Generalized Scalar PWM Algorithms for VSI Fed Direct Torque Controlled Induction Motor Drives

B Ravi Chandra Rao #1, Dr. G.V. Marutheswar*2

1Assistant Professor, Electrical and Electronics Engineering Department, GNITS, Hyderabad, T.S. India
2Professor, Electrical Engineering Department, SV University, Tirupathi, A.P. India
1rao.ravichandra@gmail.com

Abstract—Induction motor drive based on direct torque control (DTC) allows high dynamic performance to be obtained with very simple hysteresis control scheme. Direct control of the torque and flux is achieved by proper selection of inverter voltage space vector through a lookup table. Direct torque control (DTC) is quite different from vector control and has several advantages over vector control. This paper presents intelligent control scheme together with conventional control scheme. This paper presents generalized scalar PWM algorithms for VSI fed direct torque control (DTC) induction motor drives for the reduction of harmonics and switching losses. In this proposed algorithm the lookup tables are not employed for the generation of the space vector locations and hence sector identification and angle calculation is not required. This algorithm employs only the instantaneous reference phase voltages for the implementation of the space vector pulse width modulation without estimation of angle of reference vector, and the proposed algorithm calculates the switching time by using the concept of the imaginary switching times. To validate the proposed algorithm, simulation studies have been carried out on VSI fed direct torque control (DTC) induction motor drives and results are presented and compared.

Keywords—SVPWM, Scalar approach, DPWM, Harmonic Distortion, DTC

I. INTRODUCTION

In present days, the research has been focused to find out different solutions for the induction motor control having the features of precise and quick torque response and reduction of the complexity of field oriented control[1-5]. The direct torque control (DTC) technique has been recognized as the viable solution to achieve these requirements. The DTC based induction motor drives were developed and presented more than two decades ago by I.Takahashi. This technique is based on the space vector approach, where the torque and flux of an induction motor can be directly and independently controlled without any coordination transformation[6-9]. Though the DTC gives fast transient response, it gives large steady state ripples and variable switching frequency of the inverter.

To reduce the steady state ripples and to get constant switching frequency of the inverter, space vector PWM (SVPWM) algorithm has been used for DTC. However, the classical SVPWM algorithm also gives more harmonic distortion at higher modulation indices and switching losses. Moreover, the complexity involved is more due to the sector and reference voltage vector calculations. To reduce the switching losses various discontinuous PWM (DPWM) techniques were proposed in literature[10-15].

Hence, to reduce the complexity, in this paper a generalized scalar PWM (GSPWM) has been proposed. In the proposed approach, by adding a constant between 0 and 1, SVPWM and DPWM algorithms can be generated at all modulation indices. Moreover, in this work, a simplified approach for the over modulation region, in which a unique zero sequence signal is added to the sinusoidal signals to generate the all possible PWM algorithms. Thus, the proposed approach will bring all modulators under a common roof with less complexity.
II. DIRECT TORQUE CONTROL OF INDUCTION MOTOR DRIVE

The electromagnetic torque of a three-phase induction motor can be written as

\[ T_e = \frac{3}{2} \frac{P}{2} \frac{L_m}{\sigma L_s L_r} |\psi_s| |\psi_r| \sin \eta \]  

(1)

where \( \eta \) is the angle between the stator flux linkage space vector \( \psi_s \) and rotor flux linkage space vector \( \psi_r \), as shown in Fig. 1 and \( \sigma \) is the leakage coefficient given by \( \sigma = 1 - \left( \frac{L_m^2}{L_s L_r} \right) \). The expression given in (1) is valid for both the steady state and transient state conditions. In steady state both the stator flux and rotor flux moves with the same angular velocity. The rotor flux lags the stator flux by torque angle. This is the essence of “Direct Torque Control”. During a short transient, the rotor flux is almost unchanged, thus rapid changes of electromagnetic torque can be produced by rotating the stator flux in the required direction according to the demanded torque. However, the stator flux linkage space vector can be changed by the stator voltages[15-17].

![Fig. 1 \( \psi_r \) movement relative to \( \psi_s \) under influence of voltage vectors](image)

If for simplicity it is assumed that the stator ohmic drops can be neglected, then \( \bar{V}_s = \frac{d\bar{\psi}_s}{dt} \), so the inverter voltage directly impresses the stator flux. For a short time \( \Delta t \), when the voltage vector is applied, \( \Delta \bar{\psi}_s = \bar{V}_s \Delta t \). Thus, the stator flux linkage space vector moves by \( \Delta \bar{\psi}_s \) in the direction of the stator voltage space vector at a speed proportional to magnitude of voltage space vector. By selecting step-by-step the appropriate stator voltage vector, it is then possible to change the stator flux in the required direction. Decoupled control of the torque and stator flux is achieved by acting on the radial and tangential components of the stator flux linkage vector in the locus. These two components are directly proportional to the components of the stator voltage vector in the same directions.
Thus, for the torque production, the angle $\eta$ plays a very important role. By assuming a slow motion of the rotor flux linkage space vector, if a forward active voltage vector is applied then it causes rapid movement of $\psi_s$ and torque increases with $\eta$. On the other hand, when a zero voltage vector is used, the stator flux vector $\psi_s$ becomes stationary and the electromagnetic torque will decrease, since $\psi_s$ continues to move forward and the angle $\eta$ decreases. If the duration of zero voltage space vector is sufficiently long, then the rotor flux linkage space vector will overtake the stator flux linkage space vector, the angle $\eta$ will change its sign and the torque will also change its direction. Thus, it is possible to change the speed of stator flux linkage space vector by changing the ratio between the zero and non-zero voltage vectors. It is important to note that the duration of zero voltage vectors has a direct effect on torque oscillations.

By considering the three-phase, two-level, voltage source inverter (VSI), there are six non-zero active voltage space vectors and two zero voltage space vectors as shown in Fig. 2. The six active voltage space vectors can be represented as

$$V_k = \frac{2}{3} V_{dc} \exp\left[j(k-1)\pi/3\right] \quad k = 1,2,\ldots,6$$

Depending on the position of stator flux linkage space vector, it is possible to switch the appropriate voltage vectors to control both stator flux and torque. As an example if stator flux linkage space vector is in sector I as shown in Fig. 3, then voltage vectors $V_2$ and $V_6$ can increase the stator flux and $V_3$ and $V_5$ can decrease the stator flux. Similarly $V_2$ and $V_3$ can increase the torque and $V_5$ and $V_6$ can decrease the torque. Similarly the suitable voltage vectors can be selected for other sectors. Thus, as depicted in the Fig. 3, when the stator flux amplitude has to be increased, a voltage vector, phase shifted by an angle larger than 90° with respect to existing stator flux linkage space vector ($\psi_s$) is applied. In contrast, if stator flux amplitude has to be reduced, a voltage vector, phase shifted by an angle less than 90° will be applied. Similarly, by applying suitable voltage space vector in the direction of rotation of stator flux, torque ($T_e$) can be increased and by applying the voltage space vector, which is opposite to the direction of rotation of stator flux, $T_e$ can be reduced.
The Fig. 4 shows the block diagram of conventional direct torque controlled induction motor drive.

There are two hysteresis control loops, one for the control of torque and other for the control of stator flux. The flux controller controls the machine operating flux to maintain the magnitude of the operating flux at the rated value till the rated speed. Torque control loop maintains the torque close to the torque demand. Based on the outputs of these controllers and the instantaneous position of stator flux vector, a proper voltage space vector is selected.

### III. PROPOSED GENERALIZED SCALAR PWM ALGORITHM

The SVPWM algorithm divides the zero state time equally among the two possible zero states in each sampling time interval. Whereas the proposed generalized scalar approach utilizes the freedom of zero state time division in order to derive various PWM algorithms. The proposed approach distributes the zero state time as \((1-k_0)T_z\) for \(V_q(000)\) and \(k_0T_z\) for \(V_q(111)\), where \(k_0\) varies between 0 and 1. Fig. 5 correlate the SV and TC approaches based PWM algorithms, which shows the timing of gating pulses of TC method and the timing of voltage space vectors in the first sector with SV approach.
From Fig. 5, the time durations of gating pulses in the first sector for the top switches of a three-phase, two-level VSI are given by

\[ T_{ga} = T_1 + T_2 + T_7 = T_1 + T_2 + k_o T_z \]  
(3)

\[ T_{gb} = T_2 + T_7 = T_2 + k_o T_z \]  
(4)

\[ T_{gc} = T_7 = k_o T_z \]  
(5)

The active and zero state time duration can be found by using the instantaneous phase voltages as given in the following equations:

\[ T_1 = \frac{T_s}{V_{dc}} (V_{\text{max}} - V_{\text{mid}}) \]  
(6)

\[ T_2 = \frac{T_s}{V_{dc}} (V_{\text{mid}} - V_{\text{min}}) \]  
(7)

\[ T_z = T_s - T_1 - T_2 \]  
(8)

where \( V_{\text{max}}, V_{\text{mid}}, \) and \( V_{\text{min}} \) are maximum, middle and minimum values of \( V_{\text{in}} \), which is given by

\[ V_{\text{in}} = V_{\text{ref}} \cos(0 - 2(r - 1)\pi/3) \]  
(9)

for \( i = a, b, c \) and \( r = 1, 2, 3 \)

\[ v_l = \left( \frac{2t}{T_s} - 1 \right) \frac{V_{dc}}{2} \]  
(10)

Fig. 5 Correlation between TC and SC approaches (a) timing of gating pulses of TC approach (b) timing of voltage space vectors in SV approach

To correlate the TC and SV approaches, one has to find a set of modulating waveforms that, on comparison with a common triangular carrier wave, would produce the same gating pulses as would the SV approach. The equation describing the triangle wave form, which is shown in Fig. 5 is given by
where \( v_t \) is the instantaneous value of triangular carrier waveform. By adding the zero sequence signal to the old set of reference voltages \( V_{in} \), a new set of reference phase voltages, which are also known as modulating waveforms \( V_{in}^* \) can be generated as follows

\[ V_{in}^* = V_{in} + V_{zs} \quad (11) \]

where \( V_{zs} \) is the zero sequence voltage and can be calculated as below:

\[ V_{zs} = \frac{V_a + V_b + V_c}{3} = \frac{2(T_a + T_b + T_c)}{T_s} - 1 \]

\[ = \frac{V_{dc}}{2} (2k_o - 1) - k_o V_{max} + (k_o - 1)V_{min} \quad (12) \]

In order to generate the various PWM algorithms, assume a set of three-phase voltages as given in (13).

\[ V_{ir} = V_{ref} \cos(\theta - 2(r - 1)\pi/3 - \pi/6) \]

for \( i = a, b, c \) and \( r = 1, 2, 3 \)

Then, the maximum and minimum values of \( V_{ir} \) can be calculated as \( V_{max,x} = \max(V_{ix}) \), \( V_{min,x} = \min(V_{ix}) \). In the proposed Generalized Scalar PWM (GSPWM) algorithm, by varying the zero voltage vector time partition parameter \( k_o \) between 0 and 1, various PWM algorithms can be generated. Table 1 shows the various PWM algorithms with the variation of \( k_o \) value.

<table>
<thead>
<tr>
<th>PWM algorithm</th>
<th>value of ( k_o )</th>
</tr>
</thead>
<tbody>
<tr>
<td>SVPWM</td>
<td>0.5</td>
</tr>
<tr>
<td>DPWMMIN</td>
<td>0</td>
</tr>
<tr>
<td>DPWMMAX</td>
<td>1</td>
</tr>
</tbody>
</table>
| DPWM0        | if \( V_{max,x} + V_{min,x} < 0 \) \( k_o = 1 \)
|              | if \( V_{max,x} + V_{min,x} \geq 0 \) \( k_o = 0 \) |
| DPWM1        | if \( V_{max} + V_{min} \) \( < 0 \) \( k_o = 0 \)
|              | if \( V_{max} + V_{min} \geq 0 \) \( k_o = 1 \) |
| DPWM2        | if \( V_{max,x} + V_{min,x} < 0 \) \( k_o = 0 \)
|              | if \( V_{max,x} + V_{min,x} \geq 0 \) \( k_o = 1 \) |
| DPWM3        | if \( V_{max} + V_{min} \) \( < 0 \) \( k_o = 1 \)
|              | if \( V_{max} + V_{min} \geq 0 \) \( k_o = 0 \) |

The old reference phase voltage, zero sequence voltage and new reference phase voltage (modulating waveform) of various PWM algorithms are shown in Fig. 6 at a modulation index of 0.7.
IV. RESULTS AND DISCUSSION

To validate the conventional direct torque controlled induction motor drive several numerical simulations have been carried out by using Matlab/Simulink. For the simulation studies, the reference flux is taken as 1wb and starting torque is limited to 40 N·m. The steady state results for conventional direct torque controlled induction motor drive are shown in Fig 7 to Fig 30.

Fig. 7 shows the steady state plots of speed, torque, stator currents and stator flux of CDTC based IM drive. Fig. 8 gives the phase and line voltages of CDTC based IM drive during the steady state. The total harmonic distortion (THD) of the line current is shown in Fig. 9. From Fig. 7 to Fig. 9, it can be observed that the CDTC algorithm gives large steady state ripples in torque, current and stator flux. Moreover, it can be observed that the CDTC gives variable switching frequency operation of the inverter. The THD of line current is also quite high for CDTC based IM drive.

Matlab-Simulink based simulation studies have been carried out to validate the proposed GSPWM algorithm based direct torque controlled induction motor drive. Various conditions such as starting, steady state, step change in load and speed reversal are simulated. The average switching frequency of the inverter is taken as 3 kHz. For the simulation, the reference flux is taken as 1wb and starting torque is limited to 40 N·m. The simulated results at 1300 rpm (corresponding to the higher modulation region) are shown in Fig 3.52 to Fig 3.135. From the simulation results shown in Fig. 10-Fig. 30, it can be observed that the proposed GPWM algorithm generates all possible PWM algorithms and also gives good performance at all operating conditions such as starting, steady state, load change and speed reversal conditions.

From the steady state results of various PWM algorithms based DTC, it can be concluded that the SVPWM and DPWM algorithms give less steady state ripples when compared with the conventional DTC. Moreover, it can be observed that the SVPWM and DPWM algorithms give constant switching frequency operation of the inverter. Also, from the simulation results it can be observed that at higher modulation indices (motor is running at 1300 rpm, which pertains to the higher modulation region), the DPWM algorithms give reduced harmonic distortion when compared with the SVPWM algorithm. Moreover, it can be concluded that among the various DPWM algorithms, DPWM3 gives reduced harmonic distortion.
From the simulation results, it can be observed that as the SVPWM algorithm is a continuous PWM algorithm, it gives continuous pulse pattern and more switching losses of the inverter. Whereas, the DPWM algorithms clamp each phase to either positive or negative DC bus for 120 degrees over a fundamental cycle, these reduce the switching frequency and switching losses by 33.33% when compared with the SVPWM algorithm. Thus, the proposed GPWM algorithm generates a wide range of PWM algorithms at all modulation indices with reduced complexity by varying a parameter $k_o$ from 0 to 1.

Fig. 7 steady state plots of speed, torque, stator currents and stator flux for CDTC based IM drive at 1200 rpm

Fig. 8 the phase and line voltages for CDTC based IM drive during the steady state
Fig. 9 Harmonic Spectrum of stator current along with THD.

Fig. 10 Steady state plots of speed, currents, torque and flux for SVPWM based DTC

Fig. 11 modulating wave, phase and line voltages during the steady state for SVPWM based DTC
Fig. 12 harmonic spectra of line current for SVPWM based DTC drive along with the THD value

Fig. 13 Steady state plots of speed, currents, torque and flux for DPWMMIN based DTC

Fig. 14 modulating wave, phase and line voltages during the steady state for DPWMMIN based DTC
Fig. 15 harmonic spectra of line current for DPWMMIN based DTC drive along with the THD value

Fig. 16 Steady state plots of speed, currents, torque and flux for DPWMMAX based DTC

Fig. 17 modulating wave, phase and line voltages during the steady state for DPWMMAX based DTC
Fig. 18 harmonic spectra of line current for DPWMMAX based DTC drive along with the THD value

Fig. 19 Steady state plots of speed, currents, torque and flux for DPWM0 based DTC

Fig. 20 modulating wave, phase and line voltages during the steady state for DPWM0 based DTC
Fig. 21 harmonic spectra of line current for DPWM0 based DTC drive along with the THD value

Fig. 22 Steady state plots of speed, currents, torque and flux for DPWM1 based DTC

Fig. 23 modulating wave, phase and line voltages during the steady state for DPWM1 based DTC
Fig. 24 harmonic spectra of line current for DPWM1 based DTC drive along with the THD value.

Fig. 25 Steady state plots of speed, currents, torque and flux for DPWM2 based DTC.

Fig. 26 modulating wave, phase and line voltages during the steady state for DPWM2 based DTC.
Fig. 27 harmonic spectra of line current for DPWM2 based DTC drive along with the THD value

Fig. 28 Steady state plots of speed, currents, torque and flux for DPWM3 based DTC

Fig. 29 modulating wave, phase and line voltages during the steady state for DPWM3 based DTC
The DTC algorithm is becoming popular in many industrial applications due to its numerous advantages. Though the DTC gives fast transient response, it gives large steady state ripples in the torque, flux and currents. To reduce the steady state ripples, the SVPWM algorithm based DTC is used. However, the complexity involved in the classical SVPWM algorithm is more due to the angle and sector calculations. Also, as the SVPWM algorithm is a continuous PWM algorithm, it gives more switching losses of the inverter. Hence, to reduce the complexity involved and in order to generalize all the possible PWM algorithms, a generalized PWM algorithm is presented in this paper. From the simulation results it can be observed that the proposed GPWM algorithm gives all possible PWM modulators with reduced complexity. Moreover, it can be concluded that among the various DPWM algorithms, DPWM3 gives reduced harmonic distortion. Also, it can be observed that the DPWM algorithms will reduce the switching frequency and switching losses by 33.33% when compared with the SVPWM algorithm. Thus, the proposed GSPWM algorithm generates a wide range of PWM algorithms at all modulation indices with reduced complexity by varying a parameter $k_o$ from 0 to 1.

**REFERENCES**


[13] ABB Technical guide, “Direct Torque Control - the world’s most advanced AC drive technology”


