CHAPTER 1

INTRODUCTION

1.1 GENERAL

The impact of power electronics is felt in several fields of electrical engineering, such as electric drives, flexible AC transmission systems, and uninterrupted power supplies. Power electronics plays a major role in energy conservation and development of the power industry. The need for power electronics has increased over the years due to its advantages in the control of electrical energy. This surge is due to several factors; the most important being the technological advancements in the semiconductor devices industry, which has led to the introduction of high-speed and high-power integrated devices. Other factors include: (i) the advances made in the area of microelectronics that have paved the way for the development of efficient integrated circuits and growth of processors, and (ii) the ever increasing need for smaller size and lighter weight power electronic components. However, these power electronics devices impose some serious problems on the quality of power. The increasing reliance on the power converters and the need to provide quality power has mandated that all such power electronics systems should have low harmonic content, less Total Harmonic Distortion (THD), and improved input power factor.

This chapter gives a brief overview of the static AC-to-AC power and frequency conversion structures and introduces the structure and operation of the matrix converter, which is the topic of this thesis. Later, the
literature review and a new error integrative algorithm suitable for high frequency switched three-phase-to-three-phase matrix converter are presented along with the objective and organization of the thesis.

1.2 STATE-OF-THE-ART REVIEW OF THE THREE-PHASE AC-TO-AC CONVERTER

The basic function of a power converter is to process the energy using the power switches. AC power conversion can be classified as (i) direct AC-to-AC and (ii) indirect AC-to-DC-to-AC. At present, voltage source back-to-back converters (AC-DC-AC) are widely used in the AC conversion systems. These converters require bulky storage electrolytic capacitors. The operation of these converter stages (stage-1: AC-DC) and (stage-2: DC-AC) is controlled independently since they are decoupled by means of an energy storage element. Therefore, the instantaneous input power need not be equal to the instantaneous output power. The difference between the instantaneous input power and the instantaneous output power is absorbed or delivered by an energy storage element within the converter. The storage element is the bulky storage electrolytic capacitor prone to poor performance at high temperatures and susceptible to failures (Hitachi inverter instruction manual). Therefore, the use of the storage capacitor not only increases the system weight and volume but also results in reduced reliability of the system.

The energy storage element is not required in a direct converter (Gyugyi and Pelly 1976). Because of the absence of the energy storage element, the instantaneous power input must be equal to the instantaneous power output. However, the reactive power input need not be equal to the reactive power output. Figure 1.1 shows the classification of the direct AC-to-AC converter into three distinct topological approaches. The first approach is the AC voltage regulator that changes the amplitude of the AC waveform (Hashem and Darwish 2004). The second approach is the cyclo
converter, which is used if the required output frequency (Maamoun 2003) is much lower than the input source frequency. The third approach is the matrix converter, which is the most versatile and has no limits on the output frequency and the amplitude. In other words, the input may be three-phase AC and the output DC, or both may be DC, or both may be AC (Mohan et al 2003). Therefore, the matrix converter topology is promising for the universal power conversion such as AC to DC, DC to AC, DC to DC or AC to AC. The matrix converter offers some significant advantages such as adjustable power factor, inherent four-quadrant operation, high quality sinusoidal input/output waveforms and high power density. Hence, it has received extensive attention in research as a replacement for the traditional AC-DC-AC converter for variable-voltage and variable-frequency AC drive applications.

Figure 1.1 AC-to-AC converter topologies
1.3 STRUCTURE OF THE MATRIX CONVERTER

The matrix converter is the force-commutated version of the cyclo-converters (Huber and Borojevic 1989), which overcomes the disadvantage of the conventional cyclo-converter such as the limitations in the frequency conversion, rich output voltage harmonics and increased number of switches (Rashid 2005 and Fa-Hai Li et al 1994). The matrix converters can be classified as direct and indirect type matrix converters. Figure 1.2 shows the direct or the conventional matrix converter (CMC) that is an array of $3 \times 3$ bidirectional switches. The indirect or the sparse matrix converter is a cascade of the controlled rectifier and inverter topologies without a DC link in between (Ziogas et al 1986). Both the topologies directly interconnect two independent multi-phase voltage systems at different frequencies. In this research, the CMC topology is chosen and is analyzed for its performance for changes in its topology and with different pulse with modulation (PWM) techniques.

Figure 1.2 Structure of the conventional matrix converter
The matrix converter is connected to a stiff voltage source at the input and a stiff current source at the output. These externally connected voltage and current sources impose constraints on the switching of the matrix converter. At any instant, the voltage source should not be short-circuited and the current source should not be open circuited. Equation (1.1) gives the switching function of the switch $S_{ij}$ in Figure 1.2.

$$S_{ij}(t)=1, \quad S_{ij}=\text{closed}$$

$$S_{ij}(t)=0, \quad S_{ij}=\text{open, \ where \ } i \in \{A,B,C\} \text{ \& } j \in \{a,b,c\} \quad (1.1)$$

Equation (1.2) gives the constraints namely that the inputs are not short-circuited and the outputs are not open-circuited

$$d_{Aj}+d_{Bj}+d_{Cj}=1 \quad (1.2)$$

where, $d_{ij}$ is the duty-cycle of the switch $S_{ij}$. Equation (1.2) being less than one indicates an open-circuit of the current source and Equation (1.2) being greater than one indicates a short-circuit of the voltage source.

Therefore, Equations (1.3) and (1.4) represents the switching function $T$ of the matrix converter for the output voltages and the input currents

$$V_{out} = T \times V_{in} \quad (1.3)$$

where,

$$V_{out} = \begin{bmatrix} V_a \\ V_b \\ V_c \end{bmatrix}, \quad T = \begin{bmatrix} S_{Aa} & S_{Ab} & S_{Ac} \\ S_{Ba} & S_{Bb} & S_{Bc} \\ S_{Ca} & S_{Cb} & S_{Cc} \end{bmatrix} \quad \text{and} \quad V_{in} = \begin{bmatrix} V_A \\ V_B \\ V_C \end{bmatrix}$$

$$I_{in} = T^T \times I_{out}$$
\[
\begin{bmatrix}
I_A \\
I_B \\
I_C \\
\end{bmatrix} =
\begin{bmatrix}
S_{Aa} & S_{Ba} & S_{Ca} \\
S_{Ab} & S_{Bb} & S_{Cb} \\
S_{Ac} & S_{Bc} & S_{Cc} \\
\end{bmatrix}
\begin{bmatrix}
I_a \\
I_b \\
I_c \\
\end{bmatrix}
\] (1.4)

where,

\[
I_{in} =
\begin{bmatrix}
I_A \\
I_B \\
I_C \\
\end{bmatrix},
T^T =
\begin{bmatrix}
S_{Aa} & S_{Ba} & S_{Ca} \\
S_{Ab} & S_{Bb} & S_{Cb} \\
S_{Ac} & S_{Bc} & S_{Cc} \\
\end{bmatrix}
\text{ and } I_{out} =
\begin{bmatrix}
I_a \\
I_b \\
I_c \\
\end{bmatrix}
\]

\(V_A, V_B, V_C\) and \(V_a, V_b, V_c\) are the input and the output phase voltages respectively and \(I_A, I_B, I_C\) and \(I_a, I_b, I_c\) are the input and the output currents respectively.

1.3.1 Basic Components of a Practical Matrix Converter

The design of a practical matrix converter circuit shown in Figure 1.2 involves the development of three essential circuits (i) the power circuit, (ii) the input filter circuit and (iii) the clamp circuit.

The power circuit consists of nine bi-directional switches shown in Figure 1.2 with \(2^9 \times 512\) possible switching states. However, only 27 switching states are used because of the limitations imposed on the power circuit by the Equation (1.2).

The input filter is necessary to reduce the switching harmonics in the input currents (Wheeler et al 2002a). Figure 1.2 shows the input filter consisting of the source inductances \(L_A, L_B, L_C\) and the source capacitors \(C_A, C_B, C_C\).

The interruption of the inductive current during commutations leads to high voltage spikes appearing across the switches. These high voltage spikes damage the switches, and therefore a clamp circuit, shown in Figure 1.2, is required to store the inductive energy (Klumpner and Blaabjerg
2002). The clamp circuit transfers the inductive energy from the load to the clamp capacitor during the turning OFF of the converter.

1.3.2 Bidirectional Switch Configurations

The power electronics realization of the matrix converter in Figure 1.2 requires four quadrant bidirectional switches. Due to the lack of semiconductor devices capable of operating in four quadrants, four quadrant switches are constructed with two quadrant switches (Dynex Semiconductor 2007), as shown in Figure 1.3.

![Diagrams of four quadrant bidirectional switches](image)

**Figure 1.3** Four quadrant bidirectional switches (a) diode-embedded switch (b) reverse blocking IGBT (c) common emitter IGBT and (d) common collector IGBT

The main advantage of the diode-embedded switch is its simple configuration when compared to other bi-directional switch structures but the main disadvantage of this configuration is the high conduction loss. The other configurations have lower conduction losses than the diode-embedded switch. The reverse blocking IGBTs are not often used as bidirectional switches since, in practice, the IGBT has poor reverse recovery characteristics that increases the switching losses and hence the overall system efficiency decreases. Amongst the other two configurations, the common emitter configuration is widely used for high power applications (Semikron-SK60GM123 2007). Common collector configuration although
requiring a reduced number of isolated power supplies for generating the switching signals is not much preferred for high power applications (Imayavaramban 2008).

1.4 COMMUTATION TECHNIQUES FOR THE MATRIX CONVERTERS

With the constraints imposed on the matrix converter, as explained in Equation (1.2), it is found that each output phase of the matrix converter must always be connected to one and only one input phase even during the commutation. Since the matrix converter does not have an inherent freewheeling path, the commutation of its bidirectional switches is much more difficult than the commutation of switches in an inverter.

Various commutation techniques have been proposed for matrix converters namely (i) the dead-time commutation, (ii) the soft switching technique, (iii) the multi-step current commutation and (iv) the multi-step voltage commutation.

The implementation of the dead-time commutation in matrix converters leads to the interruption of inductive currents due to the absence of a freewheeling path. Use of such technique in matrix converters requires snubber circuits to provide an alternate path to the inductive currents (Sunter and Clare 1996), which increases the complexity and the size of the converters.

Soft switching techniques have been investigated in many converter topologies for reducing the switching losses. However, the implementation of soft switching techniques in matrix converters (Marcks 1995) increases the component count and complexity of the converters.
At present, the most popular methods of commutation are the multi-step current and multi-step voltage commutations. The idea of such multi-step commutations techniques first appeared in Oyama et al (1989) and Burany (1989).

1.4.1 The Four-Step Current Commutation

To determine the commutation sequence, the method relies on the direction of the load current. The bidirectional switch (BS) that will stop conducting after the commutation is known as the outgoing BS and the BS that will start conducting after the commutation is known as the incoming BS. Each BS consists of two switches, named as $S^+$ and $S^-$, which indicates the direction of the current flow through the respective switches.

Before the commutation process starts, both the switches $S^+$, $S^-$ of the outgoing BS are ON. With this condition, the safe commutation sequence of the BS to transfer the load current from one phase to another phase is explained below.

![Figure 1.4 Commutation of the bidirectional switches between the two input phases](image-url)
Step 1: When the commutation to an incoming BS is required, the current direction is used to determine the non-conducting switch in the outgoing BS. This switch is first turned OFF.

Step 2: Then, the switch in the incoming BS that would conduct the current in the same direction is turned ON. This is done to form a path for the load current to continue flowing either at the point when the next switch of the incoming BS is gated ON or when the conducting switch of the outgoing BS is turned OFF.

Step 3: The conducting switch of the outgoing BS can now be turned OFF safely, since a new path is made available for the current to flow as in Step 2.

Step 4: Finally, the other switch of the incoming BS is switched ON to complete the sequence of commutation.

For Example, in the Figure 1.4, when the $i_L > 0$,

(i) Switch $S_1^{-}$ is turned OFF, (ii) Switch $S_2^{+}$ is turned ON, (iii) Switch $S_1^{+}$ is turned OFF, and (iv) Switch $S_2^{-}$ is turned ON.

Similarly, the four-step current commutation sequence of switching can be formulated for other cases and is given in Figure 1.5.

![Figure 1.5 State diagram of the four-step current commutation](image-url)
Since the commonly used Hall effect sensors are prone to produce uncertain results in high power and low current applications, it is difficult to reliably determine the direction of current for commutation. To avoid this problem, a technique named as voltage commutation that uses the voltage across the bidirectional switch for measurement of the direction of current (Wheeler et al 2002b) has been developed. Later, a technique utilizing the zero vectors (Mahlein et al 2002) to avoid commutation error, when the line voltage is zero, was proposed for robust commutation. However, the input current in this technique was found distorted compared to the voltage and current commutation techniques because of utilizing a different switching sequence.

1.5 MODULATION TECHNIQUES FOR THE MATRIX CONVERTERS

Several modulation algorithms are reported for matrix converters in Wheeler (2002a) to achieve different control objectives; basic classifications of these modulation techniques are shown in Figure 1.6.

![Figure 1.6 Classification of the matrix converter modulation techniques](image-url)
The development of this converter and its modulation techniques started three decades back based on the complex mathematical formulation by Venturini and Alesina (1980) with the voltage transfer ratio of 0.5. Later, the Optimum Alesina Venturini (OAV) method (Alesena and Venturini 1989) was proposed, in which the modulation index was extended from 0.5 to 0.866 by using the third harmonic injection technique. It was also proved that the modulation index of 0.866 is the physical limitation for the matrix converter. The carrier based PWM method with varying amplitude triangular carrier was proposed by Yoon and Sul (2006). Later, Thuta (2007) proposed a simplified carrier PWM technique. Control technique for space vector control of the matrix converter was proposed by Huber and Borojevic (1995). Using the idea of the ‘fictitious DC Link’, a conceptually different idea (Casadei et al 2002), decoupled the control into smaller independent units. Researchers of matrix converter now predominantly use this technique. Modulation algorithm using MAX-MID-MIN technique (Oyama et al 1989) used the relative magnitudes of the input line or phase voltages for generating the switching signals. Gupta et al (2010) proposed the use of rotating space vectors for synthesizing the required outputs of the matrix converter for the elimination of the common mode voltage in the matrix converter fed induction machines. A generalized technique for modeling, analysis and control of a matrix converter using the singular value decomposition method was proposed recently by Hojabri et al (2011), which leads to a unified modulation technique that achieves the full capability for a matrix converter. In general, many of the modulation methods, established for the matrix converter, are specific cases of this technique.

Since the most commonly used matrix converter modulation techniques now is the indirect space vector modulation technique, the same is widely used in the thesis and a brief review of it is presented next.
1.5.1 **Indirect Space Vector Modulation Technique**

The Space Vector Modulation (SVM) techniques are the extension of the theory of flux in multi-phase rotating machines (Bose 2004) to the field of static power converters. Space Vector PWM (SVPWM) techniques are well known for the Voltage Source Inverters (VSI). The idea of Direct Space Vector Modulation (DSVM) for the matrix converter is difficult to understand because of the unified representation of the current and the voltage space vectors. Therefore, it would be a good approach to decouple the space vectors of the matrix converter into independent current and voltage space vectors for understanding its control (Huber and Borojevic 1995). This leads to the decoupling of the matrix converter into a fictitious converter (input converter) and a fictitious inverter (output converter) connected back to back, as shown in Figure 1.7. Later, both the current control and the voltage control are unified for the actual matrix converter and this method is termed as the Indirect Space Vector (ISVM) PWM.

![Figure 1.7 Decoupled representation of the conventional matrix converter](image)

Figure 1.7 Decoupled representation of the conventional matrix converter
The ISVM method obtains the required output voltages through two analytically independent stages or transformations of the input voltages. This requires representing the switching function $T$, given by Equation (1.3), as the product of the rectifier switching function and the inverter switching function, as given in Equation (1.5). In the first stage, the three-phase AC voltage is converted to an average DC bus voltage of constant value. In the second stage, this constant average DC bus voltage is converted to a three-phase AC voltage of the required frequency and amplitude. Equations (1.6) and (1.7) give the equations for the output voltages, output currents, input voltages, and input currents with respect to the fictitious DC bus

$$T = T_I \times T_R$$

$$\begin{bmatrix} S_{Aa} & S_{Ab} & S_{Ac} \\ S_{Ba} & S_{Bb} & S_{Bc} \\ S_{Ca} & S_{Cb} & S_{Cc} \end{bmatrix} = \begin{bmatrix} S_{ap} & S_{an} \\ S_{bp} & S_{bn} \\ S_{cp} & S_{cn} \end{bmatrix} \times \begin{bmatrix} S_{Ap} & S_{Bp} & S_{Cp} \\ S_{An} & S_{Bn} & S_{Cn} \end{bmatrix}$$  \hspace{1cm} (1.5)

where,

$$T = \begin{bmatrix} S_{Aa} & S_{Ab} & S_{Ac} \\ S_{Ba} & S_{Bb} & S_{Bc} \\ S_{Ca} & S_{Cb} & S_{Cc} \end{bmatrix}, \quad T_I = \begin{bmatrix} S_{ap} & S_{an} \\ S_{bp} & S_{bn} \\ S_{cp} & S_{cn} \end{bmatrix} \quad \text{and} \quad T_R = \begin{bmatrix} S_{Ap} & S_{Bp} & S_{Cp} \\ S_{An} & S_{Bn} & S_{Cn} \end{bmatrix}$$

$$\begin{bmatrix} V_a \\ V_b \\ V_c \end{bmatrix} = \begin{bmatrix} S_{ap} & S_{an} \\ S_{bp} & S_{bn} \\ S_{cp} & S_{cn} \end{bmatrix} \times \begin{bmatrix} V_{DC+} \\ V_{DC-} \end{bmatrix}$$  \hspace{1cm} (1.6)

$$\begin{bmatrix} I_{DC+} \\ I_{DC-} \end{bmatrix} = \begin{bmatrix} S_{ap} & S_{bp} & S_{cp} \\ S_{an} & S_{bn} & S_{cn} \end{bmatrix} \times \begin{bmatrix} I_a \\ I_b \end{bmatrix}$$

$$\begin{bmatrix} I_A \\ I_B \\ I_C \end{bmatrix} = \begin{bmatrix} S_{Ap} & S_{An} \\ S_{Bp} & S_{Bn} \\ S_{Cp} & S_{Cn} \end{bmatrix} \times \begin{bmatrix} I_{DC+} \\ I_{DC-} \end{bmatrix}$$  \hspace{1cm} (1.7)

$$\begin{bmatrix} V_{DC+} \\ V_{DC-} \end{bmatrix} = \begin{bmatrix} S_{Ap} & S_{Bp} & S_{Cp} \\ S_{An} & S_{Bn} & S_{Cn} \end{bmatrix} \times \begin{bmatrix} V_A \\ V_B \\ V_C \end{bmatrix}$$
The restrictions on the switch states that were analyzed in section 1.3 can now be applied to the rectifier and the inverter sections independently. The switches of the inverter stage, associated with lines P and N in Figure 1.7, cannot be simultaneously closed but the switches of the converter stage, associated with lines P and N, can be closed. Equations (1.8) and (1.9) express these restrictions mathematically.

\[
d_{\text{PI}} + d_{\text{NI}} = 1 \quad (1.8)
\]

\[
d_{\text{PR}} + d_{\text{NR}} < 1 \quad (1.9)
\]

For maintaining the flow of the DC bus current constant on an average basis, the DC bus voltage on the fictitious DC bus is maintained constant. This indicates that the input stage is a constant current source. Similarly, for maintaining the voltage constant (for a particular modulation index) for any load, the output stage works as a constant voltage source.

It can be assumed that the power conversion happens through the fictitious DC-link. Because of the absence of any energy storage element in the matrix converter, the averaged value of this DC-link voltage \( V_{\text{DC}} \) can be found, as shown in Equation (1.10), on the basis that the input power flow and the DC power flow are equal at any instant.

\[
P_{\text{DC}} = P_{\text{IN}}
\]

\[
V_{\text{DC}} = \frac{3}{2} V_{\text{IN}} \frac{I_{\text{IN}}}{I_{\text{DC}}} \cos(\phi_{\text{in}})
\]

\[
V_{\text{DC}} = \frac{3}{2} V_{\text{IN}} m_c \cos(\phi_{\text{in}}) \quad (1.10)
\]

In a similar manner, Equation (1.11) gives the fictitious DC-link current \( I_{\text{DC}} \).

\[
P_{\text{DC}} = P_{\text{OUT}}
\]
\( I_{DC} = \frac{3}{2} I_{OUT} \frac{V_{OUT}}{V_{DC}} \cos(\varphi_{out}) \)

\( I_{DC} = \frac{3}{2} I_{OUT} m_c \cos(\varphi_{out}) \quad (1.11) \)

From Equation (1.10), it can be seen that the DC bus voltage is a function of the \( m_c \). To make available the maximum DC bus voltage for the inverter stage, \( m_c \) is always fixed at its maximum value.

Equations (1.12) and (1.13) express the inputs \( I_{IN}, V_{IN} \) and the outputs \( V_{OUT}, I_{OUT} \) as space vectors.

\( V_{IN} = \frac{2}{3} \left( V_A + V_B e^{j\frac{2\pi}{3}} + V_C e^{j\frac{4\pi}{3}} \right), \quad I_{IN} = \frac{2}{3} \left( I_A + I_B e^{j\frac{2\pi}{3}} + I_C e^{j\frac{4\pi}{3}} \right) \quad (1.12) \)

\( V_{OUT} = \frac{2}{3} \left( V_A + V_B e^{j\frac{2\pi}{3}} + V_C e^{j\frac{4\pi}{3}} \right), \quad I_{OUT} = \frac{2}{3} \left( I_A + I_B e^{j\frac{2\pi}{3}} + I_C e^{j\frac{4\pi}{3}} \right) \quad (1.13) \)

Tables 1.1 and 1.2 present the space vectors and the relevant switching states of the converter and the inverter stages.

**Table 1.1 Switching states and input currents for the converter stage**

| \( m \) | \( S_{Cm} \) | \( I_A \) | \( I_B \) | \( I_C \) | \( |I_{IN}| \) | \( \angle I_{IN} \) |
|---|---|---|---|---|---|---|
| 1 | [1 -1 0] | \( I_{DC} \) | - \( I_{DC} \) | 0 | \( \frac{2}{\sqrt{3}} I_{DC} \) | - \( \frac{\pi}{6} \) |
| 2 | [1 0 -1] | \( I_{DC} \) | 0 | - \( I_{DC} \) | \( \frac{2}{\sqrt{3}} I_{DC} \) | + \( \frac{\pi}{6} \) |
| 3 | [0 1 -1] | 0 | \( I_{DC} \) | - \( I_{DC} \) | \( \frac{2}{\sqrt{3}} I_{DC} \) | + \( \frac{\pi}{2} \) |
| 4 | [-1 1 0] | - \( I_{DC} \) | \( I_{DC} \) | 0 | \( \frac{2}{\sqrt{3}} I_{DC} \) | + \( \frac{5\pi}{6} \) |
| 5 | [-1 0 1] | - \( I_{DC} \) | 0 | \( I_{DC} \) | \( \frac{2}{\sqrt{3}} I_{DC} \) | - \( \frac{5\pi}{6} \) |
| 6 | [0 -1 1] | 0 | - \( I_{DC} \) | \( I_{DC} \) | \( \frac{2}{\sqrt{3}} I_{DC} \) | - \( \frac{\pi}{2} \) |
| 7 | [(1, -1) 0 0] | 0 | 0 | 0 | 0 | - |
| 8 | [0 (1, -1) 0] | 0 | 0 | 0 | 0 | - |
| 9 | [0 0 (1, -1)] | 0 | 0 | 0 | 0 | - |
Table 1.2 Switching states and output voltage for the inverter stage

| m | $S_{lm}$ | $V_a$  | $V_b$  | $V_c$  | $|V_{OUT}|$ | $\angle V_{OUT}$ |
|---|---------|--------|--------|--------|-----------|-----------------|
| 1 | [100]   | $\frac{2}{3} V_{DC}$ | $-\frac{1}{3} V_{DC}$ | $-\frac{1}{3} V_{DC}$ | $\frac{2}{3} V_{DC}$ | 0 |
| 2 | [110]   | $\frac{1}{3} V_{DC}$ | $\frac{1}{3} V_{DC}$ | $-\frac{2}{3} V_{DC}$ | $\frac{2}{3} V_{DC}$ | $\frac{\pi}{3}$ |
| 3 | [010]   | $-\frac{1}{3} V_{DC}$ | $\frac{2}{3} V_{DC}$ | $-\frac{1}{3} V_{DC}$ | $\frac{2}{3} V_{DC}$ | $\frac{2\pi}{3}$ |
| 4 | [011]   | $-\frac{2}{3} V_{DC}$ | $\frac{1}{3} V_{DC}$ | $\frac{1}{3} V_{DC}$ | $\frac{2}{3} V_{DC}$ | $\pi$ |
| 5 | [001]   | $-\frac{1}{3} V_{DC}$ | $-\frac{1}{3} V_{DC}$ | $\frac{2}{3} V_{DC}$ | $\frac{2}{3} V_{DC}$ | $-\frac{2\pi}{3}$ |
| 6 | [101]   | $\frac{1}{3} V_{DC}$ | $-\frac{2}{3} V_{DC}$ | $\frac{1}{3} V_{DC}$ | $\frac{2}{3} V_{DC}$ | $-\frac{\pi}{3}$ |
| 7 | [000]   | 0       | 0       | 0       | 0       | 0 |
| 8 | [111]   | 0       | 0       | 0       | 0       | - |

Note: 1 in the $S_{cm}$ represents the upper switch in an arm being ON, -1 in the $S_{cm}$ represents the lower switch in an arm being ON, (1,-1) in the $S_{cm}$ represents both the switches in an arm being ON and 0 in the $S_{cm}$ represents both the switches in an arm being OFF.

An arbitrary voltage $V_{OUT}$ in Figure 1.8(a) can be synthesized as a vector sum of any two adjacent active vectors ($V_1$ to $V_6$) and one of the zero vectors ($V_0$, $V_7$). Similarly, the input current $I_{IN}$ in Figure 1.8(b) can be synthesized as a vector sum of any two adjacent active vectors ($I_1$ to $I_6$) and one of the zero vectors ($I_{0a}$, $I_{0b}$, $I_{0c}$).
Figure 1.8 (a) Inverter voltage hexagon and (b) rectifier current hexagon

In general, any arbitrary vector $A_v$, shown in Figure 1.9, can be expressed by the vector-time product sum of the adjacent active vectors and the zero vector as $A_v = d_x \times v_x + d_y \times v_y + d_z \times v_z$.

Figure 1.9 Representation of the vector-time product

The durations of the active vectors determine the direction of the arbitrary vector while the zero vector is used to adjust the amplitude of the arbitrary vector. Equations (1.14) to (1.16) give the duty cycle of the active vectors and the zero vector

$$d_x = \frac{T_x}{T_s} = m_i \sin \left( \frac{\theta}{3} - \theta_{AV} \right)$$  (1.14)
\[ d_y = \frac{T_y}{T_s} = m_i \sin (\theta_{AV}) \] 

\[ d_z = \frac{T_z}{T_s} = (1 - d_x - d_y) \] 

where, \( \theta_{AV} \) represents the angle of the reference arbitrary vector within that sector, \( m_i \) is the modulation index defining the magnitude of the arbitrary vector \( A_V \).

Applying this space vector concept on the voltage vectors of the inverter stage and the current vectors of the converter stage, Equations (1.17) and (1.18) give the expressions for the duty cycle of the vectors to synthesize the required output voltage and the required input current.

\[ d_{\nu \nu} = \frac{T_{\nu \nu}}{T_s} = m_\nu \sin \left( \frac{\pi}{3} - \theta_\nu \right) \]

\[ d_{\nu \beta} = \frac{T_{\nu \beta}}{T_s} = m_\nu \sin (\theta_\nu) \]

\[ d_{\nu 0} = \frac{T_{\nu 0}}{T_s} = (1 - d_{\nu \nu} - d_{\nu \beta}) \] 

\[ d_{\gamma \nu} = \frac{T_{\gamma \nu}}{T_s} = m_\gamma \sin \left( \frac{\pi}{3} - \theta_\gamma \right) \]

\[ d_{\gamma \beta} = \frac{T_{\gamma \beta}}{T_s} = m_\gamma \sin (\theta_\gamma) \]

\[ d_{\gamma 0} = \frac{T_{\gamma 0}}{T_s} = (1 - d_{\gamma \nu} - d_{\gamma \beta}) \]

\[ d_{\xi 0} = \frac{T_{\xi 0}}{T_s} = (1 - d_{\xi \nu} - d_{\xi 0}) \] 

These duty cycles correspond to the ON and OFF times of the switches of the fictitious inverter and converter circuits. Therefore, the two independent space vector modulations should be merged into one modulation for the nine bi-directional switched matrix converter.
The combination of the six active states of the rectifier with the six active states of the inverter produces a set of 36 active states. These states can be grouped into 18 pairs of equivalent states. The states in each pair are equivalent because both connect the same input to the same output. For example, the combination of the rectifier state \( m=1 \) and inverter state \( m=6 \) given in Tables 1.1 and 1.2 is equivalent to the combination of the rectifier state \( m=4 \) and inverter state \( m=3 \). Both result in the connection of the output lines a and c to the input line A and the output line b to the input line B. Both these combinations are the same state (ABA) in matrix converter, as shown in Figure 1.10.

Figure 1.10 Switching equivalence in the conventional matrix converter

Equations (1.19) to (1.23) give the duty cycles for the matrix converter, which are derived from the product of the inverter duty cycles and the rectifier duty cycles (Cha 2004).

\[
d_{V/I/c} = d_{V/I} \times d_{c/I} = m_v \sin\left(\frac{\pi}{3} - \theta_v\right) \times m_c \sin\left(\frac{\pi}{3} - \theta_c\right) = \frac{T_{V/I/c}}{T_s}
\]  

(1.19)
\[ d_{\alpha\beta} = d_{\alpha} \times d_{\beta} = m_{\alpha} \sin \left( \frac{\pi}{3} - \theta_{\alpha} \right) \times m_{\beta} \sin \left( \frac{\pi}{3} - \theta_{\beta} \right) = \frac{T_{\alpha\beta}}{T_{s}} \quad (1.20) \]

\[ d_{\beta\alpha} = d_{\beta} \times d_{\alpha} = m_{\beta} \sin \left( \frac{\pi}{3} - \theta_{\beta} \right) \times m_{\alpha} \sin \left( \frac{\pi}{3} - \theta_{\alpha} \right) = \frac{T_{\beta\alpha}}{T_{s}} \quad (1.21) \]

\[ d_{\alpha\beta} = d_{\alpha} \times d_{\beta} = m_{\alpha} \sin \left( \frac{\pi}{3} - \theta_{\alpha} \right) \times m_{\beta} \sin \left( \frac{\pi}{3} - \theta_{\beta} \right) = \frac{T_{\alpha\beta}}{T_{s}} \quad (1.22) \]

\[ d_{0} = 1 - d_{\alpha\alpha} - d_{\beta\beta} = \frac{T_{0}}{T_{s}} \quad (1.23) \]

### 1.5.2 Minimum Error Switching Strategy (MESS)

In the Minimum Error Switching Strategy (MESS), the indirect-space vectors of the matrix converter are used for the input current control and the output voltage control. The switching vectors are selected based on the integrated minimum computed square of the voltage error and the integrated minimum computed square of the current error over every sampling period. The hysteresis current controller has a constant error band and switches from one vector to another vector at different time intervals as and when the error exceeds the band limits, as shown in Figure 1.11(a). The MESS technique has a constant time band and switches from one vector to another vector at constant time intervals, as shown in Figure 1.11 (b).

![Figure 1.11 Control principle (a) hysteresis and (b) MESS](image-url)
The error in the MESS technique is limited by the envelope of the minimum error among the available space vector errors at the end of each sampling time, as shown in Figure 1.12.

![Figure 1.12 Envelope of the minimum square error](image)

**Figure 1.12 Envelope of the minimum square error**

The method minimizes the error between the actual output and the expected output in a switching period by selecting the switching vector that provides the minimum square error. In the next switching instant, the previously calculated minimum square error is added to the present square errors of the available space vectors, which becomes the input to the controller that finds the appropriate minimum error switching states. The proposed switching strategy decouples the CMC into an Output Converter (OC) and a Input Converter (IC), as explained in section 2. It selects the appropriate switching that minimizes the square of the output voltage error for the OC and the square of the input current error for the IC.

It is assumed that the constant average DC voltage and the constant average DC current are available at the Fictitious DC Bus (FDCB). During every switching period, the error quantities $V_{e_{nj}}$ for the OC and $I_{e_{nj}}$ for the IC
are calculated using Equations (1.24) to (1.27), and Tables 1.1 and 1.2. Since the input current magnitude is dependent only on the load, the modulation index \( m_c \) of the IC always operates at unity. Hence, the \( I_{DC} \) of the FDCB link is assumed to be constant with magnitude of one and used for error calculations of the current for the IC. \( V_{DC} \) is determined using the selected switching pattern of the IC and the measured input voltage, as given by Equation (1.32), which is used for the voltage error calculations of the OC. The switching states \( S_{ion} \) and \( S_{con} \), given in Equations (1.28) and (1.29), are selected that correspond to the smallest of the sum of the square of the three-phase errors and the propagated error. Finally, both the OC and the IC switching signals are processed through a digital logic circuit for generating the CMC switching signals

\[
V_{mj}(t) = S_{im} \times V_{DC_i}(t), \quad m \in \{1 \rightarrow 8\}, \quad i \in \{+,-\} \quad \& \quad j \in \{a,b,c\} \quad (1.24)
\]

\[
V_{mj}(t) = \left( V_{rj}(t) - V_{mj}(t) \right) + V_{pj}(t) + V_{mj}(t) \quad (1.25)
\]

\[
I_{mj}(t) = S_{cm}^T \times I_{DC_i}(t), \quad m \in \{1 \rightarrow 9\}, \quad i \in \{+,-\} \quad \& \quad j \in \{A,B,C\} \quad (1.26)
\]

\[
I_{mj}(t) = \left( I_{rj}(t) - I_{mj}(t) \right) + I_{pj}(t) \quad (1.27)
\]

\[
S_{on}(t) = \min_{m \in 1 \rightarrow 8} \sum_{i=a,b,c} |V_{mj}(t)|^2 \quad (1.28)
\]

\[
S_{on}(t) = \min_{m \in 1 \rightarrow 9} \sum_{i=A,B,C} |I_{mj}(t)|^2 \quad (1.29)
\]

\[
V_{pj}(t) = V_{mj}(t-1) \in S_{on}(t-1) \quad (1.30)
\]

\[
I_{pj}(t) = I_{mj}(t-1) \in S_{on}(t-1) \quad (1.31)
\]
\begin{equation}
V_{DC} = \begin{bmatrix}
S_{Ap} & S_{Bp} & S_{Cp} \\
-S_{An} & -S_{Bn} & -S_{Cn}
\end{bmatrix} \times \begin{bmatrix}
V_A \\
V_B \\
V_C
\end{bmatrix}
\end{equation}

where, $V_{mj}$ and $I_{mj}$ are respectively the output voltage and the input current corresponding to the switching matrixes $S_{Im}$ and $S_{Cm}$. $V_{rj}$ and $I_{rj}$ denote respectively the reference output voltage and the reference input current; $V_{e_{mj}}$ and $I_{e_{mj}}$ are respectively the output voltage error and the input current error corresponding to switching states $S_{Im}$ and $S_{Cm}$; $V_{e_{pj}}$ and $I_{e_{pj}}$ are respectively the output voltage error and the input current error due to the previous switching state and $S_{Ion}$ and $S_{Con}$ are respectively the switching matrices used in the present switching states for the OC and the IC.

1.5.2.1 Improved current control with the MESS

Considerably high harmonic content of the input current of the CMC is experienced with the MESS technique for lower modulation indices. Hence, a modified MESS technique is proposed in this section to improve its performance for low modulation indices. It is observed that the zero vectors of the IC do not influence the input current spectrum but zero vectors of the OC influence the input current spectrum by a considerable factor. Hence, the error of the IC is recalculated as explained in this section. The assumption that a constant average $I_{DC}$ flows in the FDCB holds good only for the active vectors of the OC. However, for the zero vectors of the OC, the above said assumption does not hold good. This is because, with the application of the zero vectors of the OC, the $I_{DC}$ becomes zero due to the isolation of the source from the load. There is a need to recalculate the current error, which could be used for calculation in the next sampling time. Hence, it is proposed, for the improvement of MESS technique, to ignore the calculated current error, $I_{e_{mj}}(t-1)$ and use $I_{e_{mj}}(t-2)$, when a zero vector is applied on the OC, as given
by Equation (1.33). Figure 1.13 shows the block diagram representation of the improved MESS technique.

\[
I_{e pj}(t) = I_{e mj}(t-1) \in S_{Con}(t-1)
\]
(when the OC uses an active vector) \hfill (1.33)

\[
I_{e pj}(t) = I_{e mj}(t-2) \in S_{Con}(t-2)
\]
(when the OC uses a zero vector)

Figure 1.13 Parallel execution sequence of the MESS algorithm

Figure 1.14 shows that for lower modulation indices, in the improved MESS technique, the input current Total Harmonic Distortion (THD) reduces by 59% when compared to the unmodified MESS algorithm.
Figure 1.14 Current harmonics (a) MESS technique and (b) improved MESS technique

The proposed technique has the inherent ability of mitigating the effects of the unbalance and the harmonics present at the input. Since the method works on the error propagation and compensation scheme, the effects of the unbalance and the harmonics are completely eliminated or reduced depending on the magnitudes of the unbalance, the harmonics and the present modulation index.

To evaluate the performance of the proposed techniques, simulation with an R–L load was performed in the MATLAB-Simulink environment for both the ISVM and the MESS techniques. The load parameters are $R_L = 20 \, \Omega$. 
$L_L = 21 \text{ mH}$ and the converter parameters are input voltage of 100 V, modulation index of 0.75, input frequency of 50 Hz, output frequency of 25 Hz and switching frequency of 7 kHz.

**Figure 1.15** Simulation results of the output phase voltage (a) ISVM technique and (b) MESS technique

Figure 1.15 shows the MESS switching states that are compared with the ISVM switching states. It can be observed in Figure 1.16 that in the ISVM method multiple switchings occur within the sampling frequency when compared to the MESS technique. This results in an increased switching loss, for the same switching frequency.
Figure 1.16 Total losses in the CMC with the ISVM and the MESS techniques

The current harmonics of the MESS are comparatively high when compared to the ISVM technique. At higher switching frequencies, the MESS technique has superior performance as compared to the ISVM, because the ISVM introduces multiple pulses within the switching interval, which increases the actual switching frequency of the device due to which the device may fail to respond. These introduce more harmonics at the output current.

An unbalance of 20% in the phase B was introduced at 0.06 s and the converter operates at the modulation index of 0.75. The MESS technique mitigates the unbalance automatically without any additional computation at the output, as seen in Figure 1.17(d).

The experiment was conducted with a balanced input voltage of 100 V, switching frequency of 7 kHz, \( R_L = 20 \, \Omega \), \( L_L = 21 \text{mH} \) and modulation index of 0.75 on a 3 kVA matrix converter prototype. The CMC was used for converting the 50 Hz input frequency to the 25 Hz output frequency using the MESS technique, as shown in Figures 1.18 (a) to 1.18 (e). The hardware results verify the effectiveness of the proposed MESS method for the CMC.
Figure 1.17 (Continued)
Figure 1.17 Simulation results (a) input phase voltage, (b) output phase voltage, (c) output line voltage, (d) output current and (e) input current
Figure 1.18 (Continued)
CERTAIN ISSUES IN THE MATRIX CONVERTERS

1.6 Over Modulation Operation of the Matrix Converter

The over modulation operation has been described as a nonlinear operation (Holtz et al 1993) since the output waveform of the converter does not follow the original sinusoidal reference waveform in the regions of higher magnitudes. The over-modulation in the DC-link converters has been widely

Figure 1.18 Hardware results (a) input phase voltage, (b) output phase voltage, (c) output line voltage, (d) output current and (e) input current
described in the literature (Bolognani and Zigliotto 1996) but only a few papers describe the detailed effects of over modulation operation of the matrix converter (Thuta 2007). The paper discusses four ways of operating the matrix converter under over modulation (i) output side over modulation, (ii) input side over modulation with power factor control, (iii) input side over modulation without power factor control and (iv) simultaneous output and input side over modulation. Using the over modulation technique, the theoretical voltage limit of the converter can be increased to 105 % of the input voltage. It has been proved in Mahlein et al (1999) that some lower order harmonics are generated at the output voltage and the input current by the over modulation operation (Wiechmann et al 1997). Over modulation operation of the matrix converter might cause the resonance of the line side filter. This might damage the converter if not controlled properly. Thus, it was concluded in Wiechmann et al (2002) that it is not advisable to operate the matrix converter under over modulation for a long time but for a short period, if demanded, for the ride-through operation.

### 1.6.2 Ride-Through Capability of the Matrix Converter

One of the desirable characteristics of the modern drives is its ride-through capability. This is a common solution for the drives during power loss. During ride-through, to magnetize the motor windings and to feed the control circuits, the drive utilizes energy from the load inertia. This is achieved by maintaining a constant voltage in the DC-link capacitor in the AC-DC-AC converters (Narayanan and Tanganathan 2002, Kim and Sul 2001 and Jounne et al 1999). However, the matrix converters are an array of controlled bidirectional switches without the DC-link capacitor and these are highly susceptible to voltage disturbances such as voltage sags, voltage swells and momentary power interruption. A new ride-through strategy for the matrix converter developed by Klumpner et al (2001) uses the zero vectors of
the matrix converter and the clamp circuit to ride-through small interruptions. In Wiechmann et al (2002), an alternative strategy was presented that enables the converter to ride-through the voltage sags and enforces constant V/f operation with the minimum reduction in the speed. Later, a new approach was presented in Cha (2004) that modified the topology of the matrix converter with three additional unidirectional switches and a ride-through capacitor.

1.6.3 Unbalanced Operation and Control of the Matrix Converter

The effect of the unbalance on the converter performance is a vital aspect in determining the overall performance of the variable speed drive, which is fed by a converter. The matrix converter, being a direct frequency conversion system, the unbalance at the utility side is immediately reflected on the load side and generates unwanted lower order input/output harmonic currents (Enjeti and Wang 1990) that may resonate with the input filter causing damage to the converter, if uncontrolled. Therefore, research has been directed to investigate and compensate for these effects of input voltage disturbance. In Nielsen et al (1996), balanced and sinusoidal output voltages were produced even when the input voltages were unbalanced. In Casadei et al (1998) and Blaabjerg et al (2002), the input current harmonic content and the limits of the voltage transfer ratio of matrix converter under unbalanced conditions were determined analytically for different operating conditions. In Zhang et al (2001) and Sunter et al (2002), the line side voltage conditions with high order voltage harmonic components are analyzed. However, it was concluded in all these techniques that the input current harmonics could not be reduced when compensated for the output harmonics under abnormal conditions of the input voltage.
1.6.4 Common Mode Effects of the Matrix Converter

The high frequency common mode voltage generated in the power converters are reported to cause potential damage to the shaft and the bearings of the electric motors. The reduction of the common mode voltage in matrix converters using proper switching sequence has been reported in Nguyen and Lee (2012), Cha and Enjeti (2003). Recently, Gupta et al (2010) presented the elimination of the common mode voltage in the matrix converter fed open-ended induction machine.

1.7 REVIEW OF TOPOLOGY CHANGES TO THE CONVENTIONAL MATRIX CONVERTER

1.7.1 Reduced Number of Switches

New topologies that are different from the conventional matrix converters, named as indirect matrix converter topologies, consisting of a rectifier/ inverter circuit without a DC-link were proposed in Ziogas et al (1986), Kim and Sul (1993), Wiechmann et al (1985) along with their PWM control and commutation procedures. However, these papers showed a low quality input current waveforms. For improving the quality of the input current, a detailed PWM control method with a synchronized switching strategy for both the line and the load side switches was proposed in Wei and Lipo (2001), Kolar et al (2002). Zero current commutation methods of the line side switches were also discussed in detail in Holtz and Boelkens (1989) for the indirect matrix converter topologies to reduce the complexity involved in the four-step commutation of bidirectional switches in the matrix converters. Moreover, in Kolar et al (2002) and Wei et al (2002), the possibility of reducing the total number of semiconductor switches required for the indirect matrix converter topology was presented. Even though constructed with reduced number of switching devices, these converters were still able to
provide unity input displacement factor, sinusoidal supply currents and load voltages that were identical to the conventional matrix converters. These topologies are referred as the “sparse matrix converters”. These sparse matrix converters were classified as (i) Simple Sparse Matrix Converter (SSMC) with 15 switches, (ii) Very Sparse Matrix Converter (VSMC) with 12 switches and (iii) Ultra Sparse Matrix Converter (USMC) with 9 switches (Meng Yeong Lee 2009). The VSMC and the USMC were designed based on the fact that the DC link current only flows in one direction. This constraint makes the VSMC and the USMC not applicable for regenerative operation.

1.7.2 Multilevel Output Voltage Operation

The evolution of the multilevel inverters (Mekhilef and Kadir 2011, Aneesh et al 2009 and Celanovic 2000) brought out the importance of the reduced voltage stresses on the power devices by using many number of lower rating power supplies and power devices. Using the same idea in matrix converters, a new family of converters called the Multilevel Matrix Converters (MLMC) evolved with different concepts; i) Replacing each bidirectional switch in CMC with n cells, each cell consisting of a capacitor connected to the centre of the H-Bridge (Erickson et al 2006). This topology generates multilevel output but at the cost of a more complicated circuit configuration and modulation strategy. ii) Modifying the topology of the IMC with additional switches which makes available two different voltage levels at the output i.e., the phase and the line voltages (Meng Yeong Lee et al 2010). This multilevel matrix converter topology is a hybrid combination of a simplified three-level neutral-point clamped voltage source inverter (Rojas et al 1993) and an indirect matrix converter topology (Ziogas et al 1986). The indirect three-level sparse matrix converter has a simpler circuit configuration than the multilevel matrix converter topologies proposed in
Meng Yeong Lee et al (2010), but was still able to generate multilevel output waveforms. A detailed analysis of this topology was presented in Meng Yeong Lee (2009).

1.7.3 Polyphase Matrix Converter Operation

The use of 3×6 matrix converters for a six phase induction machine drive system was presented in Wang et al (2011), Ghalem and Azeddine (2010). In addition to the changes in the topologies of the matrix converter, the use of the polyphase matrix converter for innovative active generator was demonstrated in Beguin (2012). It explained along with a prototype, the commutation, the modulation and the control principles of a 27×3 matrix converter.

1.8 REVIEW OF THE MATRIX CONVERTER APPLICATIONS

Neft and Schauder (1992) experimentally confirmed that a matrix converter with only nine switches can be effectively used in the vector control of an induction motor with high quality input and output currents. The compactness of the matrix converter suggested the possibility to integrate the converter and the motor in a single unit, in order to reduce the costs and to increase the overall efficiency (Klumpner et al 2002, Itoh et al 2005). Casadai et al (2001) used the matrix converter in the direct torque control (DTC) of induction machines. Podlesak et al (2005) presented the field oriented-control of the matrix converter fed induction machine.

Matrix converters find their application in the field of wind power generation in full power converter topologies and partial converter topologies for the doubly fed induction generators control (Zhang et al 1997, Lie Xu and Cartwright 2006). Research on modeling and analysis of the matrix converter
based wind energy systems was carried out in Barakathi (2008). Control of the reactive power supplied by a wind energy conversion system (WECS) based on the induction generator fed by a matrix converter was presented in Cardenas et al (2009). An increasing number of papers (Imayavaramban and Wheeler 2007, Wheeler et al 2003 and Lillo 2006) investigating the advantages/ limitations of the use of matrix converters in aircrafts are also being reported.

Today, research in matrix converter are in advanced technological and application issues such as reliable implementation of the modified topologies, operation under abnormal conditions and the design of matrix converter for control of machines with more number of phases. However, industrial applications of the converter are still limited because of some practical issues such as difficulty in implementing complex switching methods, common mode voltage effects, high susceptibility to input power disturbances and low voltage transfer ratio.

The desire to use all the advantages offered by the matrix converter has inspired me to work in this area for my PhD research. This research work attempts to investigate the existing PWM techniques and suggest modifications for improving the performance of the matrix converter. This thesis focuses on devising easier methods to implement complex switching strategies, study and mitigation of effects of the unbalance, topological changes to increase the performance indices, proper use of the modulation technique to eliminate the common mode voltage and a new direct torque control procedure for the control of induction motor fed by the modified matrix converter topology.
1.9 OBJECTIVES

The main objectives of the research study are given below.

(i) To extract a variable amplitude and a variable frequency output voltage with a relatively simple firing scheme called the Decoupled Indirect Duty Cycle (DIDC) technique suitable for a direct three-phase-to-three-phase matrix converter that strives to reduce the THD and retain the target fundamental component.

(ii) To suggest the idea of using a six-phase matrix converter in a three-phase induction motor drive system for the elimination of the common mode voltage and design a switching methodology to offer a higher modulation index and a lower THD for the output voltage.

(iii) To analyze the performance of the matrix converter for the effects of the unbalance and the harmonics at the input and to mitigate the effects of the unbalance using a refined firing algorithm termed as the Harmonic Tracking Algorithm (HTA) in a way that facilitates the reduction in THD, ensuring the maximum possible fundamental component.

(iv) To modify the matrix converter topology, termed as the Direct Three-Level Matrix Converter (DTMC) to reduce the output voltage THD. To demonstrate the ISVM method for the DTMC and its required modification to achieve reduced voltage THD and the highest fundamental value of the output voltage.

(v) To develop the Direct Torque Control (DTC) procedure for DTMC fed induction motor drive and to reduce the torque
ripple in the AC-to-AC converter fed induction motor drive with input power-factor control.

1.10 THESIS ORGANIZATION

The thesis contains seven chapters summarized as follows:

**Chapter 1** reviews the need for investigating the matrix converter and developing its control methods for applying it to the AC drive system. The role of the matrix converter as an all-silicon solution in industrial drives is explained and elucidated. This chapter enumerates the necessity to bring about topological changes in the matrix converter to ensure better quality of the output waveforms. In addition, the need for developing simple PWM methods for the matrix converter for industrial use is emphasized. The chapter also proposes a new modulation technique, suitable for high switching frequency, which uses the minimum error switching vector. At high switching frequencies, the power devices do not effectively respond to pulses smaller than one-tenth of the sampling time, as in the SVPWM. This requires that the width of the switching vector to be equal to the sampling time. In the proposed technique, constant pulse width is offered to each vector and the vector for the next switching period is selected based on the minimum error switching vector. This chapter also includes the review of the literature, research objectives, and the organization of the thesis.

**Chapter 2** develops a new PWM strategy called the Decoupled Indirect Duty Cycle (DIDC) PWM, with a focus on providing a carrier-based technique for the operation of the matrix converter with unity displacement factor. The main idea is to guarantee a simple and computation-less technique, as compared to the other techniques, that modulates the matrix converter. The matrix converter is decoupled into a fictitious converter-inverter pair and its modeling is presented. The technique extracts the duty
cycles of the CMC from the available reference signals (without the need of processors and memory) and simultaneously executes it with a simple digital logic circuit making it suitable for online (computation-less) implementation. Since the duty cycles of the fictitious converter inverter pair are combined using a digital circuit to achieve the CMC duty cycle, the input current harmonics increase. This increase in the input current harmonics is due to the non-coordinated inverter and converter zero vectors. To improve the performance of the DIDC technique, the carrier frequency adjustment method is proposed to reduce the THD in the input current. The chapter includes simulation and experimental validation to highlight the merits of the approach.

Chapter 3 aims to design a Rotating Space Vector Modulation (RSVM) PWM technique that uses the zero common mode voltage vectors (ZCMVV) to eliminate the common-mode voltage in the matrix converter. The procedure for using the ZCMVV for input current control is also established. The proposed RSVM technique eliminates Common Mode Voltage (CMV) with a low voltage transfer ratio of 0.5. A modified CMC topology, namely the Phase Shifted Dual Source Matrix Converter (PSDSMC), is proposed to increase the voltage transfer ratio to 0.866 with the modified RSVM technique. The performance of this scheme is evaluated through simulation in the MATLAB-Simulink environment. The approach involves deriving the mathematical relations for the proposed techniques. In addition, the chapter also brings out the conditions under which the approach fails to eliminate the common mode voltage. The simulation results portray the usefulness and limitations of the scheme.

In Chapter 4, the mathematical analysis of the matrix converter under unbalanced conditions is carried out and a technique to mitigate the effects of the unbalance at the output of the matrix converter is proposed. A
simple dynamic ISVM approach for the matrix converter operation for the unbalanced and the non-sinusoidal input voltage conditions are presented. Analyses of the effects of the unbalance on the FDCB of the CMC operated under ISVM PWM method is carried out. An unbalanced control method for the CMC is developed which uses the line side switching functions (converter) to track the oscillations of the average fictitious DC bus voltage and generate a dynamic modulation index for the load side (inverter). This approach is based on the simple compensation of the output voltage modulation vector with respect to the oscillating fictitious DC bus vector. The simulated results presented bring out the effectiveness of the proposed technique.

In Chapter 5, a new DTMC topology, which requires three bidirectional switches of lower ratings (rated for the phase voltage) in addition to the CMC topology, is proposed. The structure is a 4×3 matrix converter that facilitates the increase in the output voltage levels by making the input filter neutral point available to the load terminals. The DTMC topology with the modified ISVM technique reduces the THD at the output. The proposed DTMC ISVM technique uses the idea of multilevel inverter SVM technique along with the proposed neutral current balancing strategy for generating the firing pulses. The switching loss model for the DTMC is developed and the performance of the DTMC is compared with that of the CMC. The proposed DTMC ISVM technique is evaluated in simulation and validated with a hardware prototype.

Chapter 6 develops the DTC control method for the DTMC, which uses the input phase voltage vectors (short vectors) and the input line voltage vectors (long vectors). In the proposed algorithm, the large vectors are applied during torque transition states whereas the short vectors are utilized for the steady state conditions, which results in the reduction of the torque ripples.
However, the short vectors cause a serious problem of fluctuations in voltage at the input filter capacitance. Because of this problem, we obtain an output voltage from the DTMC that is asymmetric and having a non-zero average value. In this chapter, a solution to minimize this fluctuation is presented, which uses an additional voltage hysteresis band for reducing the voltage deviation at the neutral point due to the application of the short vectors. The performance of the DTC scheme for the DTMC is investigated through MATLAB-Simulink based simulation over a range of speed and torque.

Chapter 7 brings out the conclusions drawn from this research. The scope for future research in this area and the limitations of the work are also presented. The salient features of the work and the major contributions are summarized.