CHAPTER 6

DIGITAL REALIZATION OF VECTOR CONTROL SCHEME

6.1 INTRODUCTION

General purpose AC drives often feature a cost advantage over their DC counterparts and in addition, offer lower maintenance, smaller motor size and improved reliability. However, the control flexibility available with these drives is limited and their application is restricted to fans, pumps and compressors where the speed regulation is not so crucial and where transient response and low-speed performance are not critical.

More demanding drives used in machine tools, spindles, high-speed elevators, dynamometers, mine winders, rolling mills, glass float lines etc., have much more sophisticated requirements and must afford the flexibility to allow for regulation of a number of variables, such as speed, position, acceleration and torque. Until recently, such drives were almost exclusively the domain of DC motors combined with various configurations of power converters depending upon the application. With suitable control, however, induction motor drives have been shown to be more than a match for DC drives in high-performance applications. While control of the induction machine is considerably more complicated than its DC motor counterpart, with continual advancement of microelectronics, these control complexities have essentially been overcome.

6.2 SOLID-STATE AC DRIVES
AC motor drives can be broadly categorized into two types, namely thyristor-based and transistor-based drives. Thyristors possess the capability of self turn-on by means of an associated gate signal but must rely upon circuit conditions to turn-off, whereas transistor devices are capable of both turn-on and turn-off. Because of their turn-off limitations, thyristor based drives must utilize an alternating EMF to facilitate switching-off of the devices (commutation) which requires reactive volt-amperes from the EMF source to accomplish.

A brief list of the available drive types is as shown in Figure 6.1. The drives are categorized according to switching nature (natural or force commutated), converter type and motor type. The speed control of the four major drive types having differing control principles are

1. Voltage controlled induction motor drives
2. Load commutated synchronous motor drives
3. Volts per hertz and vector controlled induction motor drives and
4. Vector controlled permanent magnet motor drives.

The control principles of the remaining drive of Figure 6.1 are generally straightforward variations of one of these four drive types.
6.3 PRINCIPLE OF VECTOR CONTROL

High performance control of AC induction motors and permanent magnet synchronous motors most often relies on the principles of vector control or Field Oriented Control (FOC). Vector controllers mainly aim to maintain the flux, producing the direct component of the stator current space vector, in phase with the rotor flux space vector, under all operating conditions. The quadrature axis current component, which then lies in quadrature with the rotor flux vector, directly controls the torque developed by the machine. When properly implemented, vector control permits the independent control of the torque and flux of the AC machines, in a manner identical to that of the separately excited DC motor (Bose 2004, Krishnan 2006).

Although the large majority of variable speed applications require only speed control in which the torque response is only of secondary interest, in more challenging applications such as traction, servomotor type of drives depend critically upon the ability of the drive to provide a prescribed torque, whereupon the speed becomes the variable of secondary interest. The method
of torque control in machines is called either vector control or alternatively FOC. Vector control refers to the manipulation of terminal currents, flux linkages and voltages to affect the motor torque, while field orientation refers to the manipulation of the field quantities within the motor itself. Since it is common for machine designers to visualize motor torque production in terms of the air gap flux densities and MMFs instead of currents and fluxes, which relate to terminal quantities, generally this scheme is referred as FOC (Texas Instruments 1998).

FOC consists of controlling the stator currents represented by a vector. This control is based on projections, which transform a three-phase time and speed dependent system into a two co-ordinate (d and q co-ordinates) time invariant system. These projections lead to a structure similar to that of a DC machine control. Field orientation controlled machines need two constants as input references: the torque component (aligned with the q co-ordinate) and the flux component (aligned with d co-ordinate). As FOC is simply based on projections, the control structure handles instantaneous electrical quantities. This makes the control accurate in every working operation (steady state and transient) and independent of the limited bandwidth mathematical model (Aounis et al 2002). The FOC thus solves the classic scheme problems, in the following ways:

1. The ease of reaching constant reference (torque component and flux component of the stator current)

2. The ease of applying Direct Torque Control (DTC).

By maintaining the amplitude of the rotor flux at a fixed value, we have a linear relationship between torque and torque component \( i_{\text{sq}} \). Hence
the torque can be controlled by controlling the torque component of stator current vector.

This control structure shows an interesting advantage: it can be used to control either synchronous or induction machines by simply changing the flux reference and obtaining rotor flux position. As in Permanent Magnet Synchronous Motors (PMSM), the rotor flux is fixed (determined by the magnets) and there is no need to create one. Hence, when controlling a PMSM, the flux reference should be set to zero. As induction motors need a rotor flux creation in order to operate, the flux reference must not be zero. This conveniently solves one of the major drawbacks of the “classic” control structure, the portability from asynchronous to synchronous drives.

Figure 6.2 shows the original (a, b, c), stator-flux-oriented (α, β) and rotor-flux-oriented (d, q) reference frames, and the correct position of the rotor flux (ψr), the stator current (i_s) and stator voltage (v_s) space vector that rotates with d, q reference at synchronous speed.

The measure of the rotor flux position is different if synchronous or induction motors are considered.

- In the synchronous machine, the rotor speed is equal to the rotor flux speed. Then ☐θ (rotor flux position) is directly measured by position sensor or by integration of rotor speed.

- In the induction machine the rotor speed is not equal to the rotor flux speed (there is a slip speed), and then it needs a particular method to calculate θ.
Thus FOC makes it possible to control, directly and separately, the torque and flux of AC machines. Field oriented control of AC machines thus obtain all the advantages of the control of DC machines like instantaneous control of the separate quantities allowing accurate transient and steady state management. In addition to this advantage, FOC AC machines solve the mechanical commutation problems inherent with DC machines.

6.4 FOC SCHEME ADOPTED IN THIS WORK

In contrast to inverters which function as voltage sources, inverters which function as current sources are becoming the main power sources of high performance AC machine drives. The major advantages of these inverters are simplicity, inherent ability for regeneration and reversal, usage of even converter-grade thyristors and safety against misfire of thyristors.

There are many control schemes known for induction machine drives with impressed currents, but rotor-flux oriented control has emerged as one of the most frequently used techniques. The application of this technique yields fast dynamic response. This technique can be used in direct or indirect
approaches. When the direct method (flux-feedback control) is used, the space angle of the rotor flux-linkage space phasor is obtained by direct measurements (e.g., by using Hall sensors) or by using flux model. However when the indirect method (feed-forward control) is used, the space angle of the rotor flux-linkage space phasor is obtained as the sum of the monitored rotor angle \((\theta_r)\) and the computed reference value of the slip angle \((\theta_s)\), where the slip angle gives the position of the rotor flux-linkage space phasor relative to the rotor (or more precisely relative to the direct axis of the reference frame fixed to the rotor).

In this work, direct implementation of the rotor flux-oriented control of an induction machine supplied by a current-controlled PWM inverter is achieved (Figure 6.3).

As shown in Figure 6.3, in the direct method, the reference value of the rotor speed \((\omega_{ref})\) is compared with the monitored value of the rotor speed \((\omega_r)\), the error signal is supplied to the input of the speed controller, a PI controller, and the output of which is the reference value of the electromagnetic torque \((t_{eref})\). Comparison of \(t_{eref}\) and the actual value of the torque \(t_e\) gives an error which serves as input to the torque controller, also a PI controller, and its output is the reference value of the quadrature-axis stator current expressed in the rotor-flux oriented reference frame \((i_{syref})\).
Figure 6.3 Block diagram of vector controller designed in VHDL

The direct-axis stator current reference \( (i_{sxref}) \), which is expressed in the rotor flux oriented reference frame, is obtained as the output of the flux controller (PI controller) the input of which is the difference between the reference value of the rotor magnetizing current \( i_{mref} \) and the actual value of the rotor magnetizing current \( i_{mr} \). The reference current \( i_{mref} \) is obtained as the output of the function generator (FG), which allows field weakening to be implemented. Thus the input of FG is the monitored rotor speed. Below base speed FG gives a constant value of \( i_{mref} \), while above base speed \( i_{mref} \) is inversely proportional to the rotor speed. The relationship between the two reference frame quantities are derived as (Vas 1998)

\[
\begin{align*}
i_{sDref} &= i_{sxref}\cos \rho_r - i_{syref}\sin \rho_r \\
i_{sQref} &= i_{sxref}\sin \rho_r + i_{syref}\cos \rho_r
\end{align*}
\]

(6.1)  
(6.2)

In accordance with Equations (6.1) and (6.2), the stator current references \( i_{sxref} \) and \( i_{syref} \) are first transformed into the two-axis stator current references of the stationary reference frame \( (i_{sDref}, i_{sQref}) \) by the application of the transformation \( e^{j\theta} \) (inverse Park’s Transformation) where \( \rho_r \) is the angle of the rotor magnetizing-current space phasor with respect to the direct-axis (sD) of the stationary reference frame.
The two-axis current references are then transformed into their three-phase reference values \( (i_{sAref}, i_{sBref}, i_{sCref}) \) by the application of two-phase to three-phase transformation (inverse Clarke’s Transformation), given by Equations (6.3) to (6.5), which is indicated by the block labeled ‘2→3’ in Figure 6.3. These are used, together with the monitored three-phase currents \( (i_sA, i_sB, i_sC) \), to obtain the gate signals necessary for the inverter, which drives the induction machine.

The outputs of the flux model are the rotor magnetizing current \( (i_{mr}) \), the torque-producing stator current \( (i_{sy}) \) and the spatial position of the rotor flux-linkage space phasor \( (\psi) \). The electromagnetic torque \( (t_e) \) is obtained by multiplying \( i_{sy} \) by the constant \( c = \frac{3PL_e}{(2L_r)} \), where \( P \) is the number of pole pairs, \( L_m \) is the magnetizing inductance of the machine and \( L_r \) is the self-inductance of the rotor. This is given in Equation (6.6).

\[
t_e = \frac{3}{2} \frac{P}{L_r} i_{sy} \quad (6.6)
\]

The monitored stator currents \( (i_{sA}, i_{sB}, i_{sC}) \), together with the monitored rotor speed \( (\omega_r) \), are inputs to the flux model (Figure 6.4). This implementation utilizes the rotor time constant \( (T_r) \) for the determination of the value of \( i_{mr} \). The three phase monitored currents are transformed into their two-axis components by the application of the three-phase to two-phase
transformation (Clarke’s Transformation) as given in Equations (6.7) and (6.8).

\[
\begin{align*}
\text{(6.7)} \\
i_{sD} &= \frac{2}{3}i_{sA} - \frac{1}{3}i_{sB} - \frac{1}{3}i_{sC} \\
i_{sQ} &= \frac{1}{\sqrt{3}}i_{sB} - \frac{1}{\sqrt{3}}i_{sC} \\
\end{align*}
\]

In this work the zero sequence currents are assumed to be absent.

\[\text{Figure 6.4 Flux model in the rotor reference frame}\]

The direct and quadrature-axis stator currents, which are formulated in the stationary reference frame fixed to the stator (\(i_{sA}, i_{sB}\), \(i_{sC}\)), are then transformed into the two-axis stator current components in the rotor-flux-oriented reference frame (\(i_{sx}, i_{sy}\)), by utilizing the transformation given in Equation (6.9) (Park’s Transformation).

\[
i_{sx} + j i_{sy} = (i_{sD} + j i_{sQ}) e^{-j\beta} \quad (6.9)
\]
The current component $i_{sx}$ serves as an input to a first-order time-delay element with gain 1 and time constant $T_r$, the output of which is the rotor magnetizing current. The quadrature-axis stator current $i_{sy}$ is divided by $i_{mr}T_r$, thus yielding the angular slip frequency of the rotor flux, and when the rotor speed is added to this, finally $\omega_{mr}$ is obtained. Integration of $\omega_{mr}$ yields the angle $\beta$, which defines the position of the rotor-flux space phasor with respect to the real axis of the stationary reference frame. This angle is used in the transformation block $e^{-j\beta}$.

The behavior of the induction machine subjected to rotor-flux-oriented control described above is equal to that of the separately excited d.c machine (Vas 1998).

### 6.5 DESIGN METHODOLOGY

In recent years, motor control employing ASICs/FPGAs is receiving increased attention. This section presents a methodology of modeling, design and simulation of a reusable digital architecture for induction motor vector control, using VHDL simulation targeting FPGA implementation.

As shown in Figure 6.5, top-down methodology is adopted for the design of the vector controller. A library of reusable modules namely, multiplexers, multipliers etc., is developed which in turn is used for the development of next level of modular design and finally these sub-modules are integrated to result in the complete vector control system. The complete drive system was modeled, simulated and evaluated using VHDL.
6.6 SIMULATION RESULTS

Figures 6.6 to 6.11 show the simulation results for individual blocks of the complete vector controller as obtained from Modelsim software. The outputs are numerically verified against the theoretical results.

Figure 6.6 Flux controller – Simulation result
Figures 6.6 to 6.8 are the simulation results of flux, speed and torque controllers which are inherently PI controllers. The flux controller takes the error of magnetizing current component as input and the output is $i_{sxref}$. Similarly the torque and speed controllers are also fed with torque and speed errors as input respectively. The outputs of torque and speed controllers are $i_{syref}$ and $t_{eref}$ respectively.
Figure 6.9 Flux model – Simulation result

Figure 6.9 shows the simulation result of Flux model (Figure 6.4) in which the inputs are the three phase currents namely $i_{sa}$, $i_{sb}$ and $i_{sc}$. The outputs of the flux model are the rotor magnetizing current ($i_{mr}$), the torque-producing stator current ($i_{sy}$) and the spatial position of the rotor flux-linkage space phasor ($\psi_r$). The monitored speed is given as constant in this module. This will be later connected to the actual monitored speed.

Figure 6.10 Inverse Park’s transformation – Simulation result
Figures 6.10 and 6.11 show the simulation results of the two transformations namely inverse Park’s transformation and inverse Clarke’s transformation. Inverse Park’s transformation is realized based on the Equations (6.1) and (6.2). Similarly inverse Clarke’s transformation is realized based on Equations (6.3) to (6.5).

Figures 6.12 to 6.14 show the simulation results obtained for the SVPWM implementation. The patterns obtained for the first three sectors are shown against the variable names pulse_1, pulse_2 and pulse_3. These patterns are for the upper switches in each arm of an inverter while the inverted patterns can be applied to the lower switches.
Figure 6.12 SVPWM output for sector 1

Figure 6.13 SVPWM output for sector 2
6.7 SYNTHESIS RESULTS

Table 6.1 summarizes the space, speed and power results obtained from the synthesis, timing analysis and power analyzer software. The results given are for the individual blocks of the complete vector control system.

Table 6.1 Synthesis results – Modules of vector controller

<table>
<thead>
<tr>
<th>Module Name</th>
<th>Logic Elements required</th>
<th>Clock frequency</th>
<th>Speed (in ns)</th>
<th>Power (in milli Watts)</th>
</tr>
</thead>
<tbody>
<tr>
<td>PI controller</td>
<td>32</td>
<td>50 MHz</td>
<td>13.532</td>
<td>110</td>
</tr>
<tr>
<td>Flux controller</td>
<td>708</td>
<td>50 MHz</td>
<td>46.8</td>
<td>117.59</td>
</tr>
<tr>
<td>Speed Controller</td>
<td>759</td>
<td>50 MHz</td>
<td>114.79</td>
<td>121.21</td>
</tr>
<tr>
<td>Inverse Clarke’s Transformation</td>
<td>277</td>
<td>50 MHz</td>
<td>28</td>
<td>112.39</td>
</tr>
<tr>
<td>Inverse Park’s Transformation</td>
<td>1219</td>
<td>50 MHz</td>
<td>35</td>
<td>117</td>
</tr>
<tr>
<td>Flux Model</td>
<td>1623</td>
<td>50 MHz</td>
<td>138.87</td>
<td>133.22</td>
</tr>
</tbody>
</table>
Kharrat et al (2001) have proved that the vector modulation technique requires only 80 ns for its execution in an FPGA while a DSP executes the same algorithm in 2460 μs. This vector modulation is a part of the complete control system and the total execution time of the complete vector control system will not exceed 500 ns which is very less compared to DSP based implementation.

Hence the proposed methodology greatly improves the speed and space constraints compared to the earlier results proposed in the literature.

6.8 CONCLUSION

The basic modules necessary for vector control of an Induction Machine has been coded in VHDL and simulated. The space, speed and power results of the basic modules are also presented. This work is to mainly throw light on the advantages of FPGA based implementation of the vector control algorithm namely, less use of power and space, shorter design time and greater speed. The other advantage of FPGA based solution is that it is implemented on a single chip, which requires no control-software, leading to more reliability and faster verification time.