Chapter 3
Five-Phase Voltage Source Inverter-Modelling and Control for single motor Drive

3.1 INTRODUCTION:

Control of three-phase VSIs is nowadays, except in the highest power range, always based on pulse width modulation (PWM) schemes. A number of PWM techniques are available to control a two-level three-phase VSI. The most widely used pulse width modulation technique for three-phase inverters are the carrier-based sine-triangle PWM (SPWM) and the space vector PWM (SVPWM). These techniques have been extensively discussed in the literature. The same does not apply to multi-phase VSIs, since there are few application specific PWM techniques available. Conventional Space vector PWM for a five-phase VSI is available in the literature Levi et al (2007) and Levi (2008).

The SPWM schemes are more flexible and simpler to implement, but the maximum peak of the fundamental component in the output voltage is limited to 50% of the DC link voltage Levi (2008) and the extension of the SPWM scheme in the over modulation is difficult. For two level inverter a SVPWM like performance can be obtained with a SPWM scheme by adding a common mode voltage of suitable magnitude, to the sinusoidal reference phase voltage. A modulation scheme is presented in Zhou and Wang (2002), where a fixed common mode voltage is added to the reference phasor throughout the modulation range. It has been shown in Gopakumar et al (2003) that this common mode addition will not result in a SVPWM like performance, as it will not centre the middle inverter vectors in a sampling interval. The common mode voltage to be added in the reference phase voltages is a function of modulation index. The similar scheme for a five-phase VSI is presented in Chin and Yang (2004) where the offset addition resulted in an increase in the output voltage to 5.15% compared to the carrier-based PWM. Thus the output obtainable with offset addition is same as that of SVPWM.

The proposed SVPWM scheme does not involve checks for the sector identification and lookup tables, for switching vector determination. Thus the scheme is computationally efficient when compared to the conventional SVPWM scheme and making it simpler for real time implementation.

This chapter is devoted to the modelling and control of a five-phase voltage source inverter for single-motor drive system. The inverter is assumed to feed a five-phase ac...
machine. The model is developed on the space vector approach. The space vector model obtained is decomposed into the two orthogonal sub-spaces namely $d$-$q$ and $x$-$y$. This is owing to the fact that a five-phase system is dealt herewith and thus there exist two orthogonal planes and one plane corresponds to the zero sequence component. The zero sequence components are further neglected as isolated neutral star-connected load is assumed. The space vector model suggest 32 space vectors out of which 30 are active and two zero. These vectors are further used to develop different modulation strategies. At first simple carrier-based PWM scheme is discussed which is based on the comparison of high frequency triangular carrier-wave and the sinusoidal modulation signals. The concept of harmonic injection is employed to enhance the output voltage from the inverter. This is followed by the review of the existing space vector PWM technique employed for single motor drive system. A new proposal is given for the voltage modulation of five-phase voltage source inverter. The proposed modulation is the extension of the concept employed in three-phase voltage source inverters. The proposed modulation is extremely simple in real time implementation when compared to the existing space vector PWM. This is followed by the development of space vector PWM using artificial neural network. The complete simulation model is also provided using Matlab/Simulink. The chapter analyses both the simulation and experimental results.

3.2 MODELING OF FIVE-PHASE VOLTAGE SOURCE INVERTER

Power circuit topology of a five-phase VSI is shown in Fig. 3.1. The inverter input DC voltage is regarded as being constant. The load is taken as star-connected and the inverter output phase voltages are denoted in Fig. 1 with lower case symbol $(a,b,c,d,e)$, while the leg voltages have symbols in capital letters $(A,B,C,D,E)$. The model of five-phase VSI is developed in space vector form in Duran et al (2008), assuming an ideal commutation and zero forward voltage drop. A brief review is presented here.

There are ten switching devices and only five of them are independent, as the operation of two power switches of the same leg is complimentary. The combination of these five switching states gives out thirty two (32) space voltage vectors. Out of thirty two space vectors thirty are active vectors and two zero vectors. At any instant of time, the inverter can produce only one space vector.
The relationship between the machine’s phase-to-neutral voltages and inverter leg voltages are given with

\[
\begin{align*}
\psi_a &= \frac{4}{5}v_a - \left(\frac{1}{5}v_b + v_c + v_d + v_e\right) \\
\psi_b &= \frac{4}{5}v_b - \left(\frac{1}{5}v_a + v_c + v_d + v_e\right) \\
\psi_c &= \frac{4}{5}v_c - \left(\frac{1}{5}v_a + v_b + v_d + v_e\right) \\
\psi_d &= \frac{4}{5}v_d - \left(\frac{1}{5}v_a + v_b + v_c + v_e\right) \\
\psi_e &= \frac{4}{5}v_e - \left(\frac{1}{5}v_a + v_b + v_c + v_d\right)
\end{align*}
\]  

(3.1)

where the inverter leg voltages take the value of ± 0.5 \(V_{dc}\). As noted, lower case letters in indices define phase-to-neutral voltages. The magnitude and phase angle of vectors in \(d-q\) and \(x-y\) are listed in Table 1.

Space vector of phase voltages defined, using power variant transformation, as given in Arahal and Duran (2009)

\[
\frac{2}{5}(v_a + a^2v_b + a^4v_c + a^6v_d + a^8v_e)
\]  

(3.2)

where \(a = \exp(j2\pi/5)\), \(a^2 = \exp(j4\pi/5)\), \(a^* = \exp(-j2\pi/5)\), \(a^{*2} = \exp(-j4\pi/5)\) and * stands for a complex conjugate. The phase voltage space vectors thus obtained in \(d-q\) plane are shown in Fig. 3.2. Since it is a five-phase system, transformation is further done to obtain space vectors in \(x-y\) plane using (3.3) and the resulting space vectors are shown in Fig. 3.3.

\[
\frac{2}{5}(v_a + a^2v_b + a^4v_c + a^6v_d + a^8v_e)
\]  

(3.3)

It can be seen from Fig. 3.2 that the outer decagon space vectors of the \(d-q\) plane map into the inner decagon of the \(x-y\) plane (Fig. 3.3), the innermost decagon of \(d-q\) plane forms the outer decagon of the \(x-y\) plane, while the middle decagon space vectors map into the same region. Further, it is observed from the above mapping that the phase sequence \(a,b,c,d,e\) of the \(d-q\) plane corresponds to \(a,c,e,b,d\) sequence of the \(x-y\) plane.
Table 1. Magnitude and angle of the vectors in $d$-$q$ and $x$-$y$ axis

<table>
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<th>Decimal values</th>
<th>Equivalent binary no.</th>
<th>$d$-$q$ axis</th>
<th>$x$-$y$ axis</th>
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3.3 PULSE WIDTH MODULATION TECHNIQUES OF A FIVE-PHASE VSI

Pulse Width modulation technique is the most basic method of energy processing in power electronic converters. The purpose here is to control the inverter to generate the variable voltage and variable frequency voltages/currents. This section describes the PWM techniques employed for controlling a five-phase voltage source inverter. The basic underlying principle of PWM techniques remains the same as that of three-phase voltage source inverter. Purpose here is to generate the five-phase sinusoidal output voltage. For this purpose different existing pulse width modulation techniques and newly proposed techniques are;

*Existing PWM Techniques*
1. Carrier based sinusoidal PWM
2. Fifth harmonic injection based pulse width modulation scheme
3. Offset addition based pulse width modulation scheme
4. Space vector pulse width modulation scheme

*Proposed PWM Techniques*
5. Time equivalent space vector pulse width modulation (TESVPWM) scheme
6. Artificial neural network based space vector pulse width modulation.

3.3.1 CARRIER BASED SINUSOIDAL PULSE WIDTH MODULATION SCHEME

Carrier-based sinusoidal PWM is the most popular and widely used PWM technique because of their simple implementation in both analogue and digital realisation Iqbal et al
The principle of carrier-based PWM true for a three-phase VSI is also applicable to a multi-phase VSI. The PWM signal is generated by comparing a sinusoidal modulating signal with a triangular (double edge) or a saw-tooth (single edge) carrier signal. The frequency of the carrier is normally kept much higher compared to the modulating signal. Principle of operation of a carrier-based PWM modulator is shown in Fig. 3.4 and generation of PWM waveform is illustrated in Fig. 3.5. Modulation signals are obtained using five fundamental sinusoidal signals (displaced in time by $\alpha = 2\pi / 5$), which are summed with an appropriate zero-sequence signal. These modulation signals are compared with high frequency carrier signal (saw-tooth or triangular shape) and all five switching functions for inverter legs are obtained directly. In general, modulation signal can be expressed as:

$$v_i(t) = v'_i(t) + v_{\text{inv}}(t)$$  \hspace{1cm} (3.4)

where $i = a, b, c, d, e$ and $v_{\text{inv}}$ represents zero-sequence signal and $v'_i$ are fundamental sinusoidal signals. Zero-sequence signal represents a degree of freedom that exits in the structure of a carrier-based modulator and it is used to modify modulation signal waveforms and thus to obtain different modulation schemes. Continuous PWM schemes, analyzed in this paper, are characterized by presence of switching activity in each of the inverter legs over the carrier signal period, as long as peak value of the modulation signal does not exceed the carrier magnitude.

The following relationships hold true in Fig. 3.5:

$$t^+_n - t^-_n = v_n t_n$$ \hspace{1cm} (3.5a)

where

$$t^+_n = \left(\frac{1}{2} + v_n\right) t_n$$ \hspace{1cm} (3.5b)

$$t^-_n = \left(\frac{1}{2} - v_n\right) t_n$$ \hspace{1cm} (3.5c)

where $t^+_n$ and $t^-_n$ are the positive and negative pulse widths in the $n^{th}$ sampling interval, respectively and $v_n$ is the normalised amplitude of modulation signal. The normalization is done with respect to $V_{dc}$. Equation (3.5) is referred as the Equal volt-second principle as applied to a three-phase inverter Ward and Härer (1969). The normalised peak value of the triangular carrier wave is $\pm 0.5$ in linear region of operation. Modulator gain has the unity value while operating in the linear region and peak value of inverter output fundamental voltage is equal to the peak value of the fundamental sinusoidal signal. Thus the maximum
output phase voltages from a five-phase VSI are limited to 0.5 p.u. This is also evident in Iqbal et al (2006). Thus the output phase voltage from a three-phase and a five-phase VSI are same when utilising carrier-based PWM, Blasko (1997), Hava et al (1998).

Matlab/Simulink model is developed and the simulation results are illustrated in Fig. 3.6. The dc link voltage is kept unity so that the results are in per unit. The switching frequency is kept 5 kHz. The fundamental frequency is kept 50 Hz. The output voltage is 0.5 p.u. (typically it the same value as that of three-phase VSI).
3.3.2 HARMONIC INJECTION BASED PULSE WIDTH MODULATION SCHEME

The effect of addition of harmonic with reverse polarity in any signal is to reduce the peak of the reference signal. Aim here is to bring the amplitude of the reference as low as possible, so that the reference can then be pushed to make it equal to the carrier, resulting in the higher output voltage and better dc bus utilisation. Using this principle, third harmonic injection PWM scheme is used in a three-phase VSI which results in increase in the fundamental output voltage to 0.575\(V_{dc}\) Iqbal et al (2006). Third harmonic voltages do not appear in the output phase voltages and are restricted to the leg voltages only. Following the same principle, fifth harmonic injection PWM scheme is used to increase the modulation index of a five-phase VSI Duran et al (2008), Arahal and Duran (2009).

The reference leg voltages are given as

\[
\begin{align*}
V_{o0}^* &= 0.5M_1V_{dc} \cos(\alpha t) + 0.5M_5V_{dc} \cos(5\alpha t) \\
V_{b0}^* &= 0.5M_1V_{dc} \cos(\alpha t - 2\pi / 5) + 0.5M_5V_{dc} \cos(5\alpha t) \\
V_{c0}^* &= 0.5M_1V_{dc} \cos(\alpha t - 4\pi / 5) + 0.5M_5V_{dc} \cos(5\alpha t) \\
V_{do}^* &= 0.5M_1V_{dc} \cos(\alpha t + 4\pi / 5) + 0.5M_5V_{dc} \cos(5\alpha t) \\
V_{eo}^* &= 0.5M_1V_{dc} \cos(\alpha t + 2\pi / 5) + 0.5M_5V_{dc} \cos(5\alpha t)
\end{align*}
\tag{3.6}
\]

It is to be noted that fifth-harmonic has no effect on the value of the reference waveform when \(\alpha t = (2k + 1)\pi / 10\), since \(\cos(s(2k + 1)\pi / 10) = 0\) for all odd \(k\). Thus \(M_5\) is chosen to make the peak magnitude of the reference of (3.6) occur where the fifth-harmonic is zero. This ensures the maximum possible value of the fundamental component. The reference voltage reaches a maximum when
\[
\frac{dV^*_{ao}}{dt} = -0.5M_iV_{dc} \sin \omega t - 0.5 \cdot 5M_iV_{dc} \sin 5\omega t = 0
\]  
This yields
\[
M_s = -M_i \frac{\sin(\pi/10)}{5}; \text{ for } \omega t = \pi/10
\]  
Thus the maximum modulation index can be determined from
\[
|V^*_{ao}| = 0.5M_iV_{dc} \cos(\omega t) - 0.5 \frac{\sin(\pi/10)}{5} M_iV_{dc} \cos(3\omega t) = 0.5V_{dc}
\]  
The above equation gives
\[
M_i = \frac{1}{\cos(\pi/10)}; \text{ for } \omega t = \pi/10
\]  
Thus the output fundamental voltage is increased by 5.15% higher than the value obtainable using simple carrier-based PWM by injecting 6.18% fifth-harmonic in fundamental. The fifth-harmonic is in opposite phase to that of the fundamental. The simulation conditions are kept same as section 3.3.1 and the resulting waveforms are as shown in Fig. 3.7.

3.3.3 OFFSET ADDITION BASED PULSE WIDTH MODULATION SCHEME

Another way of increasing the modulation index is to add an offset voltage to the references. The offset voltage addition is effectively adding \(3n\) harmonic. This will effectively do the same function as above. The offset voltage is given as
\[
V_{offset} = -\frac{V_{max} + V_{min}}{2}
\]  
where \(V_{max} = \max(v_a, v_b, v_c, v_d, v_e)\) and \(V_{min} = \min(v_a, v_b, v_c, v_d, v_e)\). Note that this is the same as for a three-phase inverter.
In case of three-phase VSI the offset voltage is simply third harmonic triangular wave of 25\% magnitude of fundamental. This has been shown in literature (25\% value explicitly appears in Kazmierkowski (2002). The peak of the fundamental is 0.575 p.u., (0.406 p.u. rms), the peak of the resultant modulating signal is 0.5 p.u. (0.353 p.u. rms) and the peak of the offset is 0.147 p.u. (0.104 p.u. rms). Hence offset peak is 25\% of the fundamental peak.

In a five-phase VSI the offset is found as the fifth harmonic triangular wave of 9.55\% of the fundamental input reference. This value has been established by simulations. Offset addition requires only addition operation and hence is suitable for practical implementation.

Ojo and Dong (2005) has given a generalised formula of offset voltage, which is to be injected along with the fundamental in case of five-phase VSI. The expression is

\[ V_{no} = -0.5528(V_{max} - V_{min}) + 3/5(1-2\mu)V_{dc}/2 - 3/5(1-2\mu)(V_{max} - V_{min}) \]

where \( V_{max} \) is the maximum of the five-phase references, \( V_{min} \) is the minimum of the five-phase references and \( \mu \) is the factor which decides the placement of the two zero vector states. If it is 0.5 then the two zero states are placed equally and this corresponds to symmetrical zero vector placement.

It is important to note that not only the 5\textsuperscript{th} harmonic, but all the additional 5\( k (k = 1,3,5,\ldots) \) harmonics as well, are included in the modulation signal in this technique. Maximum modulation index has the same value as in the previous case, \( M_{IM} = 1.0515 \). The simulation results are as shown in Fig. 3.8.

(a) Filtered output voltage

(b) Harmonic spectrum phase 'a' voltage

Fig. 3.8 Simulation results of the offset addition based PWM scheme
Space vector pulse width modulation has become one of the most popular PWM techniques because of its easier digital implementation and better dc bus utilisation, when compared to the ramp-comparison sinusoidal PWM method. The principle of SVPWM lies in the switching of inverter in a special way so apply a set of space vector for specific time. SVPWM for three-phase voltage source inverter has been extensively discussed in the literature and the technology has matured (Iqbal and Levi 2005), (Homes and Lipo 2003). However, for multi-phase VSIs, there are only application specific SVPWM techniques available in the literature and more research work is needed in this area. There is a lot of flexibility available in choosing the proper space vector combination for an effective control of multi-phase VSIs because of large numbers of space vectors. With reference to five-phase VSI, there are a few examples found in the literature.

In the case of a five-phase VSI, there are in total $2^5=32$ space vectors available, of which thirty are active state vectors and two are zero state vectors forming three concentric decagons. (Toliyat et al 2000), (Xu et al 2002) and Shi and Toliyat (2002) have used only ten outer large length vectors to implement the symmetrical SVPWM. Two neighbouring active space vectors and two zero space vectors are utilised in one switching period to synthesise the input reference voltage. In total, twenty switching take place in one switching period, so that state of each switch is changed twice. The switching is done in such a way that, in the first switching half-period the first zero vector is applied, followed by two active state vectors and then by the second zero state vector. The second switching half-period is the mirror image of the first one. The symmetrical SVPWM is achieved in this way. This method is the simplest extension of space vector modulation of three-phase VSIs.

An ideal SVPWM of a five-phase inverter should satisfy a number of requirements. First of all, in order to keep the switching frequency constant, each switch can change state only twice in the switching-period (once 'on' to 'off' and once 'off' to 'on', or vice-versa). Secondly, the RMS value of the fundamental phase voltage of the output must equal the RMS of the reference space vector. Thirdly, the scheme must provide full utilisation of the available dc bus voltage. Finally, since the inverter is aimed at supplying the load with sinusoidal voltages, the low-order harmonic content needs to be minimised (this especially applies to the third and seventh harmonic). These criteria are used in assessing the merits and demerits of various SVPWM.
Conventionally two different SVPWM schemes are considered one which utilizes the outermost large set of space vectors in the $d-q$ plane and other uses four neighbouring vectors two from large set and two from the middle set. The times of active space vector application are given as Iqbal and Levi (2006).

\[
t_a = \frac{|v_r| \sin(k\pi/5 - \alpha)}{|v| \sin(\pi/5)} t,
\]

\[
t_b = \frac{|v_r| \sin(\alpha - (k-1)\pi/5)}{|v| \sin(\pi/5)} t,
\]

\[
t_o = t_o - t_a - t_b.
\]

Here $k$ is the sector number ($k = 1$ to 10), and large vector length is $|v_\alpha| = |v_\mu| = |v_r| = \frac{2}{5} V_{\text{dc}} 2 \cos(\pi/5)$ Corresponding medium vector length, which will be needed in subsequent expression, is $|v_{\alpha m}| = |v_{\mu m}| = V_{\text{DC}} = (2/5) V_{\text{dc}}$. Symbol $v_r$ denotes the reference space vector, while $|x|$ is the modulus of a complex number $x$. Switching period is denoted with $t$, and indices $a$ and $b$ denote the neighbouring space vectors to the right and to the left, respectively, of the reference space vector. Indices $l$ and $m$ stand for large and medium space vectors, respectively. The largest possible fundamental peak output voltage that can be achieved using this scheme corresponds to the radius of the largest circle that can be inscribed within the decagon. This maximum fundamental peak output voltage $V_{\text{max}}$ is

\[
V_{\text{max}} = \frac{2}{5} 2 \cos(\pi/5) \cos(\pi/10) V_{\text{dc}} = 0.61554 V_{\text{dc}}.
\]

Application of two adjacent medium active space vectors together with two large active space vectors in each switching period makes it possible to maintain zero average value in the second plane Ryu et al (2005) and consequently providing sinusoidal output. Use of four active space vectors per switching period requires calculation of four application times, labeled here $t_{la}, t_{lb}, t_{ma}, t_{mb}$. The expressions used for calculation of dwell times of various space vectors are [Silva et al (2004)],

\[
t_{la} = \frac{|v_r|}{V_m \sin(\pi/5)} \left( 1 + \frac{\tau}{\tau} \right) t, \sin\left( \frac{\pi}{5} k - \alpha \right)
\]

\[
t_{lb} = \frac{|v_r|}{V_m \sin(\pi/5)} \left( 1 + \frac{\tau}{\tau} \right) t, \sin\left( \alpha - (k-1)\frac{\pi}{5} \right)
\]
where; $t_a = t_{aw} + t_{aw}'$, $t_b = t_{am} + t_{am}'$. This is in essence allocates 61.8% more dwell times to large space vectors compared to medium space vector thus satisfying the constraints of producing zero average voltage in the $x$-$y$ plane. The maximum possible output with this approach is $0.5257V_{DC}$ which is almost 16% less than the previous method.

The simulation results of SVPWM method are provided with the ANN based method for comparison purpose.

### 3.3.4 TIME EQUIVALENT SPACE VECTOR PULSE WIDTH MODULATION SCHEME (TESVPWM)

The proposed SVPWM called here time equivalent space vector PWM (TESVPWM) utilises simply the sampled reference voltages to generate the gating time for which each inverter leg to yield sinusoidal output. This method is an extension of the technique developed for a three-phase VSI, Chung et al (1998). The reference voltage is sampled at fixed time interval equal to the switching time. The sampled amplitude is converted to equivalent time signal. The time signals thus obtained are imaginary quantities as they will be negative for negative reference voltage amplitude. Thus a time offset is added to these signals to obtain the gating time of each inverter leg. This offset addition centres the active switching vectors within the switching interval. The procedure is outlined in the Fig. 3.9 for sector I and the same approach can be used for other 9 sectors. The corresponding switching pattern for first part is illustrated in Fig. 3.10. It is evident that the switching is similar to the one used in SVPWM. The simulation results are shown in Fig. 3.11.

#### 3.3.4.1 Algorithm of the proposed TESVPWM:

1. Sample the reference voltages $V_a, V_b, V_c, V_d$ & $V_e$ in every switching period $T_s$
2. Determine the equivalent times $T_i, T_2, T_3, T_4$ & $T_5$ given by

$$
\tau_{xx} = V_{xx} \times \frac{T_s}{v_{dc}}, \quad x = a, b, c, d \text{ and } e
$$

3. Determine $T_{offset}$

$$
T_{offset} = \frac{T_s}{2} - \frac{T_{max} \times T_{min}}{v_{dc}}
$$
4. Then the inverter leg switching times are obtained as

\[ T_{gx} = T_x + T_{offset}, \quad x = a, b, c, d \text{ and } e \]

For Sector 1

\[ T_{max} = T_a; \quad T_{min} = T_d; \]

\[ T_i = T_a - T_b; \quad T_2 = T_b - T_c; \quad T_3 = T_c - T_d; \quad T_4 = T_d - T_a; \]  

\[ T_{effective} = T_{max} - T_{min} = T_a - T_d; \]

\[ T_0 = T_s - T_{effective}; \]

\[ T_{offset} = \frac{T_0}{2} - T_{min} = \frac{T_0}{2} - T_d \]

\[ T_{ga} = T_a + T_{offset} = T_a + \frac{T_0}{2} - T_{min} = T_a + \frac{T_0}{2} - T_d = \frac{T_0}{2} + T_1 + T_2 + T_3 + T_4 \]

\[ T_{gb} = T_b + T_{offset} = T_b + \frac{T_0}{2} - T_{min} = T_b + \frac{T_0}{2} + T_1 + T_2 + T_3; \]

\[ T_{gc} = T_c + T_{offset} = T_c + \frac{T_0}{2} - T_{min} = \frac{T_0}{2} + T_4; \]

\[ T_{gd} = T_d + T_{offset} = T_d + \frac{T_0}{2} - T_{min} = \frac{T_0}{2}; \]

\[ T_{ge} = T_e + T_{offset} = T_e + \frac{T_0}{2} - T_{min} = \frac{T_0}{2} + T_2 + T_3 + T_4; \]

\[ T_1 = T_{ga} - T_{gb}; \quad T_2 = T_{gb} - T_{ge}; \quad T_3 = T_{ge} - T_{gd}; \quad T_4 = T_{gd} - \frac{T_0}{2} = T_{gd}; \]

Fig. 3.10 Switching waveforms for sector 1 using the proposed TESVPWM
Fig. 3.9 Principal of TESVPWM for sector 1

(a) Harmonic spectrum phase 'a' voltage

(b) Filtered output voltage
Fig. 3.11 Simulation results of the proposed TESVPWM based scheme

Fig. 3.11(a) shows the harmonic spectrum for the output phase ‘a’ voltage with FFT and Fig. 3.11(b) shows the filtered output voltage after connecting a RL load at the output terminals. Fig. 3.11(c) shows the offset time signals as calculated after the mathematical analysis, it shows both the maximum value as well as the minimum value for the offset and the offset time signal. Fig. 3.11(d) shows the net modulating signals after adding the offset signal to equivalent time signal for each phase.

3.3.5 ARTIFICIAL NEURAL NETWORK BASED SPACE VECTOR PULSE WIDTH MODULATION.

The space vector PWM technique for a five-phase isolated neutral load of a voltage fed inverter is already discussed. Instead of implementing SVM by DSP, it is possible to implement it by a feed-forward neural network because the SVM algorithm can be looked upon as a nonlinear input/output mapping. Bhul and Lorenz (1991), Marchesi et al (1993), Pinto et al (2000), Simon (2004), Bakshai et al (1996), Brock et al (1998), Dzung et al (2006), Himavathi and Muthuramalingam (2007), Mondal et al (2002). This means that the reference voltage vector $V_r$ magnitude, and $\alpha_r$ angle can be impressed at the input of the network and the corresponding pulse width pattern of the five phases can be generated at the output. There are two approach of implementing SVM using ANN called ‘Direct method’ and ‘Indirect method’. In the so called ‘Direct method’, the feedforward backpropagation ANN directly replace the conventional SVM algorithm. Since feedforward ANN network can map only one input pattern onto only one input pattern, the sampling time is divided into $n$ subintervals. Thus each subinterval includes only one output switching pattern for every
input pattern. Thus it requires huge data set for proper training of the network. Thus this approach is limited in use. The later method uses two separate feedforward backpropagation ANN, one for the magnitude of the reference voltage and other for the reference voltage position. The magnitude network yield voltage magnitude scaling function which is linear in the linear modulation region and is non-linear function of $V_{DC}$ in the overmodulation region. The reference position network yield turn on pulse width function at unit voltage magnitude. This pulse width functions are then multiplied by a suitable bias signal and the product is compared with the up/down counter to generate appropriate switching signals for the inverter. The complete implementation block diagram is illustrated in Fig. 3.12.

![Fig. 3.12 Functional Block diagram of ANN based SVM for a five-phase VSI.](image)

### 3.3.5.1 ANN based SVM Using Large Vector (Over modulation Region)

As discussed in the previous section, there exist two conventional methods for realizing space vector PWM in a five-phase VSI. The method of using large vectors only generates higher output but the output is polluted with low-order harmonics. Thus it is proposed in this paper to use this approach above 0.5257$V_{DC}$ where the sinusoidal method fails to be implemented and it is called over modulation region in this paper. Nevertheless, this technique is simpler to implement compared to linear modulation method contrary to three-phase SVM where implementing over modulation is cumbersome. The regions of linear modulation and over modulation are depicted in Fig. 3.13.
Chapter 3: Five-phase Voltage Source Inverter - Modelling and Control for Single Motor Drive

Hence at first the technique for overmodulation is discussed followed by the method of producing sinusoidal output or liner modulation region.

The turn on time $T_{A\text{-ON}}$ can be derived from equations (3.27) and (3.28) for sector 1 as

$$T_{A\text{-ON}} = \frac{t_0}{2} = \frac{T_s}{4} + K.V^* \left[ -\sin\left(\frac{\pi}{5} - k\alpha^*\right) -\sin\alpha^* \right]$$

(3.30)

Where $K = \frac{T_s}{(4V_{DC})\left[\frac{\pi}{5}\sin\left(\frac{\pi}{10}\right) + 1.618.2 \sin\left(\frac{\pi}{10}\right)\right]}$.

Similar timing intervals can be derived for all ten sectors, and correspondingly, the phase-a turn on time can be expressed as

$$T_{A\text{-ON}} = \begin{cases} \frac{t_0}{2} = \frac{T_s}{4} + K.V^* \left[ -\sin\left(\frac{\pi}{5} - \alpha^*\right) -\sin\alpha^* \right] & \text{.........} S = 1, 2, 9, 10 \\ \frac{t_0}{2} + t_a + t_b = \frac{T_s}{4} + K.V^* \left[ -\sin\left(\frac{\pi}{5} - \alpha^*\right) +\sin\alpha^* \right] & \text{.........} S = 3 \\ \frac{t_0}{2} + t_a + t_b = \frac{T_s}{4} + K.V^* \left[ -\sin\left(\frac{\pi}{5} - \alpha^*\right) +\sin\alpha^* \right] & \text{.........} S = 4, 5, 6, 7 \\ \frac{t_0}{2} + t_a = \frac{T_s}{4} + K.V^* \left[ -\sin\left(\frac{\pi}{5} - \alpha^*\right) -\sin\alpha^* \right] & \text{.........} S = 8 \end{cases}$$

(3.31)

where the sector numbers are indicated on the right side of the equation. The turn on time for other four phases can be drawn in similar fashion. It is to be noted that the turn on time of different phases have a phase difference of 72 degrees. Because of symmetry, the corresponding turn-off time is given as

$$T_{A\text{-OFF}} = T_s - T_{A\text{-ON}}$$

(3.32)

Equation (8) can be written in general form

$$T_{A\text{-ON}} = \frac{T_s}{4} + f(V^*). g(\alpha^*)$$

(3.33)

Where $f(V^*) =$ voltage amplitude scale factor and
Chapter 3: Five-phase Voltage Source Inverter: Modelling and Control for Single Motor Drive

\[ g_A(\alpha^*) = \begin{cases} 
K.[-\sin\left(\frac{\pi}{5} - \alpha^*\right) - \sin \alpha^*] & \text{... ... ... ... } S = 1, 2, 9, 10 \\
K.[-\sin\left(\frac{\pi}{5} - \alpha^*\right) + \sin \alpha^*] & \text{... ... ... ... } S = 3 \\
K.[\sin\left(\frac{\pi}{5} - \alpha^*\right) + \sin \alpha^*] & \text{... ... ... ... } S = 4, 5, 6, 7 \\
K.[\sin\left(\frac{\pi}{5} - \alpha^*\right) - \sin \alpha^*] & \text{... ... ... ... } S = 8 
\end{cases} \] (3.34)

Where \( g_A(\alpha^*) \) is called turn on pulsewidth function at unit voltage magnitude \( f(V^*) = 1 \).

In the under modulation region the scale factor is linear, that is, \( f(V^*) = V^* \) for dc voltage and is a non-linear function in over modulation region. The turn on time of phase a can be plotted as shown in Fig. 3.14.

It is clearly seen from the Fig. 3.14, that the operational region of this technique is limited upto \( V^* = 0.6115V_{DC} \) where the turn-on time equals to \( Ts/2 \) and if the reference magnitude is further increased the application time becomes negative (dotted curve) which is not physically realizable. The lower operational limit of this technique is shown by another bold line below the \( Ts/2 \) line where the input reference voltage equal \( V^* = 0.5257V_{DC} \). At this limiting value the sinusoidal output method described in the next section becomes inoperative.

**3.3.5.2 ANN Based SVM Using Large and Medium vector (Linear Modulation)**

This section develops space vector PWM technique for obtaining sinusoidal output within the linear modulation region. This method utilises four neighbouring vectors as explained in the previous section.
Therefore, time $T_{A-ON}$ can be derived from equations (6) for sector 1 as

$$T_{A-ON} = \frac{t_a}{2} = \frac{T_S}{4} + K.V^* \left[ - \sin \left( \frac{\pi}{5} - \alpha^* \right) - k \cdot \sin \alpha^* - k \cdot \sin \left( \frac{\pi}{5} - \alpha^* \right) - \sin \alpha^* \right] \quad (3.35)$$

Where $K = \frac{0.8541 \cdot T_S}{(\pi V_{DC}^2 \sin \left( \frac{\pi}{5} \right) \cdot (1+1.618^2) \cdot 2 \sin \left( \frac{\pi}{10} \right) )}$, similar timing intervals can be derived for all ten sectors, and correspondingly, the phase-a turn on time can be expressed as

$$T_{A-ON} = \begin{cases} 
\frac{t_a}{2} = \frac{T_S}{4} + K.V^* \left[ - \sin \left( \frac{\pi}{5} - \alpha^* \right) - k \cdot \sin \alpha^* - k \cdot \sin \left( \frac{\pi}{5} - \alpha^* \right) - \sin \alpha^* \right], & S = 1, 10 \\
\frac{t_a}{2} + t_{am} = \frac{T_S}{4} + K.V^* \left[ - \sin \left( \frac{\pi}{5} - \alpha^* \right) - k \cdot \sin \alpha^* - k \cdot \sin \left( \frac{\pi}{5} - \alpha^* \right) + \sin \alpha^* \right], & S = 2 \\
\frac{t_a}{2} + t_{am} + t_{bl} = \frac{T_S}{4} + K.V^* \left[ - \sin \left( \frac{\pi}{5} - \alpha^* \right) + k \cdot \sin \alpha^* - k \cdot \sin \left( \frac{\pi}{5} - \alpha^* \right) - \sin \alpha^* \right], & S = 3 \\
\frac{t_a}{2} + t_{am} + t_{bl} + t_{al} = \frac{T_S}{4} + K.V^* \left[ - \sin \left( \frac{\pi}{5} - \alpha^* \right) + k \cdot \sin \alpha^* + k \cdot \sin \left( \frac{\pi}{5} - \alpha^* \right) + \sin \alpha^* \right], & S = 5, 6 \\
\frac{t_a}{2} + t_{bl} + t_{al} + t_{bm} = \frac{T_S}{4} + K.V^* \left[ \sin \left( \frac{\pi}{5} - \alpha^* \right) + k \cdot \sin \alpha^* + k \cdot \sin \left( \frac{\pi}{5} - \alpha^* \right) - \sin \alpha^* \right], & S = 7, 8 \\
\frac{t_a}{2} + t_{al} + t_{bm} = \frac{T_S}{4} + K.V^* \left[ - \sin \left( \frac{\pi}{5} - \alpha^* \right) + k \cdot \sin \alpha^* + k \cdot \sin \left( \frac{\pi}{5} - \alpha^* \right) + \sin \alpha^* \right], & S = 9 \\
\frac{t_a}{2} + t_{bm} = \frac{T_S}{4} + K.V^* \left[ \sin \left( \frac{\pi}{5} - \alpha^* \right) - k \cdot \sin \alpha^* - k \cdot \sin \left( \frac{\pi}{5} - \alpha^* \right) - \sin \alpha^* \right], & S = 10 \\
\end{cases} \quad (3.36)$$

Where the sector number is indicated on the right. Similar expressions can be derived for other four phases however, they will have a phase difference of 72 degrees. Because of symmetry, the corresponding turn-off time is given as

$$T_{A-OFF} = T_S - T_{A-ON} \quad (3.37)$$

Equation (3.33) can be written in general form

$$T_{A-ON} = \frac{T_S}{4} + f(V^*) \cdot g(\alpha^*) \quad (3.38)$$

Where $f(V^*) =$ voltage amplitude scale factor and
\[ g_A(\alpha^*) = \begin{cases} 
K \left[ -\sin \left( \frac{\pi}{5} - \alpha^* \right) - k \sin \alpha^* - k \sin \frac{\pi}{5} - \alpha^* \right] - \sin \alpha^* \ldots \ldots S = 1,10 \\
K \left[ -\sin \left( \frac{\pi}{5} - \alpha^* \right) - k \sin \alpha^* - k \sin \left( \frac{\pi}{5} - \alpha^* \right) + \sin \alpha^* \right] \ldots S = 2 \\
K \left[ -\sin \left( \frac{\pi}{5} - \alpha^* \right) + k \sin \alpha^* - k \sin \left( \frac{\pi}{5} - \alpha^* \right) - \sin \alpha^* \right] \ldots \ldots S = 3 \\
K \left[ -\sin \left( \frac{\pi}{5} - \alpha^* \right) + k \sin \alpha^* + k \sin \left( \frac{\pi}{5} - \alpha^* \right) + \sin \alpha^* \right] \ldots \ldots S = 4 \\
K \left[ \sin \left( \frac{\pi}{5} - \alpha^* \right) + k \sin \alpha^* + k \sin \left( \frac{\pi}{5} - \alpha^* \right) + \sin \alpha^* \right] \ldots \ldots S = 5,6 \\
K \left[ \sin \left( \frac{\pi}{5} - \alpha^* \right) + k \sin \alpha^* + k \sin \left( \frac{\pi}{5} - \alpha^* \right) - \sin \alpha^* \right] \ldots \ldots S = 7 \\
K \left[ -\sin \left( \frac{\pi}{5} - \alpha^* \right) - k \sin \alpha^* + k \sin \left( \frac{\pi}{5} - \alpha^* \right) + \sin \alpha^* \right] \ldots S = 8 \\
K \left[ \sin \left( \frac{\pi}{5} - \alpha^* \right) - k \sin \alpha^* - k \sin \left( \frac{\pi}{5} - \alpha^* \right) - \sin \alpha^* \right] \ldots \ldots S = 9 
\end{cases} \] (3.39)

Which is defined as the pulse width function at unit amplitude \( f(V^*) = 1 \). In the under modulation region the scale factor is linear, that is \( f(V^*) = V^* \) for dc voltage \( V_{dc} \) and is a non-linear function in over modulation region. The turn on time of phase \( a \) can be plotted as shown in Fig. 3.15.

![Fig. 3.15 Turn-on time of phase as a function of reference vector position in the ten sectors](image)

It is clearly seen from the Fig. 3.15, that the operational region of this technique is limited up to \( V^* = 0.5257V_{dc} \) where the turn-on time equals to \( Ts/2 \) and if the reference magnitude is further increased the application time becomes negative (dotted curve) which is not physically realizable.

A simulation model is developed in Matlab/Simulink to simulate the ANN based SVPWM for a five-phase VSI. A different approach is adopted in this paper to implement the ANN SVM. Here voltage reference magnitude network is not used from ANN, instead a
linear voltage magnitude multiplier is used. This approach simplifies the implementation and significantly reduces the computation times. The expressions of turn-on times and the corresponding turn-on pulse width function as illustrated in the preceding sections, allows ANN based space vector PWM implementation using only one subnet (Fig. 3), the angle subnet. The angle subnet inputs the reference voltage vector position $a^*$ and correspondingly generates the turn-on pulse width functions $g_i(a^*), i = A, B, C, D, E$. The angle subnet has three layers, the input, the hidden and the output with (1-15-5) neurons and using sigmoidal processing function for hidden and output layers as illustrated in Fig. 3.16. The angle training of the network was performed in the full cycle with an increment of 1 degree.
The simulation results are obtained for conventional space vector PWM and ANN based space vector PWM and they are plotted on the same curve for comparison purposes. The resulting waveforms are depicted in Fig. 3.17-3.18 for linear modulation and Fig. 3.19-3.20 for overmodulation. The switching frequency is kept equal to 10 kHz, the dc link voltage is assumed unity and the fundamental frequency is taken as 50 Hz. The results are illustrated for maximum modulation for both cases.
Fig. 3.17 Simulation result linear modulation:
a. Filtered leg voltage.
b. Filtered phase voltage.
c. Filtered Common mode voltage.

Fig. 3.18 Voltage spectrum linear modulation: a. ANN SVM. b. Conventional SVPWM
Chapter 3: Five-phase Voltage Source Inverter-Modelling and Control for Single Motor Drive

Fig. 3.19 Simulation result over modulation:
(a) Filtered leg voltage.
(b) Filtered phase voltage.
(c) Filtered Common mode voltage.

Fig. 3.20 Voltage spectrum over modulation: a. ANN SVM. b. Conventional SVPWM

The results shown in Fig. 3.17 and Fig. 3.18 depict a very good tracking of the reference voltage, both the leg and phase voltages. The results obtained using ANN and conventional SVPWM shows similar behaviour. However, there a marginal gain in the ANN based technique as suggested by the THD value of the spectrum refers to Fig. 3.19. The THD of
the phase voltage using ANN is 1.39% considering up to 20\textsuperscript{th} harmonics while it is 1.52% using conventional SVPWM at 10 kHz switching frequency.

The results obtained in over modulation region suggest once again a good tracking of the reference using ANN. The results are similar compared to the conventional method, nevertheless a marginal gain in terms of lower THD is observer in this case also. It is seen from Fig. 3.20 that the THD for ANN method is 29.25% and with that of conventional method is 29.42%.

3.4 EXPERIMENTAL INVESTIGATION

A Five-phase voltage source inverter is developed using intelligent power module from VI Micro systems, Chennai. Texas Instrument DSP TMS320F2812 is used as the processor to implement the control algorithm. A RS232 cable is used to transfer the signals generated using PC to DSP board. The complete control code is written in C++ which is compiled using Code Composer Studio 3.1 and ASCI file is transferred to DSP using the printer port of the PC and they have dedicated 16 hardware PINS to generate the desired PWM signals. The PWM circuits associated with compare units make it possible to generate up to eight PWM output channels (per Event Manger) with programmable dead band and polarity. This DSP is specifically meant for use in motor drive purposes and it can control up to 8-phase two-level inverter. The control code is written in C++ language in Code composer studio 3.1 which runs in a PC. The control signal generated by PC is transferred to the DSP board through RS 232 cable connected in parallel printer port of the PC. The DSP board is connected to the Power Module through dedicated control cable. The DSP interfacing circuit along with required A/D and D/A converter is built on the DSP board itself procured from VI Micro systems. The complete experimental set up is shown in Fig. 3.21. The experimental results obtained are illustrated in Fig. 3.22.

Switching patterns for the proposed PWM are shown in Fig. 3.22. The resulting waveform of leg voltage and harmonic spectrum of phase voltage (filtered values, C= 0.1 \textmu F, R = 20 k\Omega in parallel) are shown in Fig. 3.23 respectively. The switching frequency is 5 KHz and the output voltage frequency is 50 Hz.
Fig. 3.21 Five-phase experimental set up.

Fig. 3.22 Switching pattern for the proposed PWM

Fig. 3.23 Filtered output voltages and the harmonic spectrum for the proposed PWM