3.1 INTRODUCTION

In this chapter the space vector modulated voltage source inverter for direct self control is discussed. The basic schemes of switching to generate space voltage vectors with and without stator resistance for direct self control of induction motor drive are introduced. The discrete pulse modulation and space vector modulation switching of space voltage vectors for controlling the induction motor is also brought out. The Schematic for generating six sided and twelve sided space flux vector polygon is discussed. The author has developed a table to show the effect of space voltage vectors on flux and torque for twelve regions of the instantaneous space vector. This knowledge base is used by the author for fuzzy direct self control of induction motor drive.

3.2 SPACE VECTOR MODULATED VOLTAGE SOURCE INVERTER FOR DIRECT SELF CONTROL OF INDUCTION MOTOR

An inverter is commonly used in variable speed induction motor drives to produce a variable, three phase, and ac output voltage from a constant dc voltage. Since ac voltage is defined by two characteristics, amplitude and frequency, it is essential to work out a strategy that permits control over both these quantities. The space vector concept has led to the development of an inherently digital modulation through controlled switching of space voltage vector. This technique is nowadays commonly known as space vector modulation (SVM). Presently space
vector modulation has become one of the most important PWM methods for three phase converters [1-3].

The schematic diagram of the space vector pulse width modulated voltage source inverter for induction motor drive is shown in Fig. 3.1, where $SA, SB$ and $SC$ are the switching functions with the value “1” when the switch is set to the positive voltage and “0” when the switch is set to the negative voltage. The space vector modulated voltage source inverter output phase voltages are represented by the switching functions $SA, SB$ and $SC$ [1] as follows:

$$V_{w}(SA, SB, SC) = V_{dc} \left(2SA - SB - SC\right)$$

$$V_{b}(SA, SB, SC) = V_{dc} \left(-SA + 2SB - SC\right)$$

$$V_{c}(SA, SB, SC) = V_{dc} \left(-SA - SB + 2SC\right)$$

(3.2.1)

Similarly, the stator voltage can be expressed in a stationary d-axis and q-axis

$$V_{w}(SA, SB, SC) = V_{dc} \frac{V_{dc}}{3} \left(2SA - SB - SC\right)$$

$$V_{q}(SA, SB, SC) = \frac{(V_{b} - V_{c})}{\sqrt{3}} = V_{dc} \frac{V_{dc}}{\sqrt{3}} \left(SB - SC\right)$$

(3.2.2)

Fig. 3.1 Space vector pulse width modulated voltage source inverter for induction motor drive.
To analyses three phase quantities as a whole, the three phase induction motor voltages and currents are represented by an instantaneous space voltage vector $V_s$ and an instantaneous space current vector $I_s$ respectively [2].

The instantaneous space voltage vector $V_s$

$$V_s = \sqrt{3}(V_a + V_b e^{i\pi/3} + V_c e^{i2\pi/3})$$  \hspace{1cm} (3.2.3)

The instantaneous space current vector $I_s$

$$I_s = \sqrt{3}(i_a + i_b e^{i\pi/3} + i_c e^{i2\pi/3})$$  \hspace{1cm} (3.2.4)

where, $V_a$, $V_b$, $V_c$ are the instantaneous values of the primary line-to-neutral voltages, and $i_a$, $i_b$, $i_c$ are the instantaneous values of the primary line to neutral current as determined by equation (3.2.6) to (3.2.8).

The instantaneous space voltage vectors for the inverter are then determined by the status of switching functions $S_A, S_B, S_C$. Using these switching functions, the primary space voltage vector $V_i(S_A, S_B, S_C)$ of induction motor can be represented by [2]

$$V_i(S_A, S_B, S_C) = \sqrt{3}(S_A + S_B e^{i\pi/3} + S_C e^{i2\pi/3})$$  \hspace{1cm} (3.2.5)

where, $V_{dc}$ is the dc link voltage of the inverter.

There are eight primary space voltage vectors as shown in Fig.3.2. Six space voltage vectors are of equal magnitude and they are 60° space displaced from each other. The remaining two are of zero magnitude [2, 3, 8]. The space voltage vectors are determined by the switching functions $S_A, S_B$ and $S_C$ as shown in Table 3.1. Fig.3.3 shows the eight switching states for generating space vectors [4, 8].

The nonzero space voltage vectors ($V_i - V_s$) control the instantaneous stator current and stator flux to provide the fast dynamic response to induction motor. The trajectory of stator flux moves in the direction of primary space voltage vector $V_i$ depending upon the
Fig. 3.2 Eight primary space voltage vectors.

Table 3.1 Space voltage vectors and switching functions.

<table>
<thead>
<tr>
<th>Space voltage vectors</th>
<th>SA</th>
<th>SB</th>
<th>SC</th>
</tr>
</thead>
<tbody>
<tr>
<td>V₀</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>V₁</td>
<td>1</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>V₂</td>
<td>1</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>V₃</td>
<td>0</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>V₄</td>
<td>0</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>V₅</td>
<td>0</td>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td>V₆</td>
<td>1</td>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td>V₇</td>
<td>1</td>
<td>1</td>
<td>1</td>
</tr>
</tbody>
</table>
Fig. 3.3 Eight switching states of space voltage vectors
switching sequence of nonzero space voltage vectors. Therefore, applying a nonzero space voltage vector to an induction motor drive, the direction and amplitude of stator flux can be changed. So, by appropriately selecting the six nonzero space voltage vectors and their time duration, the trajectory of flux can be controlled to follow a specified locus.

The stator current can also be derived using the switching states and the dc link current as follows [1]:

\[ i_a = i_{dc} \quad \text{when} \quad (SA, SB, SC) = (1,0,0) \]
\[ i_a = -i_{dc} \quad \text{when} \quad (SA, SB, SC) = (0,1,1) \]  \hspace{1cm} (3.2.6)
\[ i_s = i_{dc} \quad \text{when} \quad (SA, SB, SC) = (0,1,0) \]
\[ i_s = -i_{dc} \quad \text{when} \quad (SA, SB, SC) = (1,0,1) \]  \hspace{1cm} (3.2.7)
\[ i_r = i_{dc} \quad \text{when} \quad (SA, SB, SC) = (0,0,1) \]
\[ i_r = -i_{dc} \quad \text{when} \quad (SA, SB, SC) = (1,1,0) \]  \hspace{1cm} (3.2.8)

From the above, it is obvious that one phase current is related to the dc link current at a given switching instant.

The stator current in the d-q frame can be determined from the phase currents as follows [2]:

\[ i_d = i_a, \quad i_q = \frac{2i_a + i_s}{\sqrt{3}} \]
\[ \text{or} \quad i_d = i_a, \quad i_q = -\frac{2i_s + i_a}{\sqrt{3}} \]
\[ \text{or} \quad i_d = -(i_s + i_a), \quad i_q = \frac{i_s - i_a}{\sqrt{3}} \]  \hspace{1cm} (3.1.9)

The two most recently sampled phase currents are to be used to update \( i_d \) and \( i_q \).
3.3 BASIC SCHEME OF HYSTERESIS SWITCHING TO GENERATE SPACE VOLTAGE VECTORS IGNORING STATOR RESISTANCE

Fig. 3.4 shows the basic scheme of hysteresis switching to generate space voltage vectors ignoring stator resistance for direct self control inverter fed induction motor drives. The $V_{dc}$ voltage is applied across inverter. The Schmitt trigger (ST) circuit is used to generate switching signals [5].

The shaft speed of a three phase induction motor mainly depends on the angular velocity of the rotating magnetic field. In steady state this velocity is determined by the frequency $f_s$ of stator supply and the magnitude of the magnetic field depends on the voltage to frequency ratio. The value of flux reference $\psi_{ref}$ determines the magnitude of the magnetic field. The frequency depends on the $V_{dc}$ voltage and flux reference $\psi_{ref}$ [5].

The three phase flux linkages are generated by the time integral of phase voltages as follows:

$$
\psi_{\beta a} = \int e_{\beta a} dt
$$

$$
\psi_{\beta b} = \int e_{\beta b} dt
$$

$$
\psi_{\beta c} = \int e_{\beta c} dt
$$

(3.3.1)

where, $e_{\beta a} = \frac{e_{ba}}{\sqrt{3}}$, $e_{\beta b} = \frac{e_{ca}}{\sqrt{3}}$ and $e_{\beta c} = \frac{e_{ab}}{\sqrt{3}}$.

The estimated flux linkages are compared with reference flux ($\psi_{ref}$) and are fed to ST as indicated in the schematic diagram.

At steady state, the reference flux and frequency depend on dc voltage. The relationships between dc voltage with reference flux and frequency are as follows:

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\( e_{\beta c} = \sqrt{3} e_{\beta a}, \quad e_{\alpha a} = \sqrt{3} e_{\beta b}, \quad e_{\alpha b} = \sqrt{3} e_{\beta c} \)

Fig. 3.4 Basic scheme of hysteresis switching to generate space voltage vectors ignoring stator resistance.
The speed of induction motor can be controlled by varying the dc voltage $V_{dc}$ as per the equation (3.3.2). To raise the speed above the rated speed at given dc voltage and rated flux, the value of the flux reference $\psi_{ref}$ has to be lowered. If all terminal of the induction motor are connected simultaneously to either of the dc terminals of the inverter, all the phase voltages of the induction motor will be zero. There two switching sequence can be generated with the help of an additional modulating device [5].

The hysteresis controller as shown in Fig. 3.4 though simple to implement and is of dynamic quality, this controller produces considerable flux distortions owing to the following drawbacks:

- There is no interaction between the individual hysteresis controllers of three phases and, hence, no strategy to generate the zero voltage vector.

- There is a tendency at lower speed to lock into a limit cycle of high-frequency switching sequences comprising only nonzero voltage vectors.

- The flux error is not strictly limited.

Fig.3.5 shows the track curve of space vector $\psi_s$ with time. The space vector $\psi_s$ can be expressed as a function of three phase flux linkages, $\psi_{\beta_a}$, $\psi_{\beta_b}$ and $\psi_{\beta_c}$ as follows [5]:

$$\psi_s = \frac{2V}{\sqrt{3}T} \frac{T_s}{3}$$

$$f_s = \frac{V_{dc}}{\sqrt{3} 0\psi_{ref}}$$

(3.3.2)

where $T_s$ is the period of stator quantities.

$$\psi_s = \frac{2V}{\sqrt{3}} \frac{T_s}{3}$$

$$f_s = \frac{V_{dc}}{\sqrt{3} 0\psi_{ref}}$$

(3.3.2)

where $T_s$ is the period of stator quantities.

$$\psi_s = \frac{2V}{\sqrt{3}} \frac{T_s}{3}$$

$$f_s = \frac{V_{dc}}{\sqrt{3} 0\psi_{ref}}$$

(3.3.2)
Fig. 3.5 Space vectors of stator voltages, total fluxes, and track curves with DSC.
\[ \frac{1}{3} (\psi_{aa} + a^2 \psi_{ab} + a \psi_{ac}) \]  

(3.3.3.b)

where, \( a = e^{2\pi i} \), \( \alpha \) subscript refers to line to neutral quantities, and \( \beta \) subscript refers to line to line quantities.

Usually the representation given by equation (3.3.3.b) is followed, if the lines to neutral voltages are chosen as input quantities. The time integrals of these voltages provides the \( \psi_{aa}, \psi_{ab} \) and \( \psi_{ac} \) fluxes. According to

\[ \dot{\psi}_s = V_s, \]  

(3.3.4)

\( |V_s| \) determines the tracking speed with which the head of \( \psi_s \) traverses its track curve. That means tracking speed depends only on the instantaneous value of \( V_{ac} \). To stop the motion of \( \psi_s \), the zero value space voltage vector \( V_s(0) \) is realized during a part of the pulse period. The direction of \( V_s \) gives the direction of the straight lines forming the track curve of \( \psi_s \). With DSC the distance of the straight parts of the track curve from the tail of \( \psi_s \) is determined by \( \psi_{ref} \).

The projections of the moving \( \psi_s \) on the three \( \beta_a, \beta_b \) and \( \beta_c \) axis are given by \( \psi_{\beta_a}, \psi_{\beta_b} \) and \( \psi_{\beta_c} \). These fluxes are compared to \( \pm \psi_{ref} \) to generate the switching functions \( SA, SB, SC \). With \( \psi_{ref} \) constant, the tracks curve of \( \psi_s \) forms a regular hexagon. Its center point coincides with the origin of the complex plane. At constant dc voltage and without pulse control, the track speed of \( \psi_s \) remains constant, and magnitude and angular speed of this space vector are slightly pulsating [5].

In this way direct self control of induction motor drives, the switching functions \( SA, SB, SC \) are generated from error in phase fluxes and the flux references. The extremity of space flux vector follows the hexagonal trajectory. There is no control on electromagnetic torque in this scheme.
3.4 BASIC SCHEME OF HYSTERESIS SWITCHING TO GENERATE SPACE VOLTAGE VECTORS BY TAKING INTO ACCOUNT STATOR RESISTANCE

Fig. 3.6 shows the basic scheme of hysteresis switching to generate space voltage vectors by taking accounts stator resistance for direct self control inverter fed induction motor drive [5, 6]. The signal processing of DSC is rather different from those methods directly controlling torque and stator flux. The stator flux and torque can be derived from phase currents and voltages. The control strategy of direct self control of inverter fed induction motor is described in the following paragraph wherein the role of the flux estimation, torque estimation and speed estimation are discussed with appropriate details.

3.4.1 Flux Estimation

In space vector representation the stator flux space vector $\Psi_s$ can be expressed as

$$\Psi_s = \frac{1}{2} \left[ \Psi_{s\alpha\alpha} + j \Psi_{s\beta\alpha} \right]$$

$$= |\Psi_s| \angle \theta$$

(3.4.1)

where, $\theta$ is the stator flux angle with respect to the d axis of the stationary d-q reference frame and is given by

$$\theta = \tan^{-1} \left( \frac{\Psi_{s\beta\alpha}}{\Psi_{s\alpha\alpha}} \right)$$

(3.4.2)

The stator flux in q-axis and d-axis can be obtained on integration of the phase voltage minus voltage drop in stator resistance [5-6].

$$\Psi_{s\alpha\alpha} = \int (e_{\alpha m} - R_s i_{\alpha m}) dt$$

$$\Psi_{s\beta\alpha} = \int (e_{\beta m} - R_s i_{\beta m}) dt$$

(3.4.3)

where, $e_{\alpha m} = e_{\alpha}$ and $i_{\alpha m} = i_{\alpha}$

$$e_{\beta m} = \frac{e_{\beta}}{\sqrt{3}}$$

and $i_{\beta m} = \frac{i_{\beta}}{\sqrt{3}}$

(3.4.3a)
The instantaneous values of three phase quantities are the line currents $i_a, i_b, i_c$ and the line to neutral voltages $e_a, e_b, e_c$.

Since all the stator flux projections $\psi_{s,\alpha}, \psi_{s,\beta}, \psi_{s,b}$ are required to generate switching functions $S_A, S_B, S_C$ the remaining fluxes $\psi_{s,\alpha}, \psi_{s,\beta}, \psi_{s,b}$ can be computed making use of the expressions given below [6]:

$\psi_{s,\beta} = 0.5 \psi_{s,\alpha} + 0.5 \sqrt{3} \psi_{s,\beta}$

$\psi_{s,\alpha} = 0.5 \psi_{s,\alpha} - 0.5 \sqrt{3} \psi_{s,\beta}$

(3.4.4)

wherein $\psi_{s,\alpha}$ is already available with the help of equation (3.4.3)

The control strategy for generating stator space flux vector $\psi_s$ to flow a hexagonal track curve is used with the help of appropriate switching functions as explained below:

The switching state of space vector modulated inverter has to be changed only when $\psi_s$ reaches a corner of the hexagonal track curve. Each change in direction of the flux track requires switching off of one switching element of the inverter and switching on of another one. This is called one complete switching operation, which is abbreviated as one switching function. The hexagonal track curve of $\psi_s$ demands only six different types of switching per period. Therefore, minimal numbers of six switching functions per period in practical applications are required. The additional change in direction of the stator flux track curve could be incorporated by adding additional switching function.

4.2 Torque Estimation

The electromagnetic torque can be expressed in terms of the direct and quadrature axis components of stator flux and stator current as follows [6]:

$$T_e = \frac{3P}{4} (\psi_{s,\alpha}i_{\beta} - \psi_{s,\beta}i_{\alpha})$$

(3.4.5)

where, $P$ is the number of poles of the machine.

In the TC unit of Fig.3.6, the electromagnetic torque $T_e$ is calculated according to the expression (3.4.5). This is compared with the torque reference value $T_{ref}$ to obtained error in
Fig. 3.6 Basic scheme of hysteresis switching to generate space voltage vectors by taking accounts stator resistance.

I - Integrator
CT - Coordinate transformation
ZS - Zero state select
TC - Torque control
torque $\Delta T$. In the Schmitt trigger $ST_T$ the error in torque is compared with the desired accuracy of torque constancy tolerance value of torque $\varepsilon_T$. When the absolute value of error in torque is more than $\varepsilon_T$, the switch of ESS symbolically depicted in Fig.3.6 is de-actuated by the two-point torque controlling Schmitt trigger. All switching commands $SA, SB, SC$ get the same value $S_2$, given by the zero-state select unit $ZS$. The result is that all terminals of the induction motor are commonly switched onto one of the two dc terminals of the inverter, which leads to $e=0$. This causes the stator flux space vector $\psi_s$ to halt on its track curve, whereas the rotor flux space vector $\psi_r$ continues its movement with approximately constant angular speed. This causes a decrease of the flux angle and, as a consequence a decrease in torque $T_r(t)$. When instantaneous value of torque falls below $T_{ref}$ by more than $\varepsilon_T$, the full voltage is switched again onto the induction motor via the two-point torque control. One pulse period is completed when the next zero voltage state is switched on. In this way, the mean value of tracking speed of $\psi_s$, taken over one pulse period can be varied between zero and full scale. It is automatically adjusted so that the required torque is produced by self control independent of rotor speed, stator resistance, or variations of the dc voltage $V_d$. If the direction of the flux track curve is not changed within a pulse period, only the switching state of one inverter phase has to be changed twice. This results in only two switchings per pulse period.

During a complete run around the hexagonal track curve, the direction of the path has to be changed in only six pulse periods. Therefore, only six pulse periods within a complete run need three switchings. If a circular track curve has to be approached, each pulse period requires three switchings. The switching rate of the two-level torque controller adjusted by the width of the hysteresis-band $2\varepsilon_T$ is controlled so that the allowed inverter switching frequency is fully utilized. The possible two values of $S_2$ are chosen by the zero select unit $ZS$ such that the number of needed switchings is minimal. In this way, time optimal control torque is achieved. Although this method of direct self control is convenient for practical implementation, it has the draw back of variable switching frequency in the control schematic.
3.4.3 Speed Estimation [7]

In torque controlled drives, it is necessary to close an outer speed loop. Therefore, estimation of synchronous speed or frequency and often, rotor speed becomes important.

The synchronous frequency can be found by noticing that the angle of the stator flux can be expressed as

\[ \theta = \tan^{-1}\left( \frac{\psi_{s\beta}}{\psi_{s\alpha}} \right) \]  

This implies that the synchronous frequency can be found to be

\[ \omega_s = \frac{d\theta}{dt} = \frac{d}{dt} \left( \tan^{-1}\left( \frac{\psi_{s\beta}}{\psi_{s\alpha}} \right) \right) \]  

The above equation can be rewritten using (3.4.2)

\[ \omega_s = \frac{(e_{s\beta} - R_s i_{s\beta}) \psi_{s\alpha} - (e_{s\alpha} - R_s i_{s\alpha}) \psi_{s\beta}}{\left(\psi_{s\alpha}\right)^2 + \left(\psi_{s\beta}\right)^2} \]  

Slip speed can be derived using the steady state torque speed curve for the induction motor and is given the relationship:

\[ \omega_s = K_s T_e \]  

where, \( K_s = \) rated slip frequency/rated torque and can be derived from the rated values of induction motor parameters. Accordingly the rotor speed is given by

\[ \omega_r = \omega_s - \omega_s. \]  

3.5 SCHEMATIC DETAILS OF SPACE VECTOR MODULATION SWITCHING TO GENERATE SPACE VOLTAGE VECTOR

The direct torque and flux control of induction motor using space vector modulation has the advantage of a fixed switching period. The inverter switching pattern or duty cycle is directly calculated in order to control the torque and flux in a deadbeat fashion over a constant switching period. This is accomplished by calculating the voltage space vector required to control the torque and flux on a cycle by cycle basis using the calculated flux and torque errors sampled from the previous cycle and an estimate of the back emf in the induction motor. The back emf of the induction motor is estimated from the flux and voltage vectors [8]. Voltage
space vector PWM is then used to find the switching intervals or duty cycle. A schematic block diagram of the direct torque and flux controlled induction motor is shown in Fig. 3.7.

An equivalent circuit in the stationary reference frame of the space vector modulated inverter driven induction motor is shown in Fig. 3.8. From Fig. 3.8, the change in the stator current vector $\Delta \bar{I}_s$ over a constant period $T_s$ is given by

$$\Delta \bar{I}_s = \frac{(\bar{V} - \bar{E})T_s}{L_s}$$  \hspace{1cm} (3.5.1)

The period $T_s$ is constant in this space vector modulation scheme in order to maintain a constant switching frequency. It is assumed that the stator electrical time constant is much longer than $T_s$ and therefore, the change in current over the period $T_s$ is linear. The corresponding change in electromagnetic torque over the period $T_s$ is then

$$\Delta T = \frac{3}{2} \frac{P}{2} \left( \frac{\bar{\psi}_s \times \Delta \bar{I}_s}{\psi_s} \right)$$  \hspace{1cm} (3.5.2.a)

$$= \frac{3}{2} \frac{P}{2} \left( \psi_s \times \frac{(V - E)T_s}{L_s} \right).$$  \hspace{1cm} (3.5.2.b)

Therefore, the change in torque over a period can be predicted from the stator voltage and current and the voltage behind the transient reactance $\bar{E}$. This voltage can be estimated from the stator flux and current.

From the equivalent circuit the voltage behind the transient reactance of the induction motor is

$$\bar{E} = \bar{V}_s - R\bar{I}_s - \frac{d}{dt}(L_s\bar{I}_s)$$  \hspace{1cm} (3.5.3.a)

$$= \frac{d}{dt}(\bar{\psi}_s - L_s\bar{I}_s).$$  \hspace{1cm} (3.5.3.b)

If it is assumed that $\bar{E}$ is sinusoidal, then $\bar{E} = j\omega_s(\bar{\psi}_s - L_s\bar{I}_s).$  \hspace{1cm} (3.5.4)

The excitation frequency $\omega_s$ in equation (3.4.4) can be estimated using the stator flux vector and the terminal quantities [7] as follows:

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Fig. 3.7 Schematic block diagram of direct torque and flux controlled induction motor using space vector modulation.
Fig. 3.8 Equivalent circuit of inverter driven induction motor in the dq stationary reference frame.

Fig. 3.9 Vector diagram illustrating the state selection process for transient operation.
Here, it is assumed that $\vec{\psi}_s$ is sinusoidal.

The change in flux over the period $T_s$ is given by

$$\Delta \psi_s = (\vec{V}_s - R_l\vec{I}_s)T_s$$

$$= \vec{V}T_s. \quad (3.5.6)$$

The space vector modulation control scheme is described in the following paragraphs that use the estimated or predicted values of the change in flux and torque to determine the switching state of the inverter.

### 3.5.1 Deadbeat Torque and Flux Control [8]

The space vector modulation control scheme is based on a predictive calculation of the stator voltage vector, which will drive the torque and flux magnitude to the reference value in a deadbeat fashion over a constant period. A fixed time period equal to the switching period is denoted by $T_s$. Let $t_s$ be the time at the beginning of an arbitrary $T_s$ period. The torque is controlled in a beat manner if the change in electromagnetic torque over the $T_s$ period is given by

$$\Delta T = T' - T(t_s)$$

$$= \frac{3}{2} \left( \psi_s \times \frac{(\vec{V}' - \vec{E})T_s}{L_s} \right). \quad (3.5.7.a)$$

where $T'$ is the reference or command value of torque and $\vec{V}'$ is the voltage vector which results in deadbeat control of the torque. In terms of d and q components this change in torque can be expressed as

$$\Delta T = \frac{3PT_s}{4L_s}((-\psi_d E_q + \psi_q E_d) + (\psi_d V_q' - \psi_q V_d')) \quad (3.5.7.b)$$
Let $K_e$ be defined as
\[
K_e = \frac{4}{3} \Delta T \frac{L}{P T_s} \theta + (\psi_{ad} E_d - \psi_{aq} E_q),
\]
(3.5.9)

then, the $q$ component of voltage is given by
\[
V_q^* = \frac{K_e + \psi_{ad} V_d^*}{\psi_{ad}}
\]
(3.5.10)

The stator flux magnitude is controlled such that the change in stator flux is given by
\[
\Delta \psi = \psi_i^* - |\psi(t_s)|.
\]
(3.5.11)

The $\psi_i^*$ is the command value of stator flux magnitude. Using (3.5.6) the voltage $V^*$ required for deadbeat control of the flux magnitude is found from
\[
\begin{align}
V_i^* &= |\tilde{V}_i + \psi_i^*(t_s)| \\
(\psi_i^*)^2 &= (V_q^* T_s + \psi_i^*)^2 + (V_d^* T_s + \psi_i^*)^2
\end{align}
\]
(3.5.12.1)\(3.5.12.2)

where, $\tilde{\psi}_i$ is assumed to be the value of flux vector at the beginning of the period $T_s$. Equations (3.5.7b) and (3.5.12b) represent two equations with two unknowns $V_q^*$ and $V_d^*$. Substituting (3.5.10) into (3.5.12b) gives a quadratic equation with $V_d^*$ as the unknown.
\[
\begin{align}
\left(T_s^2 + \frac{\psi_{ad}^2 T_s^2}{\psi_{ad}^2}\right) V_d^* + \left(2 \frac{K_e \psi_{ad} T_s^2}{\psi_{ad}^2} + 2 \psi_{ad} T_s + \frac{2 \psi_{ad}^2 T_s}{\psi_{ad}^2}\right) V_d^* \\
+ \frac{T_s^2 K_e^2}{\psi_{ad}^2} + \frac{2 \psi_{ad} K_e T_s}{\psi_{ad}^2} + \psi_{ad}^2 + \psi_{ad}^2 - \psi^2 = 0
\end{align}
\]
(3.5.13)

Equation (3.5.13) yields two solutions for $V_d^*$. The solution with the smallest absolute value is chosen since it represents the smallest $d$-axis voltage necessary to drive the torque and flux to their reference values. Equation (3.5.10) then gives $V_q^*$. The $\tilde{V}^*$ can be obtained from
\[
V^* = \sqrt{(V_q^*)^2 + (V_d^*)^2}
\]
(3.5.14)
Once \( \tilde{V}^* \) is known, the reference value of stator voltage is obtained by adding on the stator IR drop, that is
\[
\tilde{V}^*_s = \tilde{V}^* + R_f I_r(t_k).
\] (3.5.15)
where, \( \tilde{V}^*_s \) represents the average value of the stator voltage vector over the period \( T_s \).

The switching state of the inverter is determined from \( V_s \) using space vector modulation in article 3.2. This reference voltage is obtained by computing the duty ratio for the two voltage vectors \( \tilde{V}_{s(k)} \) and \( \tilde{V}_{s(k+1)} \), which are adjacent to \( \tilde{V}^*_s \). Let \( T_k \) and \( T_{k+1} \) be amount of time spent on \( \tilde{V}_{s(k)} \) and \( \tilde{V}_{s(k+1)} \), respectively. Then, \( T_k \) and \( T_{k+1} \) are found by solving
\[
\tilde{V}^*_s T_k = \tilde{V}_{s(k)} T_k + \tilde{V}_{s(k+1)} T_{k+1}.
\] (3.5.16)

The remainder of the switching period is spent on the zero state (\( k=0 \) or 7) and is given by
\[
T_0 = T_s - T_k - T_{k+1}
\] (3.5.17)
where, \( T_0 \) is the time spent on the zero state (\( k=0 \) or 7). Switching from the zero state to two adjacent states involves commutating each of the inverter legs exactly once. Therefore, \( T_s \) is a half period of the switching frequency. This implies that the torque and flux are controlled twice per switching cycle.

### 3.5.2 Torque and Flux Control under Transient Conditions

The dynamic response of the direct self control induction motor drive will be effected if the transients occurs in the torque command or flux command or in both of these commands. The appropriate control strategy is to be followed to minimize their effect on the dynamic response of the induction motor. With a sufficiently large torque error in one switching period, such that which occurs with a step change in flux command, the simultaneous solution of (3.5.16) and (3.5.17) yields the sum \( T_k + T_{k+1} \) greater than \( T_s \), that is \( \tilde{V}^*_s \) is too large to be synchronized in a single switching period. Therefore, an alternative control method must be derived. In the case of a transient in the torque command, the controller must cause the torque to
be driven toward its reference value while still maintaining deadbeat control of the flux. Just the opposite is true in the case of a flux reference transient. Then, the flux magnitude is driven toward the reference value while maintaining deadbeat control of the torque.

3.5.2.1 Torque Transient [8]

Consider the case of a transient condition in the torque, that is, the torque cannot be driven to the reference value in a single $T_s$ period. This is generally the most important and most frequently encountered transient condition. In this case, the inverter states are chosen a priori, which causes the torque to be driven in the desired direction while still allowing for deadbeat control of flux. Let us consider a condition where the angle of the flux vector is between $\pi/6$ and $\pi/6$ as shown in the vector diagram of Fig.3.9. States 2 and 3 cause the current in quadrature with the stator flux, and therefore the torque, to increase. Likewise, states 5 and 6 cause the torque to decrease. In addition, from Fig.3.9, states 2 and 6 cause the flux magnitude to increase, and states 3 and 5 cause the flux magnitude to decrease. Therefore, states 2 and 3 can be used to control the flux to its reference value over the $T_s$ interval while continuously increasing the torque over the entire interval. Table 3.2 summarizes the state selection for a transient in the torque command. Once the states $k$ and $k+1$ are identified, the flux is controlled by substituting $\vec{V}_{x(k)}$ and $\vec{V}_{x(k+1)}$ into (3.5.12a), that is

$$\psi_i^* = |\vec{V}_{x(k)}T_s + \vec{V}_{x(k+1)}T_s + \vec{\psi}_i(t_s)|$$

(3.5.18)

Since it is desired to drive the torque in one direction as quickly as possible, the zero state is not used under transient conditions. Therefore

$$T_s = T_s + T_{s+1}.$$  

(3.5.19)

Equations (3.4.18) and (3.4.19) are solved simultaneously for $T_s$ and $T_{s+1}$. In this way, the flux is controlled to its reference value, and the torque is driven continuously in the proper direction with maximum voltage applied to the induction motor.
Table 3.2 Selection of inverter states $k$ and $k+1$
under torque transient condition

$$(2n-3)/6 < \theta < (2n-1)/6$$

<table>
<thead>
<tr>
<th>$\mathrm{Sgn}(T - T')$</th>
<th>$K$</th>
<th>$K+1$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>$n+1$</td>
<td>$n+2$</td>
</tr>
<tr>
<td>1</td>
<td>$n+4$</td>
<td>$n+5$</td>
</tr>
</tbody>
</table>

3.5.2.2 **Flux Transient** [8]

Now, consider the case of a transient in the flux command, that is, the flux magnitude cannot be driven to the reference value in a single $T_s$ period. In this case, the state selection is again made a priori. Using the condition as shown in Fig.3.9, states 1 and 2 both increase flux, and states 3 and 4 both decrease the flux. The determination of $k$ and $k+1$ for controlling the torque while during a flux transient is also given in Table 3.3. The torque is controlled by substituting $\vec{V}_{stk}$ and $\vec{V}_{stk+1}$ into (3.5.7) that is

$$\Delta T = \frac{3}{2} \frac{1}{L_s} P (\bar{\psi}_r (\vec{V}_{stk} T_s + \vec{V}_{stk+1} T_{s+1} - \bar{E} T_s))$$

Equation (3.5.20) is solved simultaneously with (3.5.19) for $T_s$ and $T_{s+1}$. This accomplishes the deadbeat control of the torque while driving the flux in the desired direction. Here, again, the zero state is not used. This is perfectly reasonable since skipping the zero state while still switching between adjacent states is equivalent to pulse dropping in conventional sinusoidal modulation.

Table 3.3 Selection of inverter states $k$ and $k+1$
under torque transient condition

$$(2n-3)\pi/6 < \theta < (2n-1)\pi/6$$

<table>
<thead>
<tr>
<th>$\mathrm{Sgn}(\psi_r - \psi_r')$</th>
<th>$K$</th>
<th>$K+1$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>$n+1$</td>
<td>$n+2$</td>
</tr>
<tr>
<td>0</td>
<td>$n+3$</td>
<td>$n+4$</td>
</tr>
</tbody>
</table>
3.5.2.3 Torque Transient and Flux Transient [8]

The only remaining possibility is a transient in both torque and flux. In this case a single state is selected for the entire period, which drives both flux and torque in the desired direction as quickly as possible. The state selection is given in Table 3.4. The selected states cause both the torque and flux to be driven in the proper direction.

Table 3.4 Selection of inverter states $k$ and $k+1$
under torque transient condition

\[
(2n-3)\pi/6 < \theta < (2n-1)\pi/6
\]

<table>
<thead>
<tr>
<th>$\text{Sgn}(T - T')$</th>
<th>$\text{Sgn}(\psi - \psi')$</th>
<th>$K$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
<td>n+1</td>
</tr>
<tr>
<td>0</td>
<td>1</td>
<td>n+2</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
<td>n+4</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>n+5</td>
</tr>
</tbody>
</table>

Inverter state, $k = \begin{cases} 
K & K \leq 6 \\
K-6 & K > 6 
\end{cases}$

3.6 SCHEMATIC DETAILS OF DISCRETE PULSE MODULATION SWITCHING TO GENERATE SPACE VOLTAGE VECTOR

The discrete pulse modulation switching of space voltage vectors to generate six sided and twelve sided space flux vector polygon is discussed below:

3.6.1 Schematic for generating six sided space flux vector polygon

Fig. 3.10 shows the schematic diagram of discrete pulse-modulated to direct self control inverter fed induction motor drive. This technique is based on the stator flux field orientation principles. The control scheme is discussed in detail in the following paragraphs:
Fig. 3.10 Schematic block diagram of discrete pulse-modulated direct self controlled stator flux and torque regulator.
3.6.1.1 Stator Flux Field Orientation Details [11]

Stator flux field-orientation schemes are designed around a current regulated inverter whose reference current is obtained from the torque and flux magnitude references. Calculation of the current reference from the torque and flux magnitude references is sensitive to induction motor parameters and the counter EMF.

The stator space current vector \( I_s \) can be expressed in stationary reference frame in terms of stator currents \( i_a, i_b \) and \( i_c \) as follows:

\[
I_s = i_a + j i_b = i_a + j I_x + \frac{2 I_y}{\sqrt{3}}
\]  

(3.6.1)

The stator flux can be written in terms of the stator line neutral voltages \( V_a, V_b, \) and \( V_c \) and the stator \( IR \) drop as

\[
\psi_{a,b,c} = \int (V_{a,b,c} - I_{a,b,c} R) dt
\]  

(3.6.2)

The stator flux space vector is therefore given by

\[
\psi_s = \int (V_s - I_s R) dt = \psi_a + j \psi_b
\]  

(3.6.3.a)

(3.6.3.b)

This computation is shown on the upper left corner of the block diagram 3.7.

The electromagnetic torque produced by the machine is given by

\[
T = \frac{3P}{4} |\psi_s \times I_s|
\]  

(3.6.4.a)

\[
= \frac{3P}{4} (\psi_a i_b - \psi_b i_a)
\]  

(3.6.4.b)

This is represented in the schematic diagram by the block diagram indicated as torque calculator.

In terms of \( abc \) quantities, the torque can be expressed as

\[
T = \frac{\sqrt{3}P}{2} (\psi_a i_b - \psi_b i_a)
\]  

(3.6.5.a)

\[
= \frac{\sqrt{3}P}{2} (\psi_s i - \psi_s i_c)
\]  

(3.6.5.b)
The torque can also be expressed in terms of the magnitude of the stator flux and the component of stator current in quadrature with the stator flux \( I_r \) as follows

\[
T = |\psi_s| I_r
\]  

(3.6.6)

Since the stator voltage is completely determined by the dc link voltage and the inverter switching state \( k \), the stator flux can be directly controlled by inverter modulation. If the stator resistance is known, measured, or estimated, the value of the stator flux vector is the result of integration as given in equation (3.6.3). The change in the stator flux from the value at the present switching instant to the value at the next switching instant can be approximated by

\[
\Delta \psi_s = (V_s - I_s R_s) T_s
eq V_s T_s \tag{3.6.7.a}
\]

\[
\begin{cases} 
\frac{2}{3} V_s e^{(k-1) \frac{2\pi}{3}} T_s & k=1,\ldots,6. \\
0 & k=0,7 
\end{cases} \tag{3.6.7.b}
\]

3.6.1.2 Direct Flux and Torque Control [11]

The direct self control scheme (DSC) as presented in [1-10] uses a hysteresis band to directly control the flux and torque in the induction motor. In discrete pulse-modulated technique, however, a zero hysteresis sliding mode control scheme is derived for use with discrete pulse switching. Here, a single space nonzero voltage vector is switched on depending on the angle of stator space flux vector to increase the torque and the space zero voltage vector to decrease the torque. The torque is controlled in this manner with in some hysteresis band. When the magnitude of stator flux reaches the reference value, the space nonzero voltage vector is incremented. The flux then traces the well known hexagonal flux path with in the bounds of reference flux giving substantial distortion components in the induction motor current. In this way this method is used to choose nonzero voltage vectors that increase the torque and allow the
flux magnitude to be controlled such that it is nearly constant. The method of state selection is used in this scheme described below.

When the stator flux vector angle, $\theta$ is within $(2n-3)\pi/6$ and $(2n-1)\pi/6$, the flux magnitude will increase by choosing switching states $k = n, n+1, \text{and } n+2$. If $k > 6$, then the actual switching state selected is $(k-6)$. The other three switching states will cause the flux magnitude to decrease. The zero switching state causes the flux vector to remain nearly constant over the switching period. In general for $(2n-3)\pi/6 < \theta < (2n-1)\pi/6$, the switching state is selected according to Table 3.6.

Table 3.5 State selection for discrete pulse modulation DSC control

<table>
<thead>
<tr>
<th>$(2n-3)\pi/6 &lt; \theta &lt; (2n-1)\pi/6$</th>
<th>$\text{Sgn} (T - T')$</th>
<th>$\text{Sgn} (\psi_s - \psi_s')$</th>
<th>$K$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
<td>n+1</td>
<td></td>
</tr>
<tr>
<td>0</td>
<td>1</td>
<td>n+2</td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>Don't care</td>
<td>0</td>
<td></td>
</tr>
</tbody>
</table>

Inverter state, $k = \begin{cases} K & K \leq 6 \\ K-6 & K > 6 \end{cases}$

3.6.2 Schematic for Generating Twelve Sided Space Flux Vector Polygon

The direct stator flux and torque control, made possible by application of instantaneous space vector control theory, has recently become popular in induction motor variable-speed drives. With the instantaneous space vector theory, the stator instantaneous flux can be calculated and kept nearly constant. Moreover, since the quick accelerating or decelerating rotation of the stator flux is achieved, rapid torque response control can be obtained. The locus of the integrated value of the instantaneous space voltage vectors makes it very easy to
understand the movement of the stator flux and of its calculation. The stator space flux vector of induction motor can be calculated by [2]

\[
\psi_s = \int (V_s(SA,SB,SC) - R_s I_s)dt \\
= \psi_s(0) + V_s(SA,SB,SC)\Delta T - \int R_s I_s dt \tag{3.6.8}
\]

where, \(\psi_s(0)\) stator flux vector at \(t=0\)  
\(I_s\) stator current vector  
\(R_s\) stator winding phase resistance  
\(\Delta T\) time width of \(V_s\)  
\(V_s\) space voltage vector  
\(SA,SB,SC\) switching function

Assuming that the voltage drop of stator winding resistance is small, the trajectory of \(\psi_s\) moves in the direction of \(V_s(SA,SB,SC)\), and the velocity of \(\psi_s\) is proportional to the value of \(V_s(SA,SB,SC)\). It means that by applying a space nonzero voltage vector to an induction motor, the moving direction and amplitude of \(\psi_s\) will changed and that by applying a space zero voltage vector to the induction motor, the movement of \(\psi_s\) will be arrested. Therefore, by appropriately selecting the six space nonzero voltage vectors and their time widths, the trajectory of \(\psi_s\) can be made to follow a specified locus. Similarly, the insertion of one or more space zero voltage vectors between two space nonzero voltage vectors possible the regulation of the rotation velocity of \(\psi_s\).

3.6.2.1 Stator Flux Field Orientation Details

The stator flux field orientation details are explained in article 3.6.1.1. The same principle can be used in this scheme.

3.6.2.2 Direct Stator Flux Control

In order to achieve a high level of induction motor speed control, it is necessary to keep the stator flux as nearly constant as possible except for the field weakening operation.
Actually the inferiority of the induction motor in speed controllability does not lie in the induction motors themselves but in the ac power supply. In the case of sinusoidal ac power supplies, the trajectory of stator flux is a smooth circle. Unfortunately, most ac power produced by electronic inverters is not purely sinusoidal. It is therefore difficult to keep the stator flux constant in the regulated speed operation. The instantaneous space vector theory has made it possible to trace the stator flux vector and keep it almost constant over the entire speed range.

A smooth circle can be approximated by an equilateral polygon. In fact, as the number of sides of the polygon increases, the error can be made as small as desired. For a given smooth circle stator flux trajectory, there are many patterns by which the smooth circle can be approximated. It is, however, desirable that the transition loci have the properties of symmetry. An equilateral 12-sided polygon, where each side corresponds to a stator flux vector, is employed to approximate the smooth circle as shown in Fig.3.11.

It is obvious from Fig.3.11 that six out of the 12-space flux vectors have the same directions as those of the respective space nonzero voltage and are called real-flux vectors. On the other hand, another six out of 12-space flux vectors have different directions from the directions of space nonzero voltage vectors and are called pseudo-flux vectors. Consequently, the space nonzero voltage vectors can be switched to follow the real-flux vectors with particular time width. The time width can be calculated from Fig.3.12 and expressed by

$$T_s = \frac{\sqrt{6}\psi_m \tan 15^\circ}{V_d} \text{ ms}$$

(3.6.9)

where, $V_d$ is the dc link voltage of inverter in Volts,
$\psi_m$ is the rated amplitude of stator flux in milli-webers.

The pseudo-flux vectors must be synthesized by space nonzero voltage vectors by the following conditions:

- Two and only two space nonzero voltage vectors and one or more space zero voltage vectors are employed for each pseudo-flux vector.
Fig. 3.11 Primary flux vectors.

Fig. 3.12 Synthesis of the pseudo-flux vector.
• The locus synthesized by these two space nonzero voltage vectors has the same rotating direction as those of the real flux vectors.

Here, the main purpose is not to follow up the pseudo-flux vectors but the arc of the smooth circle. It can be seen from Fig.3.12 that if the time width is shorter, pulse amplitude of the stator flux will be smaller. On the other hand, the time width is longer, when pulse amplitude of the stator flux will be larger. As a compromise, four loci are selected for each pseudo-flux vector; these loci are symmetrical about the center of pseudo-flux vector. The time width $T_{ab}$ is equal to $\frac{T_{bc}}{2}$ as shown in Fig.3.12. The time widths of these flux vectors are calculated by

$$T_{ab} = \frac{\sqrt{2} \psi_m \tan 15'}{3V_{dc}} \text{ ms}$$  \hspace{1cm} (3.6.3)

$$T_{bc} = \frac{2\sqrt{2} \psi_m \tan 15'}{3V_{dc}} \text{ ms}$$ \hspace{1cm} (3.6.4)

where, $V_{dc}$, $\psi_m$ have the same meaning as in equation (3.6.9). It can be proved that the amplitude fluctuation of the stator flux is between 96 and 104% of the rated value [2]. For digital control technique, it is possible to construct two tables in which the space nonzero voltage vectors and their time widths are stored separately.

3.6.2.3 Effect of Space Voltage Vectors on Flux and Torque for Twelve Regions

The space nonzero voltage vectors ($V_1 \cdot V_6$) control the instantaneous stator current and stator flux to provide the fast dynamic response. The trajectory of stator flux moves in the direction of nonzero space voltage vector $V_i$ depending upon the switching sequence of $V_i$ to $V_6$. Therefore, applying a space nonzero voltage vector to an induction motor drive, the moving direction and amplitude of stator flux can be changed. So, by appropriately selecting the six space nonzero voltage vectors and their time widths, the trajectory of flux can be controlled to follow a specified locus. The instantaneous space vector theory is used to make stator flux sinusoidal within extended limit and the flux magnitude can be regulated. To make the locus of the stator flux vector a smooth circle, it can be approximated by an equilateral 12-sided polygon.
where each side corresponds to a stator flux vector. Six out of these 12-space flux vectors have the same directions as those of the respective space nonzero voltage vectors called as real-flux vectors. Remaining six out of the 12-space flux vectors are called pseudo-flux vector and each one of these is evolved as the combination of two nonzero voltage vectors and a zero voltage vector [2].

According to the instantaneous space vector theory, voltage, current and flux are represented by instantaneous space vectors. The vectors $V$, $i$, and $\psi$, are expressed as follows [2]:

\[
V(SA,SB,SC) = v_d + j v_q,
\]

\[
i = i_d + j i_q, \text{ and}
\]

\[
\psi = \psi_d + j \psi_q
\]

The change in stator flux vector from the present switching instant to the next switching instant can represented by (neglecting voltage drop across the stator resistance)

\[
\Delta \psi_s = (V_s - i_s R_s) \Delta T \equiv V_s \Delta T
\]

The change in stator current vector is

\[
\Delta i_s = \Delta i_d + j \Delta i_q
\]

$\Delta i_d$, $\Delta i_q$ affect the change in flux and the change in torque respectively. The plane of the instantaneous space voltage vectors is divided into 12 regions as shown in Fig.3.13 [12]. The effect of nonzero voltage vectors ($V_i - V_j$) on flux and torque for the region ‘1’ (-15° to +15°) as shown in Fig.3.14. The Table 3.6 shows the effect of space voltage vectors ($V_i - V_j$) on flux and torque for 12 regions. This knowledge base has to be used for selection of nonzero voltage vectors and zero voltage vectors at any particular instant.

**CONCLUSIONS**

In this chapter the space vector modulated voltage source inverter for direct self control of induction motor has been presented. The different direct torque and flux control strategies for direct self control are discussed. The schematic for generating six sided space
Fig. 3.13 The plane of instantaneous space voltage vectors is divided into 12-regions

Table 3.6. The effect of space voltage vectors on flux and torque for 12-regions

<table>
<thead>
<tr>
<th>Angle</th>
<th>$V_0$</th>
<th>$V_1$</th>
<th>$V_2$</th>
<th>$V_3$</th>
<th>$V_4$</th>
<th>$V_5$</th>
<th>$V_6$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\Delta \psi$ change in flux and $\Delta T_e$ change in electromagnetic torque</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$-15^\circ &lt; 0 &lt; 15^\circ$</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$\Delta \psi$ zero</td>
<td>$\Delta \psi$ zero</td>
<td>$\Delta \psi$ zero</td>
<td>$\Delta \psi$ zero</td>
<td>$\Delta \psi$ zero</td>
<td>$\Delta \psi$ zero</td>
<td>$\Delta \psi$ zero</td>
<td></td>
</tr>
<tr>
<td>$\Delta T_e$ zero</td>
<td>$\Delta T_e$ zero</td>
<td>$\Delta T_e$ zero</td>
<td>$\Delta T_e$ zero</td>
<td>$\Delta T_e$ zero</td>
<td>$\Delta T_e$ zero</td>
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</tr>
<tr>
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<td></td>
<td></td>
<td></td>
<td></td>
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<td></td>
</tr>
<tr>
<td>$\Delta \psi$ zero</td>
<td>$\Delta \psi$ zero</td>
<td>$\Delta \psi$ zero</td>
<td>$\Delta \psi$ zero</td>
<td>$\Delta \psi$ zero</td>
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<td>$75^\circ &lt; 0 &lt; 105^\circ$</td>
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<td>$\Delta \psi$ zero</td>
<td>$\Delta \psi$ zero</td>
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<td>$135^\circ &lt; 0 &lt; 165^\circ$</td>
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<td>$\Delta \psi$ zero</td>
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\( \Delta \psi \), change in stator current vector, \( \Delta \phi \) change in stator flux vector, and \( \Delta T_r \) change in torque.

Fig. 3.14 Effect of nonzero voltage vectors on flux and torque in region “1” (-15° to +15°).
vector polygon has been presented. By this scheme the performance of induction motor drive is sluggish. To improve drive performance, twelve sided space vector polygon has been brought out. This twelve sided polygon scheme has been used by the author in investigation of conventional direct self control induction motor drive. To study fuzzy direct self control induction motor drive the space vector plane has been divided into twelve regions. The effect of space voltage vectors \(V_0-V_6\) on flux and torque in these 12 regions of instantaneous space vector plane has been presented by the author in tabular manner. The author has used this knowledge base for the proposed fuzzy logic controller based direct self control of induction motor drive.

REFERENCES


