induction machines has been derived. As the drive involves variable-frequency operation, the magnitude of the stator voltage, the frequency of the stator voltage, and the frequency of the rotor will all influence the electromagnetic torque output of the machine. In the linearized model those factors are decoupled and represented in three transfer functions. Please refer to Appendix D for the detailed development.

In order to ease the stability analysis, a subsystem, consisted of a constant volts-per-hertz controller, a VSI and the induction machine, is regarded as a whole and the transfer function \( \frac{\Delta \omega_r}{\Delta \omega_{ref^*}} \) of the subsystem will be derived first. With the subsystem staying constant, different control loops can be cascaded to it and the corresponding transfer functions can be easily derived. Since the induction machine is a non-linear element, only its transfer function under the rated condition will be used hereafter.

During the startup process for an induction machine, one of the eigenvalues of the motor has a positive real part, which means the machine itself is in an unstable state. There are some control technologies which can push poles into the left-side of the s plane, making the system stable and controllable. However, it is out of the scope of this paper.

![Block diagram of the V/f control, VSI and the induction machine](image)

Fig. 3.19 Block diagram of the V/f control, VSI and the induction machine

Fig. 3.19 illustrates the subsystem mentioned above. To obtain the transfer function \( \frac{\Delta \omega_r}{\Delta \omega_{ref^*}} \) of the whole system, it needs to know the transfer function of the induction machine first.
In the following analysis, all variables are in per-unit. Another assumption is that \( \Delta T_L = 0 \). If \( u \) is the input vector of the blocks embraced by the dashed line in Fig. 3.19,

\[
\mathbf{u} = \begin{bmatrix} \Delta v \\ \Delta \omega_r \end{bmatrix}
\]

(3.9)

where \( \Delta \mathbf{v} = \begin{bmatrix} \Delta v_{qs}^e & \Delta v_{ds}^e & \Delta v_{qs}^r & \Delta v_{ds}^r \end{bmatrix}^T \). Eliminating \( \Delta T_L \) from (d.12) we have

\[
\begin{bmatrix} \frac{\Delta i}{\Delta \omega_r} \\
\end{bmatrix} = \begin{bmatrix} -Q^{-1}(r + \omega_d F + \omega_q G) & -Q^{-1}G_i_0 \\
\frac{1}{2H}i_0^T(G^T + G) & 0 \\
\end{bmatrix} \begin{bmatrix} \frac{\Delta i}{\Delta \omega_r} \\
\end{bmatrix} + \begin{bmatrix} Q^{-1} & -Q^{-1}F_0 \\
0 & 0 \\
\end{bmatrix} \begin{bmatrix} \frac{\Delta v}{\Delta \omega_r} \\
\end{bmatrix}
\]

(3.10)

where \( r, F, G, Q \) are defined in (d.2), and \( \Delta \mathbf{i} = \begin{bmatrix} \Delta i_{qs}^e & \Delta i_{ds}^e & \Delta i_{qs}^r & \Delta i_{ds}^r \end{bmatrix}^T \). If \( \Delta \omega_r \) is the only variable concerned and we eliminate \( \Delta T_L \) in the output equation of (d.13), we get

\[
\Delta \omega_r = \begin{bmatrix} 0 \\
1 \\
\end{bmatrix} \begin{bmatrix} \Delta i \\
\Delta \omega_r \\
\end{bmatrix}
\]

(3.11)

Defining

\[
\mathbf{x} = \begin{bmatrix} \Delta i \\
\Delta \omega_r \\
\end{bmatrix}
\]

(3.12)

\[
y = \Delta \omega_r 
\]

(3.13)

\[
\mathbf{A} = \begin{bmatrix} -Q^{-1}(r + \omega_d F + \omega_q G) & -Q^{-1}G_i_0 \\
\frac{1}{2H}i_0^T(G^T + G) & 0 \\
\end{bmatrix}
\]

(3.14)

\[
\mathbf{B} = \begin{bmatrix} Q^{-1} & -Q^{-1}F_0 \\
0 & 0 \\
\end{bmatrix}
\]

(3.15)

\[
\mathbf{C} = \begin{bmatrix} 0 \\
1 \\
\end{bmatrix}
\]

(3.16)

\[
\mathbf{D} = [0]
\]

(3.17)

we can express the system in standard linear system dynamical equations as

\[
px = \mathbf{A}x + \mathbf{B}u
\]

(3.18)

\[
y = \mathbf{C}x + \mathbf{D}u
\]

(3.19)

If (3.18) is solved for \( x \) and the result is substituted into (3.19), we have

\[
y = \mathbf{C}(sI - \mathbf{A})^{-1}\mathbf{B}u + \mathbf{D}u = \mathbf{C}(sI - \mathbf{A})^{-1}\mathbf{B}u
\]

(3.20)
where the operator $p$ has been replaced by Laplace operator $s$ commonly used in transfer function formulation.

If all variables are in per-unit, \( \frac{\Delta v}{\Delta \omega_e} = 1 \) is satisfied for constant volts-per-hertz control, and for volts-per-hertz control we have

\[
\Delta v = \Delta \omega_{ref^*}.
\]

Assuming $\Delta v_{qs}^*$ is in phase with $\Delta v$ and $\Delta v_{ds}^* = \Delta v_{qr}^* = \Delta v_{dr}^* = 0$, we have

\[
\Delta v = \begin{bmatrix} \Delta v_{qs}^* & \Delta v_{ds}^* & \Delta v_{qr}^* & \Delta v_{dr}^* \end{bmatrix}^T = \begin{bmatrix} \Delta v & 0 & 0 & 0 \end{bmatrix}^T
\]

(3.22)

Now we can relate $\Delta \omega_{ref^*}$ with the input vector $u$ of the induction machine by substituting (3.21) into (3.22) and we have

\[
u = \Delta \omega_{ref^*} W
\]

where $W = \begin{bmatrix} 1 & 0 & 0 & 0 \end{bmatrix}^T$. Combining (3.20) and (3.23), we have

\[
y = C(sI - A)^{-1} Bu = \Delta \omega_{ref^*} C(sI - A)^{-1} BW
\]

(3.24)

and

\[
\frac{\Delta \omega_r}{\Delta \omega_{ref^*}} = C(sI - A)^{-1} BW
\]

(3.25)

which is the transfer function of the whole system indicated in Fig. 3.19. Under the rated condition, the explicit expression of the transfer function \( \frac{\Delta \omega_r}{\Delta \omega_{ref^*}} \) is obtained as

\[
G_1(s) = \frac{\Delta \omega_r}{\Delta \omega_{ref^*}} = \frac{8.53e2s^3 + 8.41e4s^2 + 1.19e8s + 5.24e9}{s^5 + 200s^4 + 1.54e5s^3 + 1.31e7s^2 + 4.39e8s + 5.23e9}
\]

Eigenvalues of $G_1(s)$ are computed and listed below. Under the rated condition, all eigenvalues have negative real part.

| Table 3.3 Eigenvalues of open loop speed, open loop position control |
|------------------------|-----------------|-----------------|
| Eigenvalues            | -54.9±j371      | -31.7±j19.4     | -26.8           |
Fig. 3.20 Block diagram of the closed loop speed open loop position control (CSOP)

For closed loop speed open loop position control (CSOP), as shown in Fig. 3.20, the transfer function is

\[ G_2(s) = \frac{\omega_r}{\omega_{ref}} = 2 \cdot \frac{G_1(s)}{1 + G_1(s)} = \frac{1.70e3s^3 + 1.68e5s^2 + 2.37e8s + 1.05e10}{s^5 + 200s^4 + 1.54e5s^3 + 1.32e7s^2 + 5.57e8s + 1.05e10} \]

Eigenvalues of \( G_2(s) \) are listed below.

<table>
<thead>
<tr>
<th>Eigenvalues</th>
<th>-54.6±j371</th>
<th>-25.6±j35.2</th>
<th>-39.3</th>
</tr>
</thead>
</table>

For open loop speed closed loop position control, as shown in Fig. 3.21, the transfer function can be deduced as

\[(\theta_{ref} - \theta_r)G_\theta(s) + \omega_{ref} = \omega_{ref}^*, \text{ moreover } \omega_{ref} = \theta_{ref}s, \ \omega_r = \theta_r s,\]

so we get \((\theta_{ref} - \theta_r)G_\theta(s) + \theta_{ref} s = \frac{\theta_r s}{G_1(s)}\). Rearranging it we have

\[
\frac{\theta_r}{\theta_{ref}} = \frac{G_\theta(s) + s}{G_\theta(s) + \frac{s}{G_1(s)}} = \frac{852s^4 + 8.75e4s^3 + 1.19e8s^2 + 5.71e9s + 2.09e10}{s^6 + 200s^5 + 1.54e5s^4 + 1.31e7s^3 + 4.39e8s^2 + 5.70e9s + 2.09e10}
\]

where \( G_\theta(s) = K_\theta = 4 \).
Fig. 3.21 Block diagram of open loop speed closed loop position control (OSCP)

Eigenvalues of the open loop speed closed loop position control are listed in Table 3.5. Different from the previous controls, there is one more eigenvalue added caused by the closed loop of position.

Table 3.5 Eigenvalues of open speed, closed position control

| Eigenvalues | -54.4±j371 | -33.2±j16.1 | -18.4   | -5.93   |

Finally the stability of the closed loop speed closed loop position control has been checked. The block diagram is illustrated in Fig. 3.22.

Fig. 3.22 Block diagram of closed loop speed closed loop position control (CSCP)
The transfer function can be derived as
\[(\theta_{\text{ref}} - \theta_r)G_\theta(s) + \omega_{\text{ref}} + (\omega_{\text{ref}} - \omega_r) = \omega_{\text{ref}}.\]

Moreover, \(\omega_{\text{ref}} = \theta_{\text{ref}}s\), \(\omega_r = \theta_r s\), then we have
\[(\theta_{\text{ref}} - \theta_r)G_\theta(s) + 2\theta_{\text{ref}}s - \theta_r s = \frac{\theta_r s}{G_1(s)}.\]

Rearranging it we get
\[
\frac{\theta_r}{\theta_{\text{ref}}} = \frac{G_\theta(s) + 2s}{G_\theta(s) + s + \frac{s}{G_1(s)}}.
\]

For rated condition with \(G_\theta(s) = K_\theta = 4\),
\[
\frac{\theta_r}{\theta_{\text{ref}}} = \frac{1.70e3s^4 + 1.71e5s^3 + 2.38e8s^2 + 1.095e10s + 2.09e10}{s^5 + 200s^4 + 1.54e5s^4 + 1.32e7s^3 + 5.58e8s^2 + 1.09e10s + 2.09e10}
\]

Eigenvalues of the transfer function are listed below.

Table 3.6 Eigenvalues of closed loop speed, closed loop position control

<table>
<thead>
<tr>
<th>Eigenvalues</th>
<th>-54.6±j371</th>
<th>-24.7±j34.4</th>
<th>-38.9</th>
<th>-2.14</th>
</tr>
</thead>
</table>

It is indicated from the eigenvalues that all the control methods are stable. There is one more eigenvalue occurred for closed loop position control than open loop position controls. This is because one more integration block is introduced between the speed and the position.

A bode diagram of the open loop speed closed loop position control (OSCP) is given in Fig. 3.23. The actual working frequency, in terms of position, is very low for the stage system. The frequency response analysis shows the whole system has unity gain and very small lag when working in the low frequency range.
In this chapter, two acceleration-speed-position (ASP) profiles have been discussed followed by speed loop and position loop design. All those controls are straightforward and time-invariant. The time-variant characteristics of the induction machine under scalar control narrows the choice of control parameters. And it is hard to find a linear and time-invariant controller optimized for all the operating states of the machine. However, thanks to the feed-forward cascade structure, the algorithm can still achieve accurate and fast position control which fulfills the demands aroused by safe operation and riding comfortability.
Chapter 4  Implementation of the Variable-Speed Drive

4.1  Hardware Design

A good control algorithm is only the first step towards a satisfactory system. A lot of efforts have been made building up an experimental system to verify the conceptual design. From this chapter, emphasis will be given to the implementation of a prototype of the variable-speed drive from which the final implementation will be developed.

4.1.1  Hardware Overview

Fig. 4.1 depicts the basic components of the whole drive system for the stage. The drive system includes a variable-speed drive (VSD), an induction machine, and a transmission system coupled to both the motor and the stage. Among all the components, the variable-speed drive is the heart of the system. All the motion controls as well as the motor control will be realized via the VSD.

The variable-speed drive employs the commonly used AC-DC-AC topology. It converts the 3-phase, 60 Hz input into dc voltage source, which is then inverted into a 3-phase variable frequency and variable voltage output to drive the motor. In the prototype, the inverter uses six discrete IGBT power modules with built-in freewheel diode. There are also highly integrated IGBT power modules available in the market, which include...
the three-phase rectifier, the three-phase inverter, the brake chopper and a temperature sensor all into one package. Due to the high reliability, those all-in-one modules are recommended in the final implementation.

The IGBT power modules are gated by a DSP-based controller. According to the operation command, the DSP generates sinusoidally modulated PWM outputs to the driver IC. Opto-couplers provide necessary signal isolation between the DSP and driver IC. Interface circuitry between the DSP and outside sensors generates voltage shift and filters noise out from signals feeding to the DSP. In case a fault happens, the fault signal is fed back to the PDPINT channel of the DSP.

The converter-motor combination can accommodate four-quadrant operations such that during the generating mode, the induction machine converts the kinetic energy to electrical energy. In that case the current and power reverses the direction and tries to flow back to the power supply side, which will be blocked by the diodes in the rectifier and invoke a voltage rise at the dc bus. Thus, a resistor, called \( R_d \), is used to dissipate the energy during the generating mode. An extra IGBT switch in series with \( R_d \), turns on whenever the dc bus voltage exceeds a safe operation value, thus preventing overcharging.

An encoder is a position sensor coupled to the shaft of rotational objects. Speed information can be derived from the differentiation of the position signals within some time period. In the DSP, there is an interface called QEP, which is specifically designed to process the signals from the encoders.

The next discussion will be focused on each component mentioned above.

### 4.1.2 IGBT and its Drive

Fig. 4.2 illustrates a basic IGBT gate drive circuit, which converts logic level control signals into appropriate voltage and current that can drive the IGBT power module reliably and efficiently [7]. The conversion is performed by a pair of bipolar transistors alternately connecting the IGBT's gate to the appropriate on (Von) and off (Voff) voltages. The gate resistor is selected to generate a proper peak current charging or discharging the IGBT's gate. The optocoupler provides isolation between the high power component and control signal to avoid potential damage to the digital controller.
The IGBT gate drive circuits are subjected to high common mode $dv/dt$ noise produced by the fast switching, high voltage and high current IGBT power modules. To maintain the immunity to the high $dv/dt$ noise is critical for the drive circuit to function normally in an offensive environment.

4.1.2.1 Maintaining $dv/dt$ Noise Immunity

High common mode $dv/dt$ induced by the turn-on of the positive IGBTs in a two-level bridge leg is something that users often encounter. When the collector of the positive IGBT is turned on, a current is induced due to the device's collector-gate (i.e. reverse transfer) capacitance, shown in Fig. 4.3 as $C_{RES}$. The collector-to-gate current develops a voltage across the series gate resistor as well as the stray inductance in the gate circuitry. If this annoying voltage exceeds the threshold voltage of the IGBT, the off-state IGBT will be spuriously turned on, and thus, a cross-conduction will happen.

Adding an extra resistor between the gate and emitter of the IGBT device could, to some extent, cancel the offending collector-to-gate current. The principle here is that a
RC low-pass filter, consisting of the collector-gate capacitance and the gate-emitter resistor, as shown in Fig. 4.3, provides a by-pass for the high-frequency noise.

In addition, the high common-mode dv/dt noise might cause breakdown, i.e. a transient malfunction of any isolating interface for the gate drive signals. After the anti-crosscondution delay, when the complementary negative device turns on, the emitter of the positive device makes a transition to the negative bus potential with a very high negative dv/dt, which can often exceed the common-mode dv/dt noise immunity level of the optocouplers used. It is important for the positive device, which must remain "off" during this transient.

![Fig. 4.4 Noise shielding of opto-couplers](image)

Optocouplers with good immunity to high common-mode dv/dt noise could be used in the drive circuit. The immunity is normally achieved by adding shields between the "primary" and "secondary" side of the opto-coupler. Another possible solution is to lower the induced dv/dt noise by increasing the "turn-on" gate resistor on the device drivers. However this will result in a slowing down turn-on process and consequently higher "turn-on" switching loss. In addition, given the switching frequency, the turn-on process has to be fast enough compared to the whole "on" period so as to keep the duty cycle fidelity.

As an alternative to optocouplers, fiber optic cables can be used to completely eliminate the problem and to provide very high electrical isolation and creepage distance [8]. Now optic fibers are almost universally used in high-power voltage-fed IGBT converters. A side-effect benefit from using optic-fibers is that the digital controller and the IGBT driver/IGBT module could be physically separated, which makes the design and layout much more flexible.
Apart from the control-signal isolation barrier, attention must also be given to reducing $dv/dt$ capacitance coupling to or from the positive IGBT floating gate-drive power supply. In most high power applications it is necessary to use power supplies that are isolated from the input side such that they can be floating with respect to the ground because the potential of the emitter of the high side IGBT changes when the low side IGBT is switched. Although isolated power supplies are not definitely necessary for low side IGBT drives, they are recommended in order to avoid noise caused by noise voltage induced in stray inductance of the negative dc bus [7].

Normally a common-mode unwanted capacitance is intrinsic between the transformer windings of the power supplies. Appropriate electrostatic shielding of the power supply transformer windings from each other is crucial. In high power system, the positive drivers are always supplied from a shielded transformer dedicated to each IGBT. In addition, an inductor with high common-mode rejection is placed in the input side of the power supplies in order to prevent coupled high $dv/dt$ noise, shown in Fig. 4.7.
A substantial negative bias is used for IGBT drive, which provides additional \( dv/dt \) immunity and reduces turn-off losses. The additional margin to absorb "real" collector-gate capacitance coupled reverse transfer charge during high \( dv/dt \), with respect to the gate-emitter "turn-on" threshold voltage, is a significant reliability improvement, particularly when switching peak (fault) current, coincident with a high dc-bus voltage.

\[
\begin{align*}
V_{\text{ge}} &\quad V_{\text{ge}} \\
V_{\text{th}} &\quad V_{\text{th}} \\
(a) &\quad (b)
\end{align*}
\]

Fig. 4.8 Additional \( dv/dt \) immunity of negative bias turn-off voltage

4.1.2.2 IGBT Drive Schematics and Component Selection

As a result of the fast developing IC technologies, IGBT drivers are not confined to just provide drive function. Some protection and auxiliary functions are integrated into the driver IC to provide an all-in-one product. The gate drive used in the prototype, HCPL-3120, is a high-current output IGBT gate drive with built-in opto-coupler. Its main parameters are given in Table 4.1.

<table>
<thead>
<tr>
<th>Feature</th>
<th>Specification</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak output current</td>
<td>2.0 - 2.5 A</td>
<td></td>
</tr>
<tr>
<td>Common-mode rejection (CMR)</td>
<td>15 kV/\mu s</td>
<td>( V_{\text{cm}} = 1.5 \text{ kV} )</td>
</tr>
<tr>
<td>Input voltage Vcc</td>
<td>15 – 30 V</td>
<td></td>
</tr>
<tr>
<td>Under Voltage Lock-Out Protection (UVLO)</td>
<td>9.5 – 12 V</td>
<td>Hysteresis</td>
</tr>
<tr>
<td>Maximum switch frequency</td>
<td>2 MHz</td>
<td></td>
</tr>
<tr>
<td>Isolation</td>
<td>630 V peak</td>
<td></td>
</tr>
</tbody>
</table>
HCPL-3120 contains an under-voltage lockout (UVLO) feature that is designed to protect the IGBT under low voltage condition. The output voltage of HCPL-3120 switches the output off when it detects a low voltage from the power supply and turns on at a higher threshold. Other components used in the gate drive are listed in Table 4.2.

<table>
<thead>
<tr>
<th>Reference</th>
<th>Name</th>
<th>Type</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>U1</td>
<td>Dc/dc power supply</td>
<td>VASD1-S5-D15</td>
<td>5V/±15V</td>
</tr>
<tr>
<td>U2</td>
<td>Gate drive IC</td>
<td>HCPL3120</td>
<td>With built-in opto-coupler</td>
</tr>
<tr>
<td>R1</td>
<td>Resistor</td>
<td>180 Ω</td>
<td></td>
</tr>
<tr>
<td>R2</td>
<td>Gate resistor</td>
<td>51Ω</td>
<td></td>
</tr>
<tr>
<td>R3</td>
<td>Resistor</td>
<td>1.5 kΩ</td>
<td></td>
</tr>
<tr>
<td>R4,R5</td>
<td>Resistor</td>
<td>5.5 kΩ</td>
<td></td>
</tr>
<tr>
<td>C1,C2,C3</td>
<td>Capacitor</td>
<td>10 μF</td>
<td>electrolytic</td>
</tr>
</tbody>
</table>

The per-phase IGBT drive schematic is shown in Fig. 4.9.
A dc-dc converter provides the isolated ±15V power to the IGBT drive. The converter can provide 1kV dc voltage isolation across its input and output that is high enough in this application. A resistor is connected to the output of the converter which needs a minimum of 10% loading to maintain a reliable and fully-performed output.

There are several guidelines to choose a proper value for the gate resistor $R_g$, one of which is driven by the peak current of the IGBT driver, illustrated below.

![IGBT gate drive schematic](image)

**Table 4.3 Gate resistance sizing driven by peak output current of the IGBT drive**

<table>
<thead>
<tr>
<th>Vcc</th>
<th>$I_{out_pk}$</th>
<th>$R_g = \frac{Vcc}{I_{out_pk}}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gate drive voltage</td>
<td>Peak output current of the IGBT drive</td>
<td>Gate resistance</td>
</tr>
<tr>
<td>15 V</td>
<td>2 A</td>
<td>7.5 $\Omega$</td>
</tr>
</tbody>
</table>
The above calculation is somewhat coarse. Actually by properly sizing the gate resistors the switching speed of the output IGBT can be controlled [9]. Some basic rules are given below for sizing the gate resistors to obtain desired switching time. The switching time $t_{sw}$ is defined as the time spent to reach the end of the plateau voltage, as shown in Fig. 4.10.

![Fig. 4.10 IGBT turn-on sequence](image)

Here $Q_{gc}$ and $Q_{ge}$ indicate the gate to collector and gate to emitter charge respectively. Table 4.4 shows the calculation process to size the turn-on gate resistor driven by $t_{sw}$ constraint.

![Fig. 4.11 Current path when IGBT turns on](image)
Table 4.4 Turn-on gate resistor sizing driven by $t_{sw}$ constraint

<table>
<thead>
<tr>
<th>Description</th>
<th>IRG4PC50UD</th>
</tr>
</thead>
<tbody>
<tr>
<td>$Q_{ge}$</td>
<td>Gate Emitter charge (turn-on)</td>
</tr>
<tr>
<td>$Q_{gc}$</td>
<td>Gate Collector Charge (turn-on)</td>
</tr>
<tr>
<td>$t_{sw}$</td>
<td>Switching Time</td>
</tr>
</tbody>
</table>

\[
I_{av} = \frac{Q_{gc} + Q_{ge}}{t_{sw}}
\]

Average Charging Current | 172 mA |

<table>
<thead>
<tr>
<th>$V_{ge}^*$</th>
<th>Gate Plateau Voltage</th>
<th>6 V</th>
</tr>
</thead>
</table>

\[
R_{TOT} = \frac{V_{CC} - V_{ge}^*}{I_{av}}
\]

Equivalent Output Resistance of the Gate Driver | 52 Ω |

<table>
<thead>
<tr>
<th>$R_{DRp}$</th>
<th>Driver Equivalent on-resistance</th>
<th>10 – 15 Ω</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_{Gon} = R_{TOT} - R_{DRp}$</td>
<td>Gate On-resistance</td>
<td>37 - 42 Ω</td>
</tr>
</tbody>
</table>

Turn-on gate resistor can also be sized to control output slope $dV_{out}/dt$. Although the output voltage has a non-linear behavior, the maximum output slope can be approximated by \( \frac{dV_{out}}{dt} = \frac{I_{av}}{C_{RESoff}} \) [9]. The calculation of this kind of constraint is given in Table 4.5.
Table 4.5 Turn-on gate resistor sizing driven by $dV_{out}/dt$ constraint

<table>
<thead>
<tr>
<th>Description</th>
<th>IRG4PC50UD</th>
</tr>
</thead>
<tbody>
<tr>
<td>$dV_{out}/dt$</td>
<td>Output Voltage Slope</td>
</tr>
<tr>
<td>$C_{RESoff}$ (off-state)</td>
<td>Reverse Transfer Capacitance</td>
</tr>
<tr>
<td>$I_{avg} = \frac{dV_{out}}{dt} C_{RESoff}$</td>
<td>Average Charging Current</td>
</tr>
<tr>
<td>$V_{ge}$</td>
<td>Gate Plateau Voltage</td>
</tr>
<tr>
<td>$R_{TOT} = \frac{V_{CC} - V_{ge}^*}{I_{avg}}$</td>
<td>Equivalent Output Resistance of the Gate Driver</td>
</tr>
<tr>
<td>$R_{DRp}$</td>
<td>Driver Equivalent on-resistance</td>
</tr>
<tr>
<td>$R_{Gen} = R_{TOT} - R_{DRp}$</td>
<td>Gate On-resistance</td>
</tr>
</tbody>
</table>

Combining the calculation results for the two constraints, the gate resistance for turn-on should be greater than 40 ohms.

The worst condition in calculating the turn-off resistor is when the collector of the IGBT in the off state is forced to commutate by the turn-on of the companion IGBT [9]. In that case, a parasitic current through $C_{RESoff}$ will be induced by the high $dv/dt$ of the output node. If the voltage drop at the gate exceeds the threshold voltage of the IGBT, the

Fig. 4.12 Current path of the low-side IGBT drive when high-side turns on
device may be turned on by itself, which will cause cross conduction for the whole leg. If no negative bias voltage is used, condition

\[ V_{th} > V_{ge} = (R_{Goff} + R_{DRn}) \cdot C_{RESoff} \cdot \frac{dV_{out}}{dt} \]  

(4.1)

must be verified to avoid spurious turn-on. Rearrange (4.1) we get

\[ R_{Goff} < \frac{V_{th}}{C_{RESoff} \cdot \frac{dV}{dt} - R_{DRn}} \]  

(4.2)

Fig. 4.12 shows the current induced by the high \( \frac{dv}{dt} \) of the output node, where \( C_{IES} \) is the input capacitance, and \( C_{RES} \) is the reverse transfer capacitance. An example of calculating the turn-off gate resistor is given in Table 4.6.

<table>
<thead>
<tr>
<th>Description</th>
<th>IRG4PC50UD</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \frac{dV_{out}}{dt} )</td>
<td>Output Voltage Slope</td>
</tr>
<tr>
<td>( C_{RESoff} )</td>
<td>Reverse Transfer Capacitance (off-state)</td>
</tr>
<tr>
<td>( V_{th} )</td>
<td>Gate Threshold Voltage</td>
</tr>
<tr>
<td>( R_{TOT} )</td>
<td>Equivalent Output Resistance of the Gate Driver</td>
</tr>
<tr>
<td>( R_{DRn} )</td>
<td>Driver Equivalent off-resistance</td>
</tr>
<tr>
<td>( R_{Goff} = R_{TOT} - R_{DRn} )</td>
<td>Gate Off-resistance</td>
</tr>
</tbody>
</table>

From the above estimation, we find that the required resistance for turning-off contradicts the turn-on gate resistance. In order to make both turn-on and turn-off safe, the gate circuit, such as the one shown in Fig. 4.13, may be employed. The diode will block the “on” current path from flowing through the “off” resistor, while the “off” current flows through both resistors. Normally the “off” path is short-circuited to provide as quick turning-off process as possible.
If we consider the extreme case that there is a very fast step change on IGBT collector, the gate-emitter voltage can be approximated with the capacitor divider:

\[ V_{ge} = V_{ce} \frac{C_{RESoff}}{C_{RESoff} + C_{IES}} \] (4.3)

which is driven only by IGBT characteristics. Table 4.7 gives the verification for the IGBT used in the prototype.

Table 4.7 Gate voltage spike induced by high dv/dt

<table>
<thead>
<tr>
<th>Description</th>
<th>IRG4PC50UD</th>
</tr>
</thead>
<tbody>
<tr>
<td>( C_{RESoff} ) Reverse Transfer Capacitance (off-state)</td>
<td>52 pF</td>
</tr>
<tr>
<td>( C_{IES} ) Input Capacitance</td>
<td>4000 pF</td>
</tr>
<tr>
<td>( V_{ce} ) Collector Voltage</td>
<td>300 V</td>
</tr>
<tr>
<td>( V_{ge} ) Gate Voltage</td>
<td>3.9 V</td>
</tr>
<tr>
<td>( V_{th} ) Gate Threshold Voltage</td>
<td>3 - 6 V</td>
</tr>
</tbody>
</table>

It is clear that the gate voltage will be raised to a critical level that it might trigger the conduction of the IGBT at the wrong moment. Apart from the method to introduce two separate gate resistors, another way to avoid the spurious turn-on is to use negative bias voltage for the off state, which is also the solution used in building the prototype board. For a negative bias voltage of -15V, the actual gate voltage under the extreme condition will be -11V as maximum during the "off" state, which is quite far from the threshold voltage of the IGBT. The additional margin brought by negative bias voltage is really a significant reliability improvement.
The above-described methods for sizing gate resistors are intended to approximate the turn-on and turn-off phenomena of power IGBTs. More accurate sizing may rely on more precise IGBT modeling and parasitic components dependant on the layout and connection of the circuit.

4.1.3 Temperature Detection Subsystem

Thermal issue and temperature control is a problem affecting any semiconductors. The sensitivity exhibited by the NTC thermistor makes it advantageous to be used for low cost, precision temperature measurement and control in the industry. A thermistor is an electronic component that exhibits a large change in resistance with a change in its body temperature. NTC thermistors have a negative temperature-resistance characteristic that the resistance will drop as temperature rises. When NTC thermistor sensor is integrated into the power module, it can facilitate temperature detection and over-temperature protection.

But the non-linearity of the thermistor resistance-temperature characteristic sets a practical limit on the possible temperature range over which a single thermistor can be operated in a measurement circuit. Therefore an interface circuitry is employed to generate DSP-friendly signal that can be directly sampled by the DSP. The interface circuit is given in Fig. 4.14

![Fig. 4.14 NTC interface to the DSP](image)

Fig. 4.15 illustrates the temperature-resistance characteristic of the NTC sensor used in the integrated power module as well as the voltage-temperature characteristic seen from the DSP. The transformed voltage signal can easily be read by the DSP, from 0
to 2.5 V. The linearity has been improved to be suitable for measurement. The input terminal into the DSP can also be clamp to a 3.3V voltage source via a Zener diode to avoid damaging ADC module by spike voltage.

![Characteristic of NTC Thermistor](image1)

**Fig. 4.15 NTC characteristic and sampled voltage**

As the linearity of the temperature-voltage characteristic is still not ideal, a lookup table is introduced to interpret the temperature from the input voltage.

Channel ADCIN1 in the DSP is assigned to accept $v_{in}$ and RESULT1 is the variable associated with $v_{in}$ in the software. The lookup table has 64 rows of records such that the interpretation from voltage to temperature can provide enough resolution for control algorithm to know the temperature. The minimum temperature interval is 5°C in the program. Therefore only the six MSBs (most significant bits), i.e. bit 15 to bit 10, of $v_{in}$ are effective and RESULT1 is right-shifted by 10 bits to discard bit 0 to bit 9.

The actual $v_{in}$ is in the range from 0.4 – 2.35 V. Voltages beyond that range represent temperatures under 0°C or over 150°C, which are as well beyond the possible operation temperature range. In the lookup table, all input voltages between 0 to 0.4V are translated to 150°C, and voltages between 2.35V to 3.3V are translated to 0°C for simplicity.

The IGBT's body temperature should be maintained between 0°C to 125°C to assure device safety and undegraded performance. The program sends a temperature
alarm when temperature rises to 100°C. The output pulses will be shut down when temperature is over 125°C.

4.1.4 DC Bus Bulk Capacitor

Electrolytic capacitors are used to smooth the dc bus voltage. Its capacitance is an inverse function of the allowed ripple voltage $\Delta V$ and can be derived as

$$C_{\text{min}} = \frac{2P_{\text{in}}}{(V_{\text{max}}^2 - V_{\text{min}}^2)f_{\text{rect}}}$$

where $P_{\text{in}}$ is the load power in watts, $f_{\text{rect}}$ is the ripple frequency, $V_{\text{max}}$ is the maximum dc voltage and $V_{\text{min}}$ is the minimum dc voltage [10].

![Figure 4.16 DC bus ripple](image)

With a 3-phase 208V ac input, $V_{\text{max}} = \sqrt{2}V_{LL} = 281V$. Assume $V_{\text{min}} = 95\%V_{\text{max}} = 267V$, then

$$C_{\text{min}} = \frac{2P_{\text{in}}}{(V_{\text{max}}^2 - V_{\text{min}}^2)f_{\text{rect}}} = \frac{2 \times 2235}{(281^2 - 267^2) \times 360} = 1618 \mu F$$

$t_c$, the charging time, can be calculated as
\[
\cos^{-1}\left( \frac{V_{\text{min}}}{V_{\text{max}}} \right) = \frac{\cos^{-1}(0.95)}{2\pi} = 842\mu S
\]  

(4.5)

and discharging time \( t_{DC} \) is

\[
t_{DC} = \frac{1}{f_{\text{rect}}} - t_e = \frac{1}{360} - 842\mu = 1.94\text{mS}
\]  

(4.6)

The average charging current is given by

\[
I_C = C \frac{\Delta V}{t_e} = C \frac{V_{\text{max}} - V_{\text{min}}}{t_e} = 1618\mu \frac{281 - 267}{842\mu} = 27\text{A}
\]  

(4.7)

According to the calculation, at least a 1600 \( \mu \)F capacitor should be employed to maintain the dc bus ripple within 5% or less. The capacitor should also can stand 27A charging current.

4.1.5 Energy Dissipation Subsystem

During the braking period, the kinetic energy of the stage system will be reverted to electric energy by the induction machine, which is shown in Fig. 4.17. As the converter, consisting of only diodes, can not send the regenerated energy back to the grid, a dissipation system must be adopted to absorb the excessive energy, otherwise dc bus voltage might be charged to a very high and dangerous level. A braking branch is then developed to dissipate the extra current. The braking branch includes a voltage-controlled IGBT and a resistor connected in series to the dc bus. The IGBT switch will be closed and connect the braking resistor to the dc bus when the dc voltage exceeds a threshold. The control circuit disconnects the braking resistor when the dc voltage drops back to normal level.
Fig. 4.17 Current path for (a) Operation mode of motoring (b) Operation mode of generating

The braking resistance can be derived from the fact that the braking resistor should be able to dissipate all the regenerated power. Recalling the differential equation (2.1) describing the rotational motion, we have

\[ T_M - T_L = J \frac{d\omega}{dt} \]

If the deceleration rate is defined as \( \alpha = \frac{d\omega}{dt} \), \( T_M \) is the braking torque of the machine, \( T_L \) is equal to the friction torque \( T_f \), and \( J = J_{Me} \), where \( J_{Me} \) is the equivalent moment of inertia referred to the motor side, rearranging (4.8) we have

\[ T_M = J_{Me} \frac{d\omega}{dt} - T_f \]  

(4.9)

Note all the quantities in (4.9) are expressed in their magnitudes. The regenerated power is

\[ P_g = T_M \omega_r = (J_{Me} \frac{d\omega}{dt} - T_f) \omega_r = (J_{Me} \alpha - T_f) \omega_r \]  

(4.10)
The power dissipated in the braking resistor $P_d$ can be expressed as

$$P_d = \frac{V_{dc}^2}{R_b}$$  \hspace{1cm} (4.11)

The inequation $P_d \geq P_g$ has to be satisfied to maintain a constant dc bus voltage. Combining (4.10) and (4.11), we get

$$R_b \leq \frac{V_{dc}^2}{(J_{Me}\omega - T_f)}$$  \hspace{1cm} (4.12)

The right part of (4.12) reaches its minimum value when the machine is running at its maximum speed, which means the braking resistor has to be designed according to the worst case when the stage just begins to decelerate and the machine has its highest regenerated power at that moment. A detailed calculation of the braking resistor used in this application is given below in Table 4.8.

<table>
<thead>
<tr>
<th>Variable</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{dc}$</td>
<td>280 V</td>
<td></td>
</tr>
<tr>
<td>$J_{Me}$</td>
<td>0.16 kg $\cdot$ m$^2$</td>
<td>Moment of inertia of the whole system referred to the motor side</td>
</tr>
<tr>
<td>$\alpha$</td>
<td>75.4 rad/s$^2$</td>
<td>The deceleration for a 1800 rpm-running machine to stop within 2.5 seconds</td>
</tr>
<tr>
<td>$T_f$</td>
<td>0 Nm</td>
<td>Conservatively assuming no friction torque</td>
</tr>
<tr>
<td>$\omega_r$</td>
<td>188.5 rad/s</td>
<td>Assuming rotor speed equal to the maximum synchronous speed</td>
</tr>
<tr>
<td>$R_b$</td>
<td>34 $\Omega$</td>
<td></td>
</tr>
</tbody>
</table>

The voltage detection circuitry is plotted in Fig. 4.18. The attenuated dc bus voltage is sent to a voltage follower, and then compared with a reference voltage. The
voltage comparator has a hysteresis pickup/dropout characteristic, which can eliminate flipping of the braking branch.

\[ V_{\text{pickup}} = V_{\text{ref}} (1 + \frac{R_1}{R_f}) \]  

(4.14)

Dropout voltage is obtained by

\[ V_{\text{dropout}} = V_{\text{ref}} (1 + \frac{R_1}{R_f}) - \frac{R_1}{R_f} V_o \]  

(4.16)

\[ \frac{V_o - V_{\text{ref}}}{R_f} = \frac{V_{\text{ref}} - V_{\text{dropout}}}{R_1} \]  

(4.15)

According to Fig. 4.18, the pickup voltage is calculated by

\[ \frac{V_{\text{pickup}} - V_{\text{ref}}}{R_1} = \frac{V_{\text{ref}}}{R_f} \]  

(4.13)
Assuming $V_{dropout} = \frac{285}{100} V$, $V_{pickup} = \frac{295}{100} V$, $V_o = 4.5V$, we get

$$\frac{R_f}{R_1} = 45, \quad V_{ref} = 2.9V$$

Reference voltage is generated by a shunt regulator (TL431).

### 4.1.6 Encoders

In order to obtain position and machine speed, a rotary encoder is probably the most frequently used transducer. There are two types of encoders, absolute encoders and incremental encoders. An absolute encoder has multiple channels, each of which reads one binary digit. A typical 10-line encoder will generate a 10-bit binary number and one revolution will be divided into $2^{10}$ (1024) positions, each giving a unique code. By interpreting the output code, a control system can get the absolute position information associated with it.

![Fig. 4.20 An incremental encoder](image)

In an incremental encoder, pulses are generated for every increment of travel. By counting these pulses, the traveled angle or distance is obtained. Normally there are two channels of pulses, which are quadrature to each other, generated by the encoders. By comparing which channel is leading, it is quite easy to detect the rotating direction of the motor. In addition to that, a start marker is usually provided on a third channel as an ‘index’, which can be used to indicate a zero or home position on the encoder.

It is quite easy to develop speed measurement with either type of encoders. The output from an encoder is a series of pulses corresponding to the angle it travels. Speed is
the derivative of angle. In digitized form, the average speed in the sampling interval is expressed as

$$\omega(t) = \frac{1}{T} [\theta(t) - \theta(t-1)]$$

where $T$ is the sampling interval.

It is worth mentioning that with a given sampling interval $T$ the resolution of a discrete speed measurement is reduced at low speed range because fewer pulses are counted within one sampling interval. The worst case for speed estimation happens when no more pulses come during one sampling interval $T$, and calculated speed is zero. Given a constant-time window, the number of observed pulses is counted and the speed is then approximated as

$$\omega = \frac{d\theta}{dt} \approx \frac{\Delta \theta}{T_s} \approx \frac{2\pi \cdot \Delta N}{2^N \cdot T_s} [rad \cdot s^{-1}] \text{ or } f = \frac{60 \cdot \Delta N}{2^N \cdot T_s} [rpm]$$

where $N$ is the output bit number of the encoder, $\Delta N$ is the number of observed pulses, $T_s$ is the sampling time [11]. The resolution of speed measurement $\Delta \omega$ is given as

$$\Delta \omega = \frac{2\pi}{2^N \cdot T_s} [rad \cdot s^{-1}] \text{ or } \Delta f = \frac{60}{2^N \cdot T_s} [rpm]$$

The above equation indicates that the quantization resolution depends only on the resolution of the encoder and the sampling time. The resolution is independent from the motor speed. In this project, a 12-bit encoder is used and sampling time is set to 1/4000s, so the speed resolution is

$$\Delta f = \frac{60}{2^{12} \cdot (1/4000)} \approx 60 [rpm]$$

which is equal to the resolution of the speed lookup table ASP_ST.

The current stage drive system uses a wire rope to connect the motor and the stage, which will be the approach used in the new drive system as well. However, there is no such a transmission system in the experimental setup and the encoder is directly coupled to the shaft of the motor. In the real system, this configuration of the encoder may be problematic because measurement error of position, caused by slippage, will be accumulated from the motor shaft encoder. Even if there is no slippage, the transmission ratio used for calculation, i.e. the nominal ratio, will never be exactly equal to the actual
transmission ratio. The difference will be accumulated as well to ruin the position estimation.

If only one encoder would be used, it could be coupled to the shaft of the stage. Due to the transmission effect, the speed measured at the stage shaft is much slower than the speed measured from the motor's shaft. The advantage of this one-encoder configuration is that the speed estimation will not be influenced by slippage. In addition, the implementation is simple. The disadvantage is that an encoder with higher output bit is necessary. From simulation results, the system will be stable only if the sampling period of speed is less than 0.05s. Assuming that the transmission ratio is 600:1 from the motor shaft to the stage shaft and the speed resolution is 60 rpm, if we rearrange (4.14), then we have

\[ 2^N \geq \frac{60}{\Delta f \cdot T_s} \cdot \lambda = \frac{60}{60 \cdot 0.05} \cdot 600 = 12000 \, \text{pps}, \]

So at least a 14-bit encoder (16384 pulses per revolution) is needed for this system.

If we want to use the same 12-bit encoder in the stage drive system, a second encoder could be installed and coupled to the motor shaft. In such a case, the stage encoder will provide the position measurement while the motor shaft encoder will provide the speed signal to the control system. However, this two-encoder configuration will definitely make both hardware and software much more complicated than the one-encoder configuration. In addition, extra measures such as a low-pass filter for the speed estimation may be necessary to eliminate the invalid reading caused by slippage.

Apart from the above two methods, there is a third option available, which is to use open loop speed closed loop position control (OSCP). As there is no speed loop, only one encoder is required. This encoder can be coupled to the shaft of the stage. Different from the first solution, this encoder is assigned to provide only position information. As we know that the changing rate of position is slower than speed, high-bit output capability is therefore not necessarily required for this encoder.

From simulation, as well as the experimental results that will be given in the next chapter, open loop speed closed loop position control (OSCP) can achieve as same the position accuracy as obtained from closed loop speed closed loop position control.
(CSCP). From this point of view, open speed closed position control (OSCP) gets ahead of the rest methods.

An interface circuitry is added between the encoder and DSP QEP pins. Only one channel is plotted in Fig. 4.21. A voltage follower boosts the output capability for the encoder and an opto-coupler provides galvanic isolation.

![Fig. 4.21 Interface circuitry for one channel between the DSP and the encoder](image)

As 3.3V is the standard voltage for I/O of the DSP, a simple 3.3V power supply is built by using a TL431, shown in Fig. 4.22.

![Fig. 4.22 3.3V power supply using programmable shunt regulator diode](image)

4.2 **Software Design**

The code for the drive system is written in assembly language and has been tested successfully on a TMS320LF2407 DSP, which is mounted on a LF2407 Evaluation Module (EVM) from Spectrum Digital.
Some flowcharts summarize the structure of the code and are attached in Appendix G. The complete listing of the assembly code and auxiliary files are attached as well in the appendix.

4.2.1 Pulse Width Modulation with THI

The method used to produce the sine modulated PWM signal is based on look-up tables. A look-up table STBL associates PWM frequencies with different steps. A step value is added to the counter every time a new value from the sine table is to be loaded. By changing the value of the step, the program can accurately control the frequency of the sine wave [12].

From the calculation in 4.1.4, we notice that the average dc bus voltage is around 274 V. Using a sinusoidal PWM, the maximum line-to-line voltage is

\[ V_{LL} = \sqrt{3} \frac{V_{dc}}{2\sqrt{2}} = 168V \]

which is only 80% of the rated voltage of the induction machine. In order to increase the inverted voltage, third harmonic injection has been incorporated into the modulation. With third harmonic injection (THI), the maximum output line-to-line voltage of the inverter reaches

\[ V_{LL} = \frac{V_{dc}}{\sqrt{2}} = 194V \]

which has a 15% increase over the sinusoidal PWM.

4.2.2 Event Manager Setup

The program sets up the Event Manager A (EVA) to generate six symmetrical sinusoidal PWM waveforms on pin PWM1 to PWM6. Each Event Manager module has a Quadrature Encoder Pulse (QEP) circuit, which normally interfaces with optical encoder to receive position and speed information. The QEP in EVA is enabled in the software and pins QEP1 and QEP2 are configured to receive two channels of pulses from the encoder. Timer1, acting as a base frequency for the program, is configured to have a frequency of 20 kHz, which is also the switching frequency for all six IGBTs.
4.2.3 Speed Estimation and Speed/Position Recording

The information retrieved from the encoder is in the form of position. In order to obtain the speed signal, the computation

\[ \text{Speed} = K \left[ \text{T2CNT}(n) - \text{T2CNT}(n-1) \right] / \text{T} \]  

(4.20)
is required, where T2CNT is the register associated with encoder readings, T is the sampling interval and K is a coefficient to transfer the speed to a hertz format. It is found that if the sampling frequency is set to be equal to the number of pulses per revolution of the encoder, the division can be eliminated. Assume \( f_r \) is the mechanical speed of the motor in hertz, \( N_{ppr} \) is the number of pulses per revolution of the encoder, and \( f_{spl} = \frac{1}{\Delta t_s} \) is the sampling frequency. For each sampling period, the number of counted pulses \( N_{pps} = f_r \cdot N_{ppr} \cdot \Delta t_s = \frac{f_r \cdot N_{ppr}}{f_{spl}} \). If \( N_{pps} = f_{spl} \), we have \( N_{pps} = f_r \), which means the number of counted pulses is the actual mechanical speed. In the software, the refreshing frequency of speed estimation is set to 4,000, which is quite approximated to the number of pulses per revolution of the encoder.

The register T2CNT storing the pulses overflows when it reaches the maximum count in an up-counting mode, which is associated with a positive rotating direction of the motor. Likewise the register underflows when it counts down to zero and winds back to its maximum value.

Fig. 4.23 Two counting modes for T2CNT (a) Up-counting mode (b) Down-counting mode
A special program is designed to correct the measurement when overflow or underflow occurs. It is assumed the motor can only be rotating towards one direction in one shot of the operation. If the rotating direction is positive and the QEP is in the up-counting mode, once $T2CNT(n) - T2CNT(n-1) < 0$, which represents an overflow, the speed reading will be recalculated as

$$\text{Speed} = T2CNT(n) - T2CNT(n-1) + N_{ppr}$$

Similarly, the program negatives the speed reading when the motor is assigned a negative rotating direction. It is worth mentioning that jerks in the T2CNT reading caused by mechanical oscillation may happen, which will ruin the reading and make the control algorithm invalid. A low pass filter is then added to overcome the bad reading. If the possible maximum speed for the motor is $f_{r\_max}$, because $N_{ppr} = f_{spl}$, then $N_{pps\_max} \leq f_{r\_max}$. Consequently, any reading of $N_{pps}$ greater than $N_{pps\_max}$ will be discarded.

The measured variables are stored for future analysis. In the software variables are stored every other 0.1 second. Two arrays SPD and POS are claimed to record real-time speed and position information. Some of the experimental results saved in SPD and POS will be redraw in Chapter 5.

### 4.2.4 Deadtime Setting

It is necessary to include a blank time, called deadtime, to avoid cross conduction between the high side and low side devices in one leg because any real power electronic devices do not turn on or off instantaneously. In the EVA of the LF2407 DSP, a programmable dead-band unit DBTCONA is built-in to add a deadtime into the PWM signals. It has been found that the deadtime causes a reduction in the fundamental component of the output voltage and introduces low order harmonics [13]. In variable frequency drive systems the magnitude drop of the voltage subsequently leads to a reduction of the output electromagnetic torque. Therefore the deadtime should not be set excessively long compared with the total turn-off time of the IGBT devices. In the software the deadtime is set to 1μs.
4.2.5 Look-up Tables

Look-up tables are widely used in the program which provides an easy way to implement non-linear functions such as trigonometric operations. Speed/position operation curves are also stored in several tables. A brief introduction for all the look-up tables used in the program is attached for reference. ADCHZ and ADCVOL are designed to implement the volts-per-hertz curve. Speed and position profiles are stored in two look-up tables, ASP_S and ASP_P. Each table has 256 entries. A timer controls how frequently the speed and position loops are performed. In the program the period of the timer is set to 0.05s.

Table 4.9 Look-up tables used in the program

<table>
<thead>
<tr>
<th>Name</th>
<th>No. of Entries</th>
<th>Format</th>
<th>Range</th>
<th>Step size /entry</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>STBL</td>
<td>256</td>
<td>Q15</td>
<td>0 – 360°</td>
<td>1.41 °</td>
<td>Sine waveforms with third harmonic injection</td>
</tr>
<tr>
<td>ADCHZ</td>
<td>128</td>
<td>Hexadecimal</td>
<td>-64 – 63 Hz</td>
<td>1 Hz</td>
<td>Associating frequencies with the incremental step</td>
</tr>
<tr>
<td>ADCVOL</td>
<td>128</td>
<td>Q15</td>
<td>-64 – 63 Hz</td>
<td>1 Hz</td>
<td>Associating frequencies with voltages, controlling the V/Hz ratio</td>
</tr>
<tr>
<td>VLTPRE</td>
<td>64</td>
<td>Unsigned integer</td>
<td>0 – 150°</td>
<td>5 °</td>
<td>Interpreting the sampled NTC voltage to temperature</td>
</tr>
<tr>
<td>ASP_ST</td>
<td>256</td>
<td>Unsigned integer</td>
<td>0 – 12.8 s</td>
<td>0.05 s</td>
<td>Storing the desired speed profile</td>
</tr>
<tr>
<td>ASP_PT</td>
<td>256</td>
<td>Unsigned integer</td>
<td>0 – 32320 pulse</td>
<td>1 pulse</td>
<td>Storing the desired position profile</td>
</tr>
</tbody>
</table>
Chapter 5  Experimental Results

5.1  IGBT Drive

A clean and noise-free gate signal can assure the IGBTs to be turned on or off accordingly as they are required. Before the full dc voltage is applied, the IGBT gate signals are examined. Both low-side and high-side signals are plotted in Fig. 5.1 and Fig. 5.2 respectively.

![Low side IGBT gate signal, with Vdc = 0V and no load](image)

As the two figures are recorded separately, their time origins are not necessarily the same. Additionally, the gate signals might have different duty cycles.

![High side IGBT gate signal, with Vdc = 128 V and no load](image)

The gate signals are checked and recorded again with dc voltage applied and load connected. A dc voltage of 128V was obtained from 10 vehicle batteries connected in
series to achieve high power and high current output. A RL load was connected between one phase output and the negative dc bus to set up a half bridge topology.

![Graphs](a) V_{GE}(V) vs. Time(s)

Fig. 5.3 Low-side IGBT gate signal for (a) R = 7.5 Ω (b) R = 4 Ω (c) R = 2 Ω

Fig. 5.3 shows the gate signals obtained under different load currents. The arrows show where the high common mode dv/dt noise is coupled into the gate signals. The amplitude of the coupled voltage is about 3 volts and close to the estimation in Chapter 4. Due to the negative biased voltage during the off state, the spike is pushed far from the positive voltage region to avoid false IGBT conduction. The magnitude of the switching noise is current independent.
Fig. 5.4 High-side IGBT gate signal for (a) $R = 7.5 \, \Omega$ (b) $R = 4 \, \Omega$ (c) $R = 2 \, \Omega$

The gate signal of the high-side IGBT are also recorded and plotted in Fig. 5.4. Different from the low-side, switching noise can not be observed. This may be explained that during the deadtime when high-side and low-side IGBTs are both off, the charge accumulated at the emitter of the high-side IGBT is discharged or partially discharged.

Fig. 5.5 Half bridge setup and the discharge path during the deadtime
through the RL load, as shown in Fig. 5.5. When it comes to the moment that the low-side IGBT turns on, the voltage at the emitter of the high-side IGBT may have already dropped to a low level, which can not induce a high dv/dt noise coupled to the high-side gate signals. If the RL load is connected to the positive dc bus, a noise voltage will occur at the high-side IGBT gate signals while the gate signals of the companion IGBT remain quite clean.

The tested load currents are illustrated in Fig. 5.6. As the third harmonic injection is utilized, the phase-to-ground voltage is no longer sinusoidal.

Fig. 5.6 Load current for (a) $R = 7.5 \, \Omega$ (b) $R = 4 \, \Omega$ (c) $R = 2 \, \Omega$
Finally the line current is recorded and plotted below for an induction machine connected to the variable-frequency-drive. The same induction machine is used for all the motion control tests afterwards.

![Graph showing line current of the induction motor for different load torques](image)

Fig. 5.7 Line current of the induction motor for (a) $T = 0\ \text{Nm}$ (b) $T = 0.2\ \text{Nm}$

### 5.2 Motion Control

A validation experiment has been carried out to verify the conceptual design of the control system. The tested system consists of an induction machine, an optical encoder, the IGBT power module and its drive board, and a DSP Evaluation board. Parameters of the induction machine used in the experiment are listed below in Table 5.1. A 12-bit (4096 pulses per revolution) optical encoder is coupled directly to the motor shaft.
Table 5.1 Parameters of the induction machine used for the experiment

<table>
<thead>
<tr>
<th>Item</th>
<th>Specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated Voltage</td>
<td>120 V</td>
</tr>
<tr>
<td>Rated Power</td>
<td>175 W</td>
</tr>
<tr>
<td>Rated Current</td>
<td>1.5 A</td>
</tr>
<tr>
<td>Rated Speed</td>
<td>1670 rpm</td>
</tr>
<tr>
<td>Number of Poles</td>
<td>4</td>
</tr>
</tbody>
</table>

Four different control algorithms have been tested. Table 5.2 describes briefly the four control methods.

Table 5.2 Different control algorithms

<table>
<thead>
<tr>
<th>Control algorithm</th>
<th>Abbreviation</th>
</tr>
</thead>
<tbody>
<tr>
<td>Open loop speed, open loop position control</td>
<td>OSOP</td>
</tr>
<tr>
<td>Closed loop speed, open loop position control</td>
<td>CSOP</td>
</tr>
<tr>
<td>Open loop speed, closed loop position control</td>
<td>OSCP</td>
</tr>
<tr>
<td>Closed loop speed, closed loop position control</td>
<td>CSCP</td>
</tr>
</tbody>
</table>

For each control algorithm, different load conditions are tested, including no load, half the rated load, full load and variable load. The variable load test is designed to simulate the fluctuation of the load torque existed in the real system.

Table 5.3 Tested load conditions

<table>
<thead>
<tr>
<th>Item</th>
<th>Set Torque</th>
</tr>
</thead>
<tbody>
<tr>
<td>No load</td>
<td>0 Nm</td>
</tr>
<tr>
<td>Half the rated load</td>
<td>0.5 Nm</td>
</tr>
<tr>
<td>Full rated load</td>
<td>1 Nm</td>
</tr>
<tr>
<td>Variable load</td>
<td>1 - 0 - 1 - 0 - 1 Nm</td>
</tr>
</tbody>
</table>

Experimental results for open loop speed, open loop position control (OSOP) are shown in Fig. 5.8.
Fig. 5.8 Experimental results for open loop speed, open loop position control (OSOP)

The results show that noticeable errors exist in both speed and position control. The heavier the load, the greater the deviation in the speed and position. Moreover, in the
speed subplot, obvious sag of speed occurs for the variable load condition when the load happens to increase to its maximum value.

After first testing with both loops open, the speed loop is closed to compare the difference. The results are shown in Fig. 5.9

Fig. 5.9 Experimental results for closed loop speed, open loop position control (CSOP)
For CSOP control, both speed error and position error have been reduced thanks to the closed loop of speed. However, the speed error is still noticeable as well as the position error.

After the closed loop speed control, the closed loop position control was tested while the speed loop is open (OSCP). The results are illustrated in Fig. 5.10.
Fig. 5.10 Experimental results for open loop speed, closed loop position control (OSCP)

Since the position loop is closed, the position error has been dramatically reduced. In addition to the position, the speed error drops as well while no closed-loop is applied to speed control. Considering the position profile is derived from the speed curve, both of
them are mutually tied to each other. When the position loop tries to reduce its own error, it involuntarily regulates the speed as well, and thus accompanied with an improved speed performance.

The closed loop speed and closed loop position control (CSCP) is finally tested. The experimental results are given in Fig. 5.11.
Fig. 5.11 Experimental results for closed loop speed, closed loop position control (CSCP)

In terms of position or speed error, no improvement is observed between open loop speed, closed loop position control (OSCP) and closed loop speed, closed loop position control (CSCP). However, speed error reduces when compared to OSCP.
For all the tested scenarios, actual speed lags the required value during the acceleration process while it leads the latter during the deceleration period. That’s why from the results of both closed loop position controls, the position error experiences a trapezoidal change, i.e. it first increases, then keeps constant and finally drops to a very low level during the whole process.

The test results of all four controls are re-plotted in Fig. 5.12 for comparison. For each control, only those obtained under the worst case, i.e. with the rated load, is plotted for simplicity.

Since the tested system and the stage system are quite different, they will definitely exhibit different dynamic behaviors. In order to make the comparison between the experimental and simulation results more reasonable and consistent, a simulation based on the tested system has been carried out and the results are plotted in Fig. 5.13. We find the theoretical model of the tested system veraciously predicts the dynamic behavior of the real system.
Fig. 5.12 Testing results for all the control algorithms.
Fig. 5.13 Simulation results based on the tested system

Apart from the above experiments, an OSCP control with medium position error correction ($K_p=4$) is tested under rated load condition in order to obtain more accurate position control. The results are plotted in Fig. 5.14.
Fig. 5.14 Experimental results of OSCP control with medium position error correction

Table 5.4 summarizes the test of all the four control methods. In Table 5.4 as well as in Fig. 5.8 to Fig. 5.14, position and position errors are assumed to be referred to the
stage side and expressed in the form of pulses. One pulse is equivalent to a length of 6 mm measured at the edge of the stage.

Table 5.4 Summary of the test for all control methods

<table>
<thead>
<tr>
<th>Control algorithm</th>
<th>Speed Error Acceleration/Constant Speed /Deceleration (Hz)</th>
<th>Position Error (pulse)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Open loop speed, open loop position control (OSOP)</td>
<td>4/2.8/-0.1</td>
<td>107</td>
</tr>
<tr>
<td>Closed loop speed, open loop position control (CSOP)</td>
<td>3/1.5/-0.2</td>
<td>65</td>
</tr>
<tr>
<td>Open loop speed, closed loop position control (OSCP)</td>
<td>1.5/0/-1</td>
<td>3 (K_0=1)</td>
</tr>
<tr>
<td>Closed loop speed, closed loop position control (CSCP)</td>
<td>1.5/0/-1</td>
<td>1 (K_0=4)</td>
</tr>
</tbody>
</table>

Some conclusions can be drawn based on the simulation as well as the experimental results. Firstly, if the position loop is open, huge position error is inevitable to occur. As position is the main control target of the system, open loop position controls, both OSOP and CSOP, can not meet the requirements, thus are not recommended. Secondly, if closed loop position control is employed, position error accumulated in the acceleration process will be made up and cancelled during the deceleration period. The effect of closed position control is obvious. Thirdly, between open loop speed closed loop position control (OSCP) and closed loop speed closed loop position (CSCP) control, no big difference has been observed. However, OSCP is recommended as the control algorithm since it is simpler to implement.

In terms of the performance, the position error can be maintained at a level of around 1 pulse when an OSCP control with medium position error correction is used. The error level is very close to the control requirements stated in Chapter 3. Although the results are not obtained from the real stage system, we can still expect the final
implementation can exhibit as similar performance as obtained from the experimental setup and approach the desired control level.

In this chapter, some experimental results are presented. From the results, it can be concluded that the hardware as well as the software in the prototype system works well. The tested system can be operated in the way that is predefined by the acceleration-speed-position (ASP) profiles. Both speed and position errors are confined in an acceptable level.
Chapter 6  Conclusions and Recommendations

6.1  Conclusions

In this thesis the development of a DSP-controlled variable-speed drive has been considered. The design and implementation of a prototype involving motion control and machine drive have been proposed, analyzed and validated experimentally. From the observation, some points can be concluded as follows.

The cascaded structure for motion control exhibits several advantages including transparent and straightforward structure, flexibility for step-by-step design and commissioning, and the use of standard controllers. Thanks to the feed-forward signals produced by a reference generator, the slow response of the cascaded control with multiple loops is easily improved without any stability being sacrificed. Another merit of the reference signal is that the system can be controlled and operated following the desired speed and position curves along all the travel process.

Four different motion control algorithms, i.e. open loop speed open loop position (OSOP) control, closed loop speed open loop position (CSOP) control, open loop speed closed loop position (OSCP) control, and closed loop speed closed loop position (CSCP) control are simulated, and tested on an experimental system. It is observed both closed loop control methods can work well with the system while open loop speed closed loop position control is recommended for its simple implementation and good dynamic performance.

The controller is designed and optimized under the rated condition of the induction machine. However, as induction machines exhibit some non-linear characteristics in the low-speed region under scalar control, the controller must be verified for all the operating conditions. An aggressive controller, i.e. a system with strong error correction, is not suitable for this application.

The tested IGBT drive can work well under all the testing conditions with dc bus voltage up to 300V and maximum current up to 27A. When designing the gate drive, much effort has been made to maintain the immunity to high common-mode dv/dt noise for IGBT gate drives. A negative bias voltage during the off state for the gate drive brings remarkable safe operation margins for the IGBTs. Other measures, such as a well-
shielded interface between any high-voltage and low-voltage circuits, play important roles to assure IGBT modules work in a healthy condition. In addition, the gate resistor can be properly selected to control the switching speed of the output IGBT. There are different rules for sizing turn-on and turn-off resistors. If one gate resistor can not meet the requirements for both on and off operation, either two separate resistors and separate current paths must be employed, or we can use the turn-on gate resistor assisted with the negative bias voltage.

6.2 Recommendations

As a continuation to this work, emphasis could be focused on improving the performance, reliability and practicality of the whole system. Some more features can be incorporated to the IGBT drive, such as the anti-crossover protection, short-circuit/over current protection and fault feedback. Dc bus voltage and line current could also be used to realize more precise control. For the final work of the project, more accurate and authentic data of the stage mechanical system need to be measured, and accordingly tune-up the motion control system.
References


Appendix A: Estimating the Mass and Moment of Inertia of the Stage

The moment of inertia of a complex structure, such as the stage, containing steel frame and wood surface can in practice only be determined by approximation. Here an estimation of the moment of inertia of the stage is given for simulation.

The moment of inertia of the stage can be calculated as \( J = M \frac{R^2}{2} \) with \( M \) being the mass of the stage, \( R \) being the radius of the stage. The stage is a steel-framed wood-covered structure. The estimation is given by the table below.

Table a.1 Calculation of the moment of inertia of the stage

<table>
<thead>
<tr>
<th>Item</th>
<th>Calculation</th>
<th>Note</th>
</tr>
</thead>
<tbody>
<tr>
<td>Radius of the stage</td>
<td>( R = \frac{8.23}{2} = 4.115 \text{m} )</td>
<td>diameter of the stage is 27-foot.</td>
</tr>
<tr>
<td>Wood thickness</td>
<td>5cm</td>
<td>estimated</td>
</tr>
<tr>
<td>Volume of the wood part</td>
<td>( V = l \cdot \pi R^2 = 0.05 \times 3.14 \times 4.115^2 = 2.66 \text{m}^3 )</td>
<td></td>
</tr>
<tr>
<td>Mass density for wood</td>
<td>( \rho = 0.7 \times 10^3 \text{kg/m}^3 )</td>
<td></td>
</tr>
<tr>
<td>Mass of the wood</td>
<td>( M = V \rho = 2.66 \times 0.7 = 1.86 \times 10^3 \text{kg} )</td>
<td></td>
</tr>
<tr>
<td>Mass of the steel frame, scene setup and actors on the stage</td>
<td>1 to 2 ( \times 10^3 \text{kg} )</td>
<td>estimated</td>
</tr>
<tr>
<td>Total mass of the stage</td>
<td>3 to 4 ( \times 10^3 \text{kg} )</td>
<td></td>
</tr>
<tr>
<td>Total moment of inertia of the stage</td>
<td>25 to 34 ( \times 10^3 \text{kg} \cdot \text{m}^2 )</td>
<td></td>
</tr>
</tbody>
</table>

In most simulations, \( 25 \times 10^3 \text{kg} \cdot \text{m}^2 \) is used as the moment of inertia of the stage, including the scenery setups and actors standing on it.
Appendix B: Induction Motor Test and Parameter Estimation

Parameters on the nameplate of the machine are given in Table b.1.

Table b.1 Nameplate data of the induction machine

<table>
<thead>
<tr>
<th>Variable</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated output</td>
<td>3 hp (2.24kW)</td>
</tr>
<tr>
<td>Voltage</td>
<td>230 V</td>
</tr>
<tr>
<td>Full load current</td>
<td>8.2 A</td>
</tr>
<tr>
<td>Frequency</td>
<td>60 Hz</td>
</tr>
<tr>
<td>Speed</td>
<td>1760 rpm</td>
</tr>
<tr>
<td>Power factor</td>
<td>0.77</td>
</tr>
<tr>
<td>Efficiency</td>
<td>89.5%</td>
</tr>
</tbody>
</table>

Some tests have been carried out to estimate the detailed parameters of the machine. Tests include dc test, no load test and block rotor test. Testing results and parameter derivation are given below:

1) DC Test: \( r_s = 1/2 = 0.5 \Omega \)

2) No-load Test

\[
V_{as} = \frac{V_{i-1,NL}}{\sqrt{3}} = \frac{205.2}{\sqrt{3}} = 118.47V, \quad I_{NL} = 3.577A
\]

\[
S_{NL} = 1257.66VA, \quad P_{NL} = 108W, \quad Q_{NL} = 1253VAR, \quad Pf = 0.086
\]

\[
X_{NL} = \frac{Q_{NL}}{3I_{NL}^2} = \frac{1253}{3 \times 3.577^2} = 32.64\Omega
\]

3) Blocked-rotor Test

3.1) \( V_{as} = \frac{V_{i-1,NL}}{\sqrt{3}} = \frac{28.35}{\sqrt{3}} = 16.37V, \quad I_{BR} = 4.0015A \)

\[
R_{BR} = \frac{P_{BR}}{3I_{BR}^2} = \frac{56}{3 \times 4^2} = 1.166\Omega
\]

\[
X_{BR} = \sqrt{(V_{as} / I_{BR})^2 - R_{BR}^2} = \sqrt{(16.37 / 4.0015)^2 - 1.166^2} = 3.92\Omega
\]
\[ X_{ls} = X_{lr} = \frac{X_{BR}}{2} = 1.96\Omega \]

3.2) \[
V_{as} = \frac{V_{r-I,NL}}{\sqrt{3}} = \frac{38.65}{\sqrt{3}} = 22.31V, \quad I_{BR} = 5.995A
\]

\[
R_{BR} = \frac{P_{BR}}{3I_{BR}^2} = \frac{129}{3 \times 5.995^2} = 1.196\Omega
\]

\[
X_{BR} = \sqrt{(V_{as} / I_{BR})^2 - R_{BR}^2} = \sqrt{(22.31 / 5.995)^2 - 1.196^2} = 3.524\Omega
\]

\[
X_{ls} = X_{lr} = \frac{X_{BR}}{2} = 1.762\Omega
\]

3.3) \[
V_{as} = \frac{V_{r-I,NL}}{\sqrt{3}} = \frac{48.75}{\sqrt{3}} = 28.15V, \quad I_{BR} = 8.029A
\]

\[
R_{BR} = \frac{P_{BR}}{3I_{BR}^2} = \frac{230}{3 \times 8.0295^2} = 1.189\Omega
\]

\[
X_{BR} = \sqrt{(V_{as} / I_{BR})^2 - R_{BR}^2} = \sqrt{(28.15 / 8.0295)^2 - 1.189^2} = 3.298\Omega
\]

\[
X_{ls} = X_{lr} = \frac{X_{BR}}{2} = 1.649\Omega
\]

3.4) Average value:
\[
X_{ls} = X_{lr} = 1.79\Omega
\]

\[
X_M = X_{NL} - X_{ls} = 32.64 - 1.79 = 30.85\Omega
\]

4) Estimate \( r' \) from the torque-speed curve
\[
V_{as,th} = \frac{X_M}{\sqrt{r_s^2 + (X_{ls} + X_M)^2}} V_{as} = \frac{30.85}{\sqrt{0.5^2 + 32.64^2}} \frac{230}{\sqrt{3}} = 0.945 \times 132.8 = 125.5V
\]

\[
r_{s,th} + jX_{ls,th} = \frac{jX_M (r_s + jX_{ls})}{r_s + j(X_{ls} + X_M)} = 0.4466 + j1.6987
\]

\[
T_e = 3 \frac{V_{as,th}^2}{\omega_{syn}} \frac{r'_s / s}{(r_{s,th} + r'_s / s)^2 + (X_{ls,th} + X'_s)^2} = 12.19, \text{ find } r'_s = 0.422\Omega
\]
Appendix C: Induction Motor Qd0 Model

The equivalent circuit of an induction motor in arbitrary reference frame is developed in Fig. c.1. The rotor is short-circuited and the zero sequence circuit is neglected in the simulation process.

The following formulas are used to develop the equivalent circuit and model.

(a) Flux linkage per-second and reactance equations

\[
\frac{d\psi_{qs}}{dt} = \sigma_b (-r_s i_{qs} - \frac{\sigma}{\sigma_b} \psi_{ds} + v_{qs}) \quad \text{q-axis stator-side} \quad (c.1)
\]

\[
\frac{d\psi_{qr}}{dt} = \sigma_b (-r_r i_{qr} - \frac{\sigma - \omega_r}{\sigma_b} \psi_{dr} + v_{qr}) \quad \text{q-axis rotor-side} \quad (c.2)
\]
\[
\frac{d\psi_{ds}}{dt} = \sigma_b (-r_s i_{ds} + \sigma_b \psi_{qs} + \nu_{ds}) \quad \text{d-axis stator-side} \quad \text{(c.3)}
\]

\[
\frac{d\psi_{dr}}{dt} = \sigma_b (-r_s i_{dr} - \sigma_b \omega_r \psi_{qr} + \nu_{dr}) \quad \text{d-axis rotor-side} \quad \text{(c.4)}
\]

\[
\frac{d\psi_{os}}{dt} = \sigma_b (-r_s i_{os} + \nu_{os}) \quad \text{0-axis stator-side} \quad \text{(c.5)}
\]

\[
\frac{d\psi_{or}}{dt} = \sigma_b (-r_s i_{or} + \nu_{or}) \quad \text{0-axis rotor-side} \quad \text{(c.6)}
\]

(b) Output equations

\[
i_{qs} = \frac{1}{x_{ls}} (\psi_{qs} - \psi_{mq}) \quad \text{q-axis stator-side} \quad \text{(c.7)}
\]

\[
i_{ds} = \frac{1}{x_{ls}} (\psi_{ds} - \psi_{md}) \quad \text{d-axis stator-side} \quad \text{(c.8)}
\]

\[
i_{qr} = \frac{1}{x_{lr}} (\psi_{qr} - \psi_{mq}) \quad \text{q-axis rotor-side} \quad \text{(c.9)}
\]

\[
i_{dr} = \frac{1}{x_{lr}} (\psi_{dr} - \psi_{md}) \quad \text{d-axis rotor-side} \quad \text{(c.10)}
\]

\[
\psi_{mq} = x_{aq} (\frac{\psi_{qs}}{x_{ls}} + \frac{\psi_{qr}}{x_{lr}}) \quad \text{q-axis} \quad \text{(c.11)}
\]

\[
\psi_{md} = x_{ad} (\frac{\psi_{ds}}{x_{ls}} + \frac{\psi_{dr}}{x_{lr}}) \quad \text{d-axis} \quad \text{(c.12)}
\]

\[
x_{ad} = x_{aq} = \frac{1}{\frac{1}{x_m} + \frac{1}{x_{ls}} + \frac{1}{x_{lr}}} \quad \text{(c.13)}
\]

(c) Torque equations

\[
T_e = \frac{P}{2} \omega_b \left( \frac{3}{2} i_{qs} \psi_{qs} - i_{ds} \psi_{qs} \right) \quad \text{(c.14)}
\]

\[
J \frac{d\omega_m}{dt} = -T_{frc} + T_e \quad \text{(c.15)}
\]
\[
\omega_r = \left(\frac{P}{2}\right) \omega_m \quad \text{(c.16)}
\]

\[
J = J_{\text{rotor}} + J_{\text{load}} \quad \text{(c.17)}
\]

(d) Abcs-to-qd0 transformation

\[
V_{qd0s} = K_s \cdot V_{abcs} \quad \text{(c.18a)}
\]

\[
K_s = \frac{2}{3} \begin{bmatrix}
\cos \theta & \cos(\theta - \frac{2}{3} \pi) & \cos(\theta + \frac{2}{3} \pi) \\
\sin \theta & \sin(\theta - \frac{2}{3} \pi) & \cos(\theta + \frac{2}{3} \pi) \\
\frac{1}{2} & \frac{1}{2} & \frac{1}{2}
\end{bmatrix}, \theta = \theta(0) + \int \omega dt \quad \text{(c.18b)}
\]

(e) Qd0s-to-abcs transformation

\[
V_{abcs} = K_s^{-1} V_{qd0s} \quad \text{(c.19a)}
\]

\[
K_s^{-1} = \begin{bmatrix}
\cos \theta & \sin \theta & 1 \\
\cos(\theta - \frac{2}{3} \pi) & \sin(\theta - \frac{2}{3} \pi) & 1 \\
\cos(\theta + \frac{2}{3} \pi) & \sin(\theta + \frac{2}{3} \pi) & 1
\end{bmatrix}, \theta = \theta(0) + \int \omega dt \quad \text{(c.19b)}
\]
Appendix D: Linearized Induction Machine Model
under Variable-Frequency Operation

The voltage equation for induction machine with current as state variables may be
written in the synchronously rotating frame as

$$
\begin{bmatrix}
    v_{qs}^e \\
    v_{ds}^e \\
    v_{qr}^e \\
    v_{dr}^e
\end{bmatrix} =
\begin{bmatrix}
    r_s & \frac{\omega_e}{\omega_b} X_{ss} & \frac{p}{\omega_b} X_M & \frac{\omega_e}{\omega_b} X_M \\
    -\frac{\omega_e}{\omega_b} X_{ss} & r_s + \frac{p}{\omega_b} X_{ss} & \frac{\omega_e}{\omega_b} X_M & \frac{p}{\omega_b} X_M \\
    \frac{p}{\omega_b} X_M & s \frac{\omega_e}{\omega_b} X_M & r'_r + \frac{p}{\omega_b} X'_{rr} & s \frac{\omega_e}{\omega_b} X'_{rr} \\
    -s \frac{\omega_e}{\omega_b} X_M & \frac{p}{\omega_b} X_M & -s \frac{\omega_e}{\omega_b} X'_{rr} & r'_r + \frac{p}{\omega_b} X'_{rr}
\end{bmatrix}
\begin{bmatrix}
    i_{qs}^e \\
    i_{ds}^e \\
    i_{qr}^e \\
    i_{dr}^e
\end{bmatrix}
$$

(d.1)

Assume $v = [v_{qs}^e, v_{ds}^e, v_{qr}^e, v_{dr}^e]^T$ and $i = [i_{qs}^e, i_{ds}^e, i_{qr}^e, i_{dr}^e]^T$. Rewrite (d.1) we have

$$
v = ri + \omega_e Fi + \omega_r Gi + pQi
$$

(d.2)

where

$$
\begin{bmatrix}
    r_s & 0 & 0 & 0 \\
    0 & r_s & 0 & 0 \\
    0 & 0 & r'_r & 0 \\
    0 & 0 & 0 & r'_r
\end{bmatrix}, \quad
\begin{bmatrix}
    0 & L_{ss} & 0 & L_M \\
    -L_{ss} & 0 & -L_M & 0 \\
    0 & L_M & 0 & L_{rr} \\
    -L_M & 0 & -L_{rr} & 0
\end{bmatrix}, \quad
\begin{bmatrix}
    0 & 0 & 0 & 0 \\
    0 & 0 & 0 & 0 \\
    0 & -L_M & 0 & -I_{rr} \\
    L_M & 0 & L_{rr} & 0
\end{bmatrix}, \quad
\begin{bmatrix}
    L_{ss} & 0 & L_M & 0 \\
    0 & L_{ss} & 0 & L_M \\
    L_M & 0 & L_{rr} & 0 \\
    0 & L_M & 0 & L_{rr}
\end{bmatrix}
$$

Rearrange (d.2) we get

$$
pi = -Q^{-1}ri - \omega_e Q^{-1}Fi - \omega_r Q^{-1}Gi + Q^{-1}v
$$

(d.3)

If we expand the above equation at an operation point $(i_0, \omega_{e0}, \omega_{r0}, v_0)$, we have

$$
p(i_0 + \Delta i) = -Q^{-1}r(i_0 + \Delta i) - (\omega_{e0} + \Delta \omega_e) Q^{-1}F(i_0 + \Delta i) \\
- (\omega_{r0} + \Delta \omega_r) Q^{-1}G(i_0 + \Delta i) + Q^{-1}(v_0 + \Delta v)
$$

(d.4)

Substitute (d.3) to (d.4) we get

$$
p\Delta i = -Q^{-1}r\Delta i - \omega_{e0} Q^{-1}F\Delta i - \Delta \omega_e Q^{-1}Fi_0 - \omega_{r0} Q^{-1}G\Delta i - \Delta \omega_r Q^{-1}Gi_0 + Q^{-1}\Delta v
$$

(d.5)
The electromagnetic torque is expressed as

\[ T_e = X_M (i_{qf}^e i_{d}^e - i_{d}^e i_{qf}^e) \] or \[ T_e = i^T G i \] (d.6)

The relationship between torque and speed in per-unit is

\[ T_e - T_L = 2H p \omega_r \] (d.7)

Substitute (d.6) into (d.7) we get

\[ p \omega_r = \frac{1}{2H} (T_e - T_L) = \frac{1}{2H} i^T G i - \frac{1}{2H} T_L \] (d.8)

where \( H \), the inertia constant expressed in seconds, is

\[ H = \left( \frac{1}{2} \right) P \frac{J \omega_r^2}{P_b} \] (d.9)

Expand the above equation at an operation point \((\omega_r, i_0, T_{e0}, T_{L0})\), we have

\[ p (\omega_r + \Delta \omega_r) = \frac{1}{2H} (i_0 + \Delta i)^T G (i_0 + \Delta i) - \frac{1}{2H} (T_{L0} + \Delta T_L) \] (d.10)

Substitute (d.8) into (d.10), and \( \Delta i^T G i_0 = i_0^T G^T \Delta i \), we get

\[ p \Delta \omega_r = \frac{1}{2H} i_0^T (G^T + G) \Delta i - \frac{1}{2H} \Delta T_L \] (d.11)

Therefore the input equation of the induction motor and the mechanical system is

\[ p \begin{bmatrix} \Delta i \\ \Delta \omega_r \end{bmatrix} = \begin{bmatrix} -Q^{-1} (r + \omega_e F + \omega_r G) \\ \frac{1}{2H} i_0^T (G^T + G) \end{bmatrix} \begin{bmatrix} \Delta i \\ \Delta \omega_r \end{bmatrix} - \begin{bmatrix} -Q^{-1} G i_0 \\ 0 \end{bmatrix} \begin{bmatrix} \Delta v \\ \Delta \omega_e \end{bmatrix} \] (d.12)

Output equation is

\[ \begin{bmatrix} \Delta T_e \\ \Delta \omega_r \end{bmatrix} = \begin{bmatrix} i_0^T (G^T + G) \end{bmatrix} \begin{bmatrix} \Delta i \\ \Delta \omega_r \end{bmatrix} + \begin{bmatrix} 0 \\ 0 \end{bmatrix} \begin{bmatrix} \Delta v \\ \Delta \omega_e \end{bmatrix} - \begin{bmatrix} 0 \\ 0 \end{bmatrix} \begin{bmatrix} \Delta T_L \end{bmatrix} \] (d.13)

Assume

\[ M = -Q^{-1} (r + \omega_e F + \omega_r G) \] (d.14a)

\[ N = i_0^T (G^T + G) \] (d.14b)

\[ R = -Q^{-1} G i_0 \] (d.14c)
\[ K = -Q^{-1}F_i \]  

Then (d.12) can be rewritten as

\[ \Delta \Delta \Delta = M \Delta \Delta + R \Delta \omega_r + Q^{-1} \Delta \Delta + K \Delta \omega_e \]  

\[ \Delta \omega_r = \frac{N}{2H} \Delta \Delta - \frac{1}{2H} \Delta T_L \]  

Rearrange (d.15a) we have

\[ \Delta \Delta = s \Delta - M \Delta ^{-1} (R \Delta \omega_r + Q^{-1} \Delta \Delta + K \Delta \omega_e) \]  

where the operator \( p \) has been replaced by the Laplace operator \( s \) commonly used in transfer function formulation and \( I \) is the unity diagonal matrix. Substitute (d.14b) and (d.16) into (d.13) we get

\[ \Delta T_e = N \Delta \Delta = N \Delta \Delta ^{-1} (R \Delta \omega_r + Q^{-1} \Delta \Delta + K \Delta \omega_e) \]  

Define \( G_{or} = N \Delta \Delta ^{-1} R \)  
\( G_v = N \Delta \Delta ^{-1} Q^{-1} \)  
\( G_{oe} = N \Delta \Delta ^{-1} K \)

we can rewrite the torque equation (d.17) as

\[ \Delta T_e = G_{or} \Delta \omega_r + G_v \Delta \Delta + G_{oe} \Delta \omega_e \]  

Expand (d.7) at the operation point \( (\omega_{r0}, T_{e0}, T_{L0}) \) we have

\[ \Delta T_e - \Delta T_L = 2H \Delta \Delta \Delta \]  

or

\[ \Delta \omega_r = \frac{1}{2H} \int \frac{1}{s} (\Delta T_e - \Delta T_L) \]  

![Fig. d.1 Transfer function for variable-frequency operated induction motor](image)

Using (d.19) and (d.21), Fig. d.1 gives the decoupled transfer function of a variable-frequency operated induction motor around a given operating point.
Appendix E: Computing Stator and Rotor Currents for Induction Machines in qd0 System

From the per-phase equivalent circuit for steady-state operation of a symmetrical induction machine, the input impedance of the equivalent circuit shown in Fig. e.1 with rotor side short-circuited is

\[
Z = \frac{r_s r_r' + \omega_e^2 (L_M^2 - L_s L_r') + j \omega_e (r_r' L_s + r_s L_r')}{s} + j \omega_e L_r'
\]

Then we have

\[
\tilde{I}_{as} = \frac{\tilde{V}_{as}}{Z} \tag{e.2}
\]

\(\tilde{I}_{bs}\) and \(\tilde{I}_{cs}\) have the same magnitude as \(\tilde{I}_{as}\), except that they have \(\frac{2}{3}\pi\) and \(-\frac{2}{3}\pi\) phase shift to \(\tilde{I}_{as}\), i.e.

\[
\tilde{I}_{bs} = \tilde{I}_{as} [\cos(-\frac{2}{3}\pi) + j \sin(-\frac{2}{3}\pi)] \tag{e.3}
\]

and

\[
\tilde{I}_{cs} = \tilde{I}_{as} [\cos(\frac{2}{3}\pi) + j \sin(\frac{2}{3}\pi)] \tag{e.4}
\]

For single fed machine, \(V'_{ar} = 0\) whereupon \(\tilde{I}_{ar}' = -\frac{j \omega_e L_M}{r_s} \tilde{I}_{as} \tag{e.5}\)
\( \tilde{I}_{br} \) and \( \tilde{I}_{cr} \) have the same magnitude as \( \tilde{I}_{ar} \), except that they have \( \frac{2}{3} \pi \) and \( -\frac{2}{3} \pi \) phase shift to \( \tilde{I}_{ar} \), i.e.

\[
\tilde{I}_{br} = \tilde{I}_{ar} [\cos(-\frac{2}{3} \pi) + j \sin(-\frac{2}{3} \pi)] \tag{e.6}
\]

and

\[
\tilde{I}_{cr} = \tilde{I}_{ar} [\cos(\frac{2}{3} \pi) + j \sin(\frac{2}{3} \pi)] \tag{e.7}
\]

After obtaining all the variables in the abc system, we can transfer them to the qd0 system by

\[
[I_q s \quad I_d s \quad I_o s]^T = K_s [\tilde{I}_{ar} \quad \tilde{I}_{br} \quad \tilde{I}_{cr}]^T \tag{e.8}
\]

where

\[
K_s = \frac{2}{3} \begin{bmatrix}
\cos \theta & \cos(\theta - \frac{2}{3} \pi) & \cos(\theta + \frac{2}{3} \pi) \\
\sin \theta & \sin(\theta - \frac{2}{3} \pi) & \cos(\theta + \frac{2}{3} \pi) \\
\frac{1}{2} & \frac{1}{2} & \frac{1}{2}
\end{bmatrix} \tag{e.9}
\]

In (e.9), we can assume \( \theta = 0 \) to simplify the computation if synchronous frame is used, and assume the q-axis and a-axis are in-phase.

Similar to the transfer of stator current, rotor current in the qd0 system can be obtained by

\[
[I_q r \quad I_d r \quad I_o r]^T = K_s [\tilde{I}_{ar} \quad \tilde{I}_{br} \quad \tilde{I}_{cr}]^T \tag{e.10}
\]
Appendix F: Modified Limited-Jerk ASP Profile

A big trouble, found from the simulation results of both time-optimal ASP profile and limited-jerk ASP profile, is that there is offending oscillation of the acceleration during the first second of the operation.

![Fig. f.1 Acceleration of the time-optimal ASP profile](image1)

![Fig. f.2 Acceleration of the limited-jerk ASP profile](image2)

Table f.1 gives the jerk for each ASP profile.

<table>
<thead>
<tr>
<th></th>
<th>Time-optimal ASP</th>
<th>Limited-jerk ASP</th>
</tr>
</thead>
<tbody>
<tr>
<td>Jerk $\gamma$</td>
<td>-1150 to 1280 rad/s$^3$</td>
<td>-1100 to 1284 rad/s$^3$</td>
</tr>
</tbody>
</table>
The oscillation reveals the intrinsic disadvantage of scalar control. In scalar control, the torque component and field component at the stator current are not decoupled. The air-gap flux needs several seconds to be built up. However, the scalar-controlled drive does not have any knowledge of that. The stator voltage seen from a qd system is not in phase with the air-gap flux, so there must be considerable fluctuation in the air-gap flux, which is then transferred to the electromagnetic torque of the machine.

Amin proposed a quantitative way to analyze the operating conditions of scalar control [4], which is based on the observation of the time-constants of induction machines. Time-constants can be easily derived from eigenvalues of the machine. Eigenvalues satisfy the characteristic equation of matrix A of the state equation, i.e.

\[
\text{det}(A - \lambda I) = 0
\]  

(f.1)

where I is the identity matrix. The eigenvalues could provide behavior of an induction machine at any balanced operating conditions. Negative real parts are the sign to show any state variables decrease exponentially with time, thus the system will gradually run to its steady-state. Table f.2 lists the eigenvalues of the 3-hp induction machine under different conditions.

<table>
<thead>
<tr>
<th>Eigenvalue</th>
<th>Stall</th>
<th>Rated Speed</th>
<th>No Load</th>
</tr>
</thead>
<tbody>
<tr>
<td>(\lambda_{1,2})</td>
<td>(-97.21\pm j377)</td>
<td>(-54.55\pm j371)</td>
<td>(-54.56\pm j371)</td>
</tr>
<tr>
<td>(\lambda_{3,4})</td>
<td>(-2.85\pm j377)</td>
<td>(-26.31\pm j32.2)</td>
<td>(-22.64\pm j33.4)</td>
</tr>
<tr>
<td>(\lambda_5)</td>
<td>0.45</td>
<td>-37.95</td>
<td>-45.26</td>
</tr>
</tbody>
</table>

If we express the eigenvalues as \(\lambda_{1,2} = -\frac{1}{\tau_1} \pm j\omega_1\), \(\lambda_{3,4} = -\frac{1}{\tau_2} \pm j\omega_2\) and \(\lambda_5\), \(\tau_1\) and \(\tau_2\) represent the two time-constants of the machine, one referred to the stator side and the other referred to the rotor side. There are five eigenvalues, two from the stator, two from the rotor and one from the mechanical system. As the electromagnetic torque is
concerned in this chapter, discussion will be focused on the variables associated with electromagnetism.

Plot of the electromagnetic time-constants versus motor mechanical speed in rpm are shown in Fig. f.3.

![Electromagnetic time-constant vs. rotor mechanical speed](image)

Fig. f.3 Electromagnetic time-constants of the induction machine

One of the two time-constants is from the stator side and the other is from the rotor side. The major time-constant is contributed by the stator side, while it is also the dominant time-constant for the whole speed range. At the low speed area, especially the stall state, the major time-constant varies from 0.1 – 0.3 s. Compared with the rotor side time-constant, the stator side time-constant is much bigger, which casts a huge influence on the performance of the scalar control at low speed range, i.e. the dynamic response is by and large slow. For medium and high rotor speed area, the major time-constant decreases quickly and it keeps at the range of 0.02 – 0.04s, which makes the response much faster at that speed range. From the perspective of physical essential, the stator side time-constant can be approximated by $\frac{L_M}{r_s}$ (0.16s) [14], because the mutual magnetizing inductance is large, it decides that the time-constant can’t be small during the build-up of
the air-gap field. However, the time-constant of the rotor side is always smaller than the stator side time-constant, so it is a minor time-constant in the system and can be neglected. We may estimate the rotor side time-constant by using

\[ \tau_r = \frac{L_{is} + L_{ip}}{r_i + \frac{r_s}{s}} \]  

which, as a function of slip, ranges from 0.01 to 0 second.

For scalar control method, the process of calculating the magnitude and phase of stator voltage must be continually repeated at the end of some step-time \( \Delta t \) [1]. If it is assumed that the length of period \( \Delta t \) is at least five times of the major electromagnetic time-constant, \( \tau_{em, ma} \), of the motor, i.e.

\[ \Delta t \geq 5\tau_{em, ma} \]  

then we have

\[ T_e - \frac{T_f}{\lambda} = (J_M + \frac{J_L}{\lambda^2}) \frac{d\omega_{rm}}{dt} \approx (J_M + \frac{J_L}{\lambda^2}) \frac{\Delta\omega_{rm}}{\Delta t} \]  

We also assume that the relative angular speed must not have changed too much during the period of \( \Delta t \), for example 10% when motoring or -10% when braking, therefore

\[ \frac{\Delta\omega_{rm}}{\omega_{rm}} = \pm 0.1 \]  

Combining (f.3), (f.4) and (f.5) we have

\[ T_e = \frac{T_f}{\lambda} + (J_M + \frac{J_L}{\lambda^2}) \frac{\omega_{rm}}{50\tau_{em, ma}} \]  

(f.6) defines the admissible electromagnetic torque output for induction motors under scalar control [4].

A simulation based on MATLAB/Simulink has been developed to analyze the admissible torque, actual torque together with the flux linkage per-second of the air-gap during the first few seconds of the acceleration period. The results, plotted together in Fig. f.4, show that actual torque exceeds the admissible torque between 0.3 – 1.3 s. And right during that period the electromagnetic torque has huge fluctuation.
If we look at the air-gap flux, we find that the flux fluctuates dramatically during that period as well. Theoretically the flux transfers its fluctuation to the output torque. If we could prevent that from happening, we might regulate the torque to have a smoother output. Note there is friction existing in the mechanic system, if the ASP profile is redesigned to make the motor try to output a torque that does not exceed the maximum static friction, the friction torque will contradict the electromagnetic torque and the stage will be held standstill during that period. Although there is no speed output from the motor, the air-gap flux will gradually be established. After the air-gap flux grows up to some certain level, its fluctuation decreases and then the commanded torque can resume to what is desired in the limited-jerk curve. Therefore a modified limited-jerk ASP profile has been developed from the original limited-jerk ASP profile.

In the new ASP profile, an inflection point is introduced to separate the acceleration period into two steps, as shown in Fig. f.5
Fig. 6.5 Modified limited-jerk ASP profile

Between 0 to $t_5$, a gentle slope for the acceleration is adopted and in the second phase between $t_5$ to $t_1$, a slope with normal value is used to accelerate the system.
From the simulation illustrated in Fig. f.6 we can find the start-up transient has been improved such that only small oscillations can be observed when a modified ASP profile is used. Table 3.4 provides a comparison for the jerk obtained under three different ASP profiles.

Table 3.4 Jerks for different ASP profiles

<table>
<thead>
<tr>
<th></th>
<th>Time-optimal ASP</th>
<th>Limited-jerk ASP</th>
<th>Modified limited-jerk ASP</th>
</tr>
</thead>
<tbody>
<tr>
<td>Jerk $\gamma$</td>
<td>-1150 to 1280 rad/s$^3$</td>
<td>-1100 to 1284 rad/s$^3$</td>
<td>-200 to 400 rad/s$^3$</td>
</tr>
</tbody>
</table>
Fig. f.7 Admissible torque, actual torque and flux linkage per-second of the air-gap with limited-jerk ASP profile during the acceleration period

Torques and air-gap flux are plotted again in Fig. f.7 for the modified limited-jerk ASP profile. The moment that the actual torque exceeds the admissible torque is delayed for 0.8s, which gives time for the air-gap flux to build up to about 60% of its rated level.

An induction motor can only output satisfactory torque when an appropriate magnetic flux level in the air-gap has been established. The modified limited-jerk ASP profile, by avoiding accelerate the system during the period when the air-gap is experiencing its severest oscillations, is expected to generate much smoother electromagnetic torque, and thus, improves the riding comfortability.
Appendix G: Flowchart of the Software

The main program consists of three parts, an initialization setup, an endless main loop and several interrupt service routines. Detailed code is eliminated in the thesis. Here are some flowcharts summarizing the program.

Fig. g.1 Flowchart of the main program – CPU initialization, Event Manager modules setup and interrupt setup
Initialize tables, pointers and variables for specific application

- Initialize pointers for phase A, B and C
- Assign TOPTABLE to the beginning address of wave table STBL
- Initialize step size, both high bits and low bits
- Initialize MODREGL/H registers for three phases
- Initialize TOPHZ to the beginning address of look-up table ADCHZ
- Initialize TOPVOL to the beginning address of table ADCVOL
- Initialize TOPTP to the beginning address of table VLTPRE
- Initialize pointers for speed/position control ASP profiles
- Initialize variables associated with operation direction, revolutions, timers, look-up table indices, stop sign, output frequencies, measured speed and positions

- Setup I/O pins for brake leg control

Main loop

- Endless loop waiting for interrupt

Fig. g.2 Flowchart of the main program – initialization of other variables, I/O pins setup and main loop
In all the Interrupt Service Routines, ADCISR homes the several functions, including temperature detection, speed estimation, speed and position recording and motion control. Interrupt Service Routine SINE generates the PWM waveform with the frequency and magnitude assigned by motion control block.
MODREGA = MODREGA + FRQSTEPA

Discard (Zero) the highest 8 bits of MODREGHA

TABLEA = MODREGHA

Read the value from wave table with the offset address TABLEA and store it to SINEVALA

Do the same operations for phase B as in phase A

Do the same operations for phase C as in phase A

LOADMA

Fig. 4.4 Flowchart of the sub-function SINE
Sub-function SINE generates PWM sinusoidal waveforms and transfers the results to sub-function LOADMA which normalizes the sine wave values to a proper format that used by EVA registers.

LOADMA

\[ \text{SMAVAL} = \text{MAVAL} \times \text{normal} \]
\[ \text{in Q15 format} \]

\[ \text{CMPAR1} = \text{SINEVALA} \times \text{SMAVAL} \]

\[ \text{CMPR1} = \text{CMPAR1} + \text{normal} \]
\[ \text{in Q15 format} \]

Do the same operations for phase B as above in phase A

Do the same operations for phase C as above in phase A

STOP = 1?

Yes

Turn off output PWM

No

return

Fig. 6.6 Flowchart of the sub-function LOADMA
ADCISR

Detect power module temperature

TPRE > 100?

Yes → STOP = 1

No →

Estimate mechanical speed and position

Record mechanical speed (or position error) and position along time

Speed control

Position control

Speed conversion

return

Fig. g.6 Flowchart of the sub-function ADCISR