

Advanced Control Techniques for a Direct Matrix Converter

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*If we knew what we were doing it would not be called research,
would it?*

- Albert Einstein

Abstract

Model Predictive Control (MPC) has gained popularity in the recent years for control of power converters. One of the drawbacks of this control method is the variable switching frequency operation. By obtaining a fixed switching frequency operation while maintaining the advantages of MPC, the performance of these controllers can be enhanced. This thesis investigates the implementation of two novel control methods to control a direct matrix converter. The proposed control methods combine the features of classical MPC and the Space Vector Modulation technique to develop a hybrid predictive controller with an inbuilt modulation scheme. The resulting methods are known as Modulated Model Predictive Control (M2PC) and Direct Predictive Current-error Vector Control (DPCVC). The M2PC maintains all the characteristics of MPC (such as fast transient response, multi-objective control using only a single feedback loop, easy inclusion of non-linearities and constraints of the system, the flexibility to include other system requirements within the controller) adding the advantages of operating at a fixed switching frequency and improving the quality of the controlled waveforms. Simulation and experimental results employing the control method to a direct matrix converter feeding an RL load and an induction motor drive are presented. The DPCVC results in fixed switching frequency operation and improved current quality compared to M2PC, owing to the advanced modulation technique it utilises to generate the desired voltage vectors. It also inherits features such as fast transient response and multi-objective control without cascaded loops from the MPC. This method is implemented to control the load currents of a direct matrix converter feeding an RL load and an induction machine and the simulation and experimental results depicting the same are presented in this work. Finally a comparison of the different control methods for instance PI control, MPC, M2PC and DPCVC when used to control the load currents of a direct matrix converter are analysed and included in this thesis.

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List of Terms

V_o	- Output converter voltage
T_s	- Sampling time
I_s	- Supply Current
V_s	- Supply Voltage
V_l	- Inductor Voltage
L	- Load inductance
R	- Load resistance
C_f	- Input filter capacitance per phase
L_f	- Input filter inductance per phase
R_d	- Damping resistance of input filter
d	- Duty cycle for each bidirectional switch
q	- Voltage transfer ratio of the converter
λ	- Rotor flux angle
I_s^*	- Current reference
S_{ij}	- Switching function of ij_{th} switch
$V_{s\alpha}$	- α axis component of supply voltage
$V_{s\beta}$	- β axis component of supply voltage
V_{sd}	- d-axis component of stator voltage
V_{sq}	- q-axis component of stator voltage
I_{sd}	- d-axis component of stator current
I_{sq}	- q-axis component of stator current
V_{cd}	- d-axis component of converter voltage
V_{cq}	- q-axis component of converter voltage
τ_r	- Rotor time constant

τ_s - Stator time constant

R_s - Stator resistance

R_r - Rotor resistance

L_s - Stator inductance

L_r - Rotor inductance

L_{ls} - Stator leakage inductance

L_{lr} - Rotor leakage inductance

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Chapter 1

Introduction

1.1 Motivation

Energy efficiency is crucial now more than ever before due to the alarming rate of depletion of conventional energy resources. The energy from fossil fuels is not sustainable and hence there is an immediate need to shift to renewable energy systems and electrification of aircrafts and vehicles. A number of potential alternative, more sustainable sources are now being used for instance, wind, solar, tidal and hydro power. However most of these options produce less energy per unit area compared to their non renewable counter parts due to their low energy density and high power loss during power conversion and transmission. An inevitable part of any electrical system is the power conversion stage and it is crucial that this stage is highly efficient.

Nowadays, power converters using advanced power semiconductor switches are used for various applications for instance, power supplies, motor drives and renewable energy systems. Electrical drives with advanced power converter topologies are an integral part of modern day transportation systems where tremendous effort is being made to develop electric vehicles. For example *Tesla Motors*, *Nissan*, *BMW* and *Volkswagen* have made outstanding progress in electric vehicle technology by in-

corporating existing converter topologies with high end design and manufacturing. This proves that with advances in power electronics including packaging, reliability of devices and reduction in the volume of converters, the possibilities are limitless as new technologies are introduced. This highlights the need for advanced control and modulation techniques for power converters for drive applications and also to study new/non-conventional converter topologies. New converter technologies like modular multi-level converters and matrix converters are becoming popular nowadays due to the particular advantages that they offer and research is being carried out in academia and industry to improve these topologies. The objective of this work is to develop an advanced control strategy for direct matrix converters.

1.2 Literature Review - Matrix Converter

The concept of matrix converters were first introduced in literature by Gyugyi and Pelly in 1976 [1]. Matrix converters are capable of generating sinusoidal input currents and output voltages. A conventional matrix converter is capable of AC to AC conversion without the a DC link capacitor and are called direct matrix converters. The frequency of the output of a matrix converter can be higher or lower than the supply side which makes it suitable for a wide range of drive applications including aerospace [2].

Direct matrix converters are AC-AC power converters with nine bidirectional voltage-blocking current-conducting switches capable of single stage four-quadrant power conversion. The attractive feature of a matrix converter is the absence of any large energy storage devices which makes the converter smaller and as reliable compared to any other multilevel converter [3]. Figure 1.1 shows a three phase direct matrix converter(MC). In a direct MC, the switches are arranged in the form of a matrix such that at any time; any input phase can be connected to any output phase. The matrix converter is often known as an all silicon converter as there are no bulky and heavy energy storage devices [4]. The only passive components required are a small LC filter

on the input to the converter to attenuate the switching frequency components.

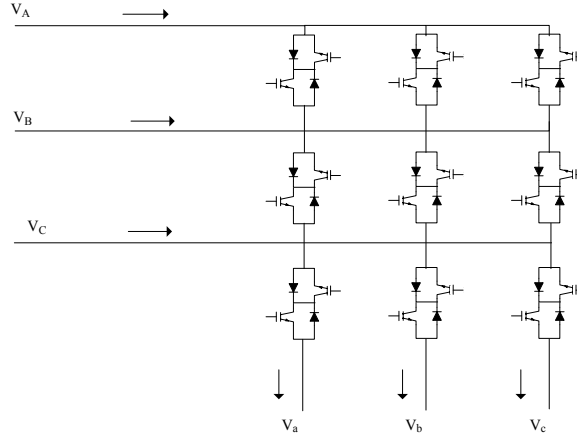


Figure 1.1: Three phase direct matrix converter

A variation on the conventional matrix converter was proposed in 2001 by Wei and Lipo [5] which is more similar to the traditional AC/DC/AC converter systems. These converters also known as indirect matrix converters consists of three phase current source converter which is used as a controlled bridge rectifier and a three phase voltage source converter connected to a common virtual DC bus without any energy storage element on the DC side [6, 7]. Six bidirectional switches form the rectifier stage and the inverter stage consists of six unidirectional switches. In this converter, the rectification and inversion stages have independent modulation strategies [8]. The rectifier stage generates sinusoidal input currents and a virtual DC-link voltage and the desired output voltage is generated by the inverter by using traditional modulation methods. Figure 1.2 shows a three phase indirect matrix converter.

Direct matrix converters are more efficient for high power high speed applications like drives and pumps compared to an indirect MC whereas an indirect MC is more efficient where the load has a high output current with a low power factor such as low speed and high loading applications. Hence the focus of this thesis will be on the direct matrix converter topology. They have reached a good level of technological maturity that allows their practical and industrial implementation in a variety of applications such as industrial drives, power supplies and aerospace applications [9]. The need for industrial drives with improved performance in terms of robustness, reliability

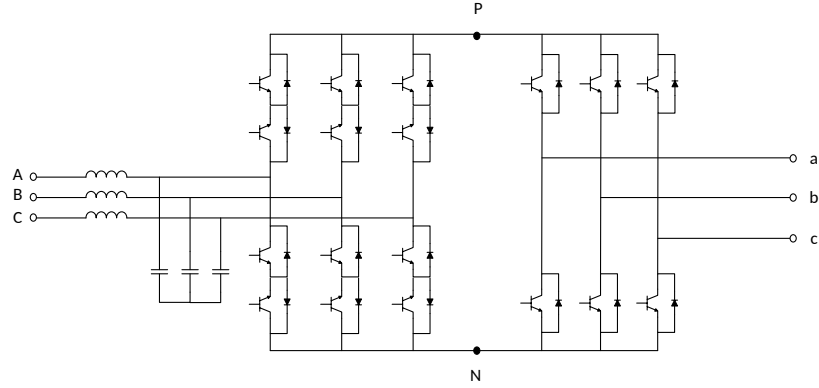


Figure 1.2: Three phase indirect matrix converter

and dynamic response further compounds the need to improve the controllability of matrix converters given the specific advantages in terms of reliability and robustness that they offer. This demands better control strategies and modulation techniques compared to the conventional ones. Several modulation [10] and control methods have been developed for a direct MC which will be discussed briefly in the next section.

1.2.1 Background study of existing modulation strategies for matrix converters

The procedure to produce the appropriate firing pulses to the nine bidirectional switches in the matrix converter is called modulation [11]. In the case of MC, this involves firing the devices at high speed in order to generate variable frequency and variable magnitude sinusoidal output voltages from fixed frequency, fixed magnitude input voltages. Modulation methods for matrix converters can be classified into two broad categories : scalar methods and vector modulation methods.

The foremost modulation method proposed for the generation of input and output waveforms of a matrix converter was proposed by Alesina and Venturini [12]. Although the method was first of its kind in terms of being able to control both output voltage and input current, it was not free from drawbacks. The major disadvantage of this method was the low voltage transfer ratio. The maximum voltage ratio that can

be achieved by this method was 0.5.

In 1985, Ziogas et al expanded the modulation method in [13] where the input voltages are rectified to form an imaginary DC link [14]. The DC link voltages are later inverted to produce the desired voltage waveforms. The main advantage of this method over the first method was its ability to achieve higher voltage ratio of 0.86. However it was limited by the poor control of input power factor. In 1988, Alesina and Venturini introduced another modulation strategy which rectifies the problems in the first method [15]. This method is called Optimal Venturini method [16]. The voltage ratio achieved by this method is 0.86 which is the theoretical maximum possible voltage ratio for a matrix converter to achieve control of both input current and output voltage. Roy and April proposed another scalar PWM modulation technique in 1989; that generated the switching signals directly from the input voltages and the reference output voltages of a matrix converter [17]. Other scalar modulation methods were proposed in [18, 19].

Another category of modulation method is the Vector Modulation which can be divided into carrier based PWM and space vector pulse width modulation (SVPWM). Carrier based PWM method applied to matrix converter proposed in many works in the literature [20, 21, 22, 23, 24, 25].

The SVPWM technique was first introduced by Huber and Borojevic in 1989 [26]. They proposed a two stage or indirect method in which the first is the rectification stage which results in a fictitious DC link and the second stage is the inversion stage to produce the desired voltage waveforms. Although it is able to achieve a voltage ratio of 0.86, control of input power factor is not possible using this approach. Later in 1991, the same authors proposed an advanced SVPWM method which can control the input power factor [27, 28]. In 1993 Casadei et al proposed a direct SVM method which does not involve any fictitious DC link [3]. Even though other modulation techniques which produce the same voltage transfer ratio as Optimum Venturini modulation have been proposed, SVM technique was generalized for matrix converters [3, 29, 30, 31, 32, 33, 34]. Since then, SVM has been a very popular approach for matrix converters

to obtain sinusoidal input currents and output voltages and has been applied in many applications in conjunction with feedback control strategies in order to regulate specific control variables [35, 36].

1.2.2 Background study of existing control strategies for MC

The control of matrix converter is very different from the conventional AC-AC converters with a DC-link capacitor. The absence of such an energy storage component in the circuit makes it difficult to control the bidirectional power flow in case of a matrix converter. This together with high number of switching states and the presence of rotating vectors makes the analysis and control of an induction motor using a MC even more complex [37]. Ideally MC should feed the load with sinusoidal currents while generating sinusoidal input currents with controlled power factor at the main supply [38]. This is typically achieved by using PI controllers for the output current and speed together with the SVM strategy for modulation. Behaviour of a matrix converter drive when subjected to polluted supply voltages has led to the research in modifying the existing modulation techniques [39, 40, 41, 42, 43].

Another control method that has been applied to matrix converter drives to control the flux and torque of induction machines is Direct Torque Control (DTC) [44, 45]. Even though DTC offers good control performance, the resonance at the input filter is a significant problem with this method. Efforts to improve the waveform quality of DTC for matrix converter permanent magnet drives have been carried out by Ortega et al [46].

Predictive control on the other hand was developed in the late seventies [47]. Different variations of predictive control associated with modulation techniques have been used for current control [48], PF correction [49] and drives [50, 51]. Model Predictive Control (MPC), considers a model of the system in order to predict its future behaviour over a time horizon. A cost function represents the desired behaviour of the system. MPC is an optimization problem where a sequence of future actuations is obtained

by minimizing the cost function in order to find the most suitable converter state to use. This approach has been successfully applied in recent years for the control of power converters and drives, such as for current control and power control of three phase Voltage Source Inverter (VSI) [52, 53] and for current and torque control for induction motor and permanent magnet motor drives [54, 55, 56, 57, 58]. It has also been exploited for other purposes, for instance common mode voltage reduction [59] and reduction of switching losses [60]. A predictive based controller with the aspects of a modulation strategy is the Dead-beat control[61]. This method has been used widely for control of power converters. However it does not offer the advantages of multi-objective control without cascaded loops.

Modulated Model Predictive Control (M2PC) is a control strategy that has an inbuilt modulation method proposed in [62] for a cascaded H-bridge converter. Due to a modulation scheme within the control method, it results in fixed switching frequency operation unlike other predictive control methods. This strategy has been applied successfully to an active-front end rectifier [63], 7-level cascaded H-bridge converter [64, 65, 66, 67], VSI [68] and an indirect matrix converter [69, 70].

A summary of the existing control and modulation strategies for a matrix converter is shown in Figure 1.3 [11].

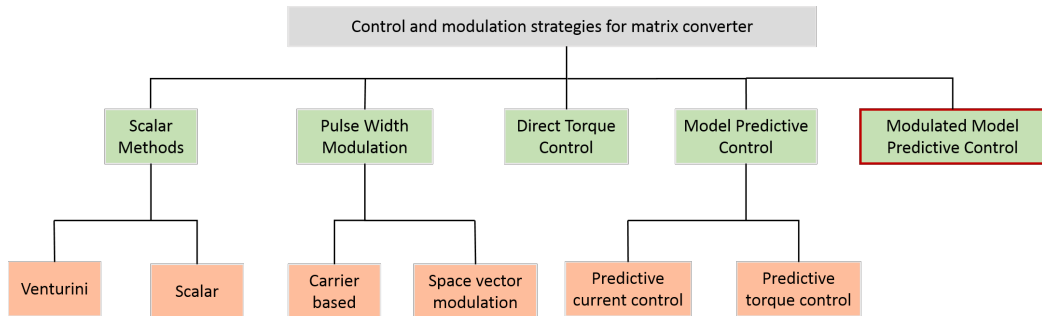


Figure 1.3: A summary of control and modulation strategies for matrix converter

1.2.3 Stability of Matrix Converter

Matrix converter has many advantages compared to other converter topologies in terms of unity input displacement factor, absence of large energy storage devices and sinusoidal input current and output voltage waveforms [71]. However, like any other converter, it distorts the input current and hence the supply voltage. The harmonics caused are in the range of the switching frequencies. The lower order harmonics are less compared to other converters that use rectifier front end solutions. In order to make the input current smoother, an input filter is necessary. However addition of the input filter is likely to cause instability issues in the system. Therefore design of an input filter is a crucial part as far as the stability of the system is concerned. Instability issues generally arise when the output power exceeds a certain limit [72]. The problems start with small oscillations in the input voltage waveforms of the converter and when these erroneous voltages are used to calculate the duty cycles of the converter which in turn creates larger disturbances which then again feed into the modulator even increasing the oscillations. Various methods such as small signal and large signal analysis have been used to study the stability issues associated with a matrix converter. Casadei et al in 2002 studied the effect of different types of input filter on the stability of a matrix converter drive [71]. A small signal analysis of the system is also performed and an equation for the maximum output power limit for stable operation of the system is obtained. It also showed that the power limit depends on the system parameters such as line impedance and the LC filter parameters. The paper shows that the LC filter with a damping resistor would be a more stable option than the conventional LC filter.

Casadei et al later developed this analysis and proposed that if the voltages used for the calculation of duty cycles in the MC are filtered using a digital filter; the stability of the system will be improved [73]. It also proved that this increases the maximum power transfer of the converter and the power limit of the system. This idea is supported and similar results are shown in [74]. Unstable operation of MC when feeding a constant load and a drive is discussed in [75] in detail. This paper uses the same small signal analysis method used in [73] and [74] to determine the

stability of the system in both the cases. This shows that input filter parameters have a significant influence where stability of the system is concerned. Effects of filtering the magnitude and angle of the MC input voltage vector is also studied and it is concluded that filtering the angle of the vector has no effect on the stability of the system whereas the amplitude filtering will make the system more stable. The stability of matrix converter drives is discussed in [75].

Even though some research has been done in the area of matrix converter stability as mentioned earlier, a robust input filter design without any stability issues is yet to be identified. The most widely used input filter design for MC is the LC filter with a damping resistor. The different factors that may affect MC stability also needs to be studied extensively. Hence the topic of stability of a matrix converter system is beyond the scope of this thesis.

1.3 Application of Matrix Converters

Matrix converters are predominantly considered for variable frequency AC drives with four quadrant operation. Due to the prospects for a light weight and concise implementation of MC, they are a strong candidate for aircraft applications. Wheeler et al proposed an aircraft actuator system based on a matrix converter permanent magnet motor drive [76]. Due to the absence of bulky DC link capacitors, integration of MC to the motors has been investigated by Klumpner et al in 2000 where a Matrix Converter Motor (MCM) is demonstrated [77]. Further more, a 30kW integrated drive with a matrix converter mounted on the motor end plate for industrial drive applications has been designed and tested by Apap et al [36].

In addition to industrial drives and aerospace, matrix converters have also been used for wind generation systems [78, 79, 80, 81], ac power utility units [82] and wireless energy transmission [83].

The world's foremost commercial matrix converter drive was developed and marketed

by the Japanese manufacturer Yasakawa in 2005. The drive has a direct matrix converter made of 1200V RB-IGBTs manufactured by Fuji Electric [2]. The main features of this drive are low harmonic distortion and power regeneration capability. Later Yasakawa extended the MC drive for medium voltage applications with input voltages of 3.3kV and 6.6kV. Eventhough matrix converters have been extensively researched in academia, it has achieved little market penetration mainly due to the reduced voltage ratio of 0.8666 [2].

1.4 Thesis Objectives

Model Predictive Control, (MPC) has been used to control power converters in the recent years with some examples mentioned in section 1.2.2. Even though control of currents, torque and flux are achieved by MPC, the switching frequency is variable and not fixed which causes problems for the design of a typical system such as input filter performance, component sizing and thermal management since the precise switching frequency and hence losses is unknown. Various methods are proposed in the literature to improve the applied vector sequence [84] or to introduce a modulation scheme inside the MPC algorithm [85]. However these methods involve complicated expressions for the switching time patterns and are not flexible enough to include other system requirements in the cost function. This is overcome by the M2PC which is proposed in [62]. This thesis investigates the performance of M2PC with a direct matrix converter. Another control method with a modulation scheme and results in fixed switching frequency operation has been discussed for a two-level inverter in [86]. Inspired from [86], this thesis will also look at the development of a novel control strategy for load current control of a direct matrix converter called Direct Predictive Current-error Vector Control (DPCVC). Another feature that is desirable for a control method is the fast transient response. The ability of MPC to deliver very fast transients has been exploited by using a predictive control algorithm together with a modulation method as in the case of M2PC and DPCVC. Thus the new control methods ensure fixed switching frequency together with very fast transients which is

difficult to be attained if a conventional PI controller is used. The objectives of this thesis can be summarised as follows:

- Design and implementation of M2PC for load current control of a direct matrix converter.
- Design and implementation of DPCVC for load current control of a direct matrix converter.
- Implementation of closed loop current control of a direct matrix converter using PI controllers.
- Implementation of MPC for load current control of a direct matrix converter.
- Comparison between PI controller, MPC, M2PC and DPCVC for a direct matrix converter.

1.5 Thesis Structure

The thesis is divided into two sections. The first section includes literature review and the theory of matrix converters which is discussed in chapters 1 and 2. The second section comprising chapters 3-6 focuses on the application of different control strategies for a direct matrix converter.

Chapter 2 focuses on the topology and operation of a direct matrix converter. This includes the different components of a matrix converter circuit such as input filter, clamp circuit and bidirectional switches. The commutation strategy commonly used in a matrix converter is the four-step commutation strategy which is also explained in detail. The output voltages in a matrix converter can be generated using Space Vector Modulation or Venturini Modulation which are also discussed in this chapter.

Chapter 3 gives a detailed account of the control of a direct matrix converter using the conventional Proportional-Integral (PI) controllers. The SVM strategy is used

in conjunction with PI to attain load current control of the converter feeding a RL load and an induction machine. The control of induction machine involves cascaded loops to control both speed and stator currents. This chapter explains the concept and theory behind indirect rotor flux orientation model of the machine which is used to control the speed and stator currents. Simulations are carried out for load current control using PI in Matlab Simulink and the results verifying the controller design are discussed in this chapter.

Chapter 4 explains the concept of Predictive control and the different types of predictive control. A detailed account of Model Predictive Control (MPC) and the various steps involved are discussed. The application of MPC for a direct matrix converter is also included in this chapter. To elaborate on the control procedure, application of MPC for a Voltage Source Inverter is explained. Load current control algorithm for a direct matrix converter is developed and simulation tests are carried out to verify the model. Simulation results for load current control of matrix converter feeding a RL load and an induction machine by MPC are also included.

Chapter 5 proposes a new predictive control strategy with an intrinsic modulation scheme called Modulated Model Predictive Control (M2PC). The objective of this method is to develop a control strategy that inherits the positive features of MPC and SVM resulting in fixed switching frequency operation. The concept and control procedure for a direct matrix converter is proposed and simulation results to verify the concept are also presented in this chapter. Another control strategy based on the features of predictive control called Direct Predictive Current-error Vector Control (DPCVC) is also proposed for matrix converter. The simulation results of this method applied to a matrix converter feeding an RL load and induction machine are presented in this chapter.

Chapter 6 presents the description of the matrix converter prototype in the laboratory. The experimental results for the four controllers such as PI Controller, MPC, M2PC and DPCVC for a direct matrix converter feeding an RL load are discussed. Implementation of closed loop control of an induction motor under no load condition

using the four control strategies discussed in this thesis are also included in this chapter. A comparison of the steady state and transient response when the load currents of a direct matrix converter are controlled by four different control methods are also analysed.

Chapter 7 describes the conclusions inferred from the simulation and experimental results of the different control strategies discussed in this thesis such as PI Controller, MPC, M2PC and DPCVC. The scope of M2PC and DPCVC and the future modifications and extensions to these control strategies are also addressed in this chapter.

Chapter 2

Matrix Converter Topology

2.1 Introduction

This chapter introduces the direct matrix converter (MC) topology and describes the operating principle. The generation of voltages in a MC using the mathematical model of the converter is presented. The advantages and disadvantages of a matrix converter compared to other converters like voltage source inverters are also addressed in this chapter. This chapter presents a review of the modulation methods and the commutation techniques that are used with a MC. The most common modulation strategy employed for a matrix converter known as the Space Vector Modulation(SVM) method is also explained in this chapter.

2.2 Matrix Converter

A typical three-phase to three-phase matrix converter consists of nine bidirectional switches connected in the form of a matrix as shown in Figure 2.1. In this converter, any input phase can be connected to any output phase to produce the desired output

voltage. The attractive feature of this converter is the absence of a DC-link capacitor for energy storage, enabling direct AC to AC conversion [4].

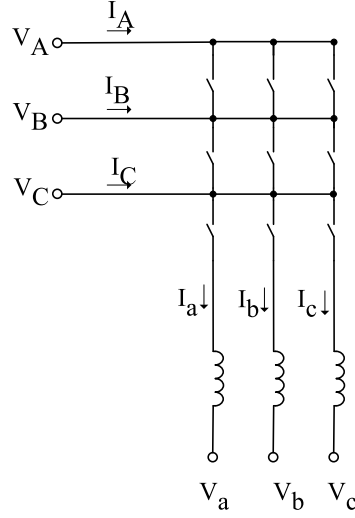


Figure 2.1: Direct matrix converter

Matrix converters have many advantages compared to conventional variable voltage, variable frequency topologies such as:

- The bidirectional switches enable power flow in both directions.
- Sinusoidal input and output currents can be obtained depending on the modulation strategy selected.
- The absence of a bulky DC-link capacitor makes the converter compact and offers potential for higher power density and higher operating temperatures.
- Unity input displacement factor can be achieved according to the modulation technique employed.

Like any other converter topology a matrix converter has its shortcomings. They are mainly :

- The theoretical maximum voltage transfer ratio between the input and output that can be achieved using a matrix converter is 0.866. This is explained in

detail in section 3.2. However this would not be a problem for applications like integrated drive [36] where the motor winding voltage can be selected.

- Due to the direct coupling between the input and output, a matrix converter is prone to instability i.e any change in the output current of the converter can affect the input side and vice versa.

These drawbacks together with high number of devices have influenced the adoption of this converter for industrial applications such as drives. However the advanced semiconductor modules designed specifically for matrix converters have generated considerable commercial interest for this converter [9]. Even though there are no commercial matrix converter modules available on the market to date, considerable interest has been shown over the recent years by companies like Siemens and Yaskawa.

2.3 Circuit topology

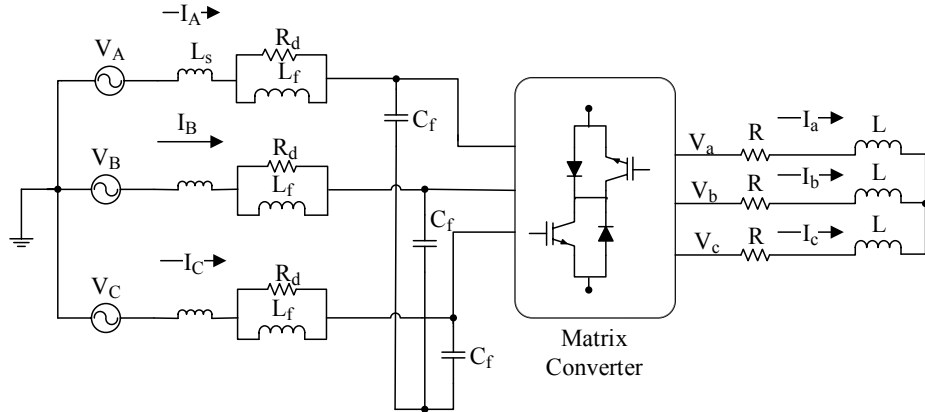


Figure 2.2: Schematic of matrix converter feeding a resistive inductive load

A schematic of a direct matrix converter system feeding a resistive inductive (RL) load is shown in Figure 2.2. The impedance of the input filter is L_f and L_s is the grid impedance if known. With an inductive load, this converter can be compared to a Voltage source inverter(VSI) at the output side and a current source inverter at the input side. An input filter is essential to remove the switching frequency harmonics

at the supply side and a clamp circuit to de-energize the load in case of overvoltage. The clamp circuit is also necessary due to the absence of a DC link capacitor. The clamp circuit provides a path for the currents if mis-commutation or open circuit occurs at the load side.

2.3.1 Input Filter

One of the main features of a matrix converter is its ability to produce sinusoidal input current and output voltage waveforms. The input current has less low order harmonics and sub-harmonics but may contain harmonics in the range of switching frequency and its multiples. The non sinusoidal currents flowing through the input impedances will distort the voltages and cause power quality issues at the supply side. The magnitude of these harmonic current components may be comparable with the fundamental and an input filter is necessary in this case to reduce these harmonics. An input filter is an important part of the MC circuit and the size of the filter is inversely proportional to the switching frequency. Therefore while designing input filters certain conditions should be taken into account. Some of these conditions are as follows:

- The filter should have a cut-off frequency much lower than the switching frequency of the converter.
- The voltage drop at the filter capacitor has to be as minimum as possible as this will affect the voltage ratio.
- It is desirable to minimize the size and volume of the input filter components.
- The phase shift between the input phase voltages and currents should be negligible.
- The input filter should not affect the stability of the system.

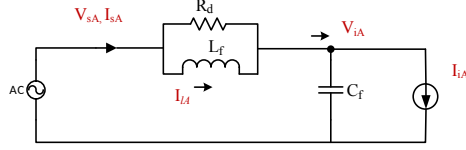


Figure 2.3: Input filter schematic

Considering all these conditions a simple LC filter can be used for the matrix converter system. Research has been done in order to design the optimum input filter for MC and some of these configurations are mentioned in the literature [75], [87]. If the load of MC is inductive, the input currents to or from the grid will be commutated and this leads to a discontinuous current with high harmonic content which needs to be filtered before injecting into the grid. The most common solution is to use a LC low-pass filter as shown in Figure 2.3. It can also include a damping resistor R_d , connected in parallel with the inductor [88].

A damping resistor is essential to reduce the oscillations at resonant frequency. The resonant frequency of the filter, f_{res} is the frequency at which the input signal's harmonic is amplified to the maximum [89]. It can be determined if the filter parameters are known from 2.1.

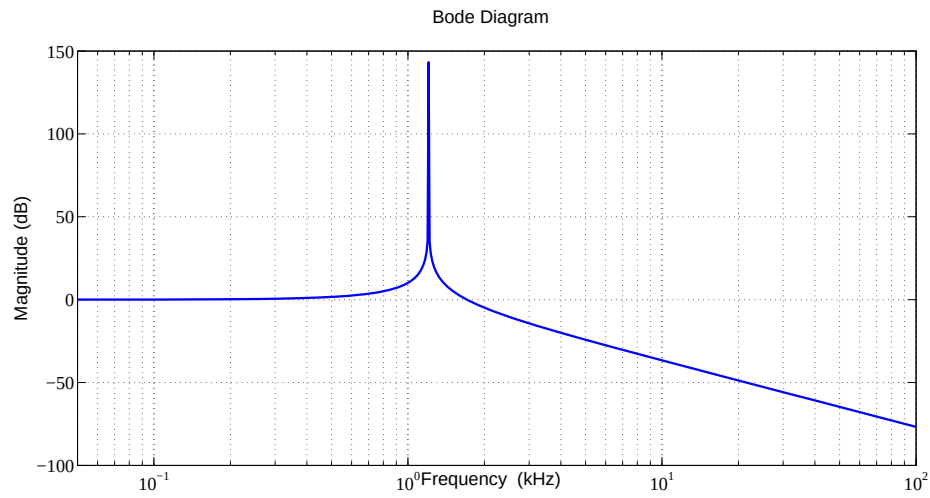
$$f_{res} = \frac{1}{2\pi\sqrt{L_f C_f}} \quad (2.1)$$

The effect of damping resistor on the input filter response can be seen by observing the frequency response of the input filter model in 2.2.

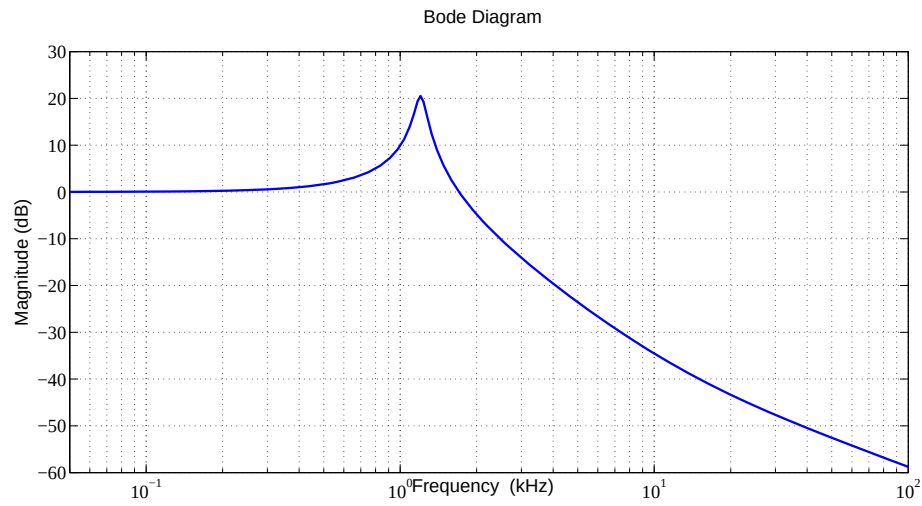
$$\dot{x}(t) = \begin{bmatrix} -\frac{1}{R_d C_f} & \frac{1}{C_f} \\ -\frac{1}{L_f} & 0 \end{bmatrix} x(t) + \begin{bmatrix} \frac{1}{R_d C_f} & -\frac{1}{C_f} \\ -\frac{1}{L_f} & 0 \end{bmatrix} u(t) \quad (2.2)$$

where $x(t) = \begin{bmatrix} V_i(t) \\ I_l(t) \end{bmatrix}$ and $u(t) = \begin{bmatrix} V_s(t) \\ I_i(t) \end{bmatrix}$

Figure 2.4 shows the frequency response of the input filter. From Figure 2.4 it is evident that the magnitude of the filter response at resonant frequency reduces with the inclusion of the damping resistor.



(a)



(b)

Figure 2.4: Bode plot of input filter response(a) without damping resistor (b)with damping resistor

2.3.2 Clamp Protection Circuit

To avoid overvoltages from the load or grid side, the converter needs to be equipped with a clamp protection circuit. It consists of two three-phase diode bridge rectifiers as shown in Figure 2.5 with a DC link capacitor and a balancing resistor [90]. As there are no free-wheeling diodes in a MC, when an overvoltage occurs at the grid or load side there is no alternative path for the load current if all the switches are turned off. This could lead to possible destruction of the switches. The clamp circuit protects the converter by transferring the energy to the capacitor from the point of overvoltage.

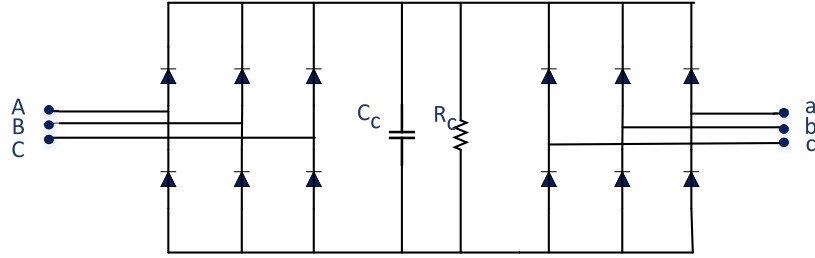


Figure 2.5: Clamp circuit

The clamp capacitor, C_c is selected such that it is able to receive the energy stored in the load (inductive) without exceeding the maximum voltage rating of the switches. The clamp resistor which is merely a balancing resistor is included to distribute the voltage across the clamp capacitor, C_c realised by connecting two capacitors in series evenly. The value of the balancing resistor can be calculated using 2.3.

$$R_c = \frac{V_c}{i_l} \quad (2.3)$$

where i_l is the leakage current of the capacitor and V_c is the voltage at C_c . The basic idea to design a clamp circuit is to calculate the energy stored in the load. For example, the energy stored in a three phase inductive load(balanced) is given by

$$\begin{aligned} E_{load} &= \frac{1}{2} L (i_a^2 + i_b^2 + i_c^2) \\ &= \frac{3}{4} L I^2 \end{aligned} \quad (2.4)$$

where L is the load inductance and I is the sum of squares of the three phase load currents i_a, i_b and i_c . The energy stored in the capacitor, E_c is given by

$$E_c = \frac{1}{2} C_c V_c^2 \quad (2.5)$$

where V_c is the voltage at C_c . During over voltage, the energy of the load will be transferred to the clamp capacitor given by E_{final} .

$$E_{final} = E_{load} + E_c$$

$$\frac{1}{2} C_c V_{cl}^2 = \frac{1}{2} C_c V_c^2 + \frac{1}{2} L I^2 \quad (2.6)$$

where V_{cl} is the maximum allowed clamp voltage. Based on V_{cl} and the rms load current, C_c can be calculated using (2.5).

2.3.3 Bidirectional Switch

The matrix converter allows bidirectional power flow which is attributed to the presence of bidirectional switches. These switches are capable of conducting current and blocking voltage in both directions. They are made up of discrete devices to fulfil this requirement [4]. There are different topologies for bidirectional switches out of which the most common ones are common emitter anti-parallel Insulated-gate Bipolar Transistor (IGBT) configuration and common collector anti-parallel IGBT configuration.

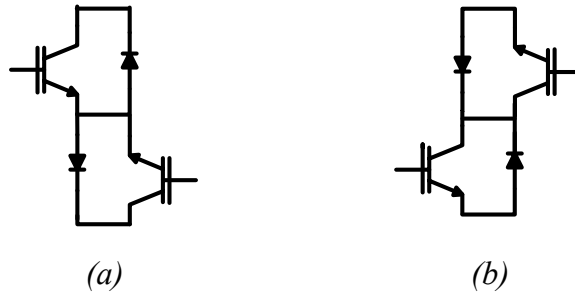


Figure 2.6: Bidirectional switches: (a) Common emitter configuration, (b) Common collector configuration

The common emitter configuration shown in Figure 2.6(a) is made of two diodes and two IGBTs connected in anti-parallel manner. The diodes enable reverse blocking capability in this topology. The main advantages of this configuration includes reduced conduction losses as only two devices conduct current at any time allowing independent control of each current path and the ability to drive two IGBTs with respect to the common emitter. However this configuration requires isolated power supplies for each switch, hence a total of nine power supplies are needed. Therefore it is not possible to drive all the switches with respect to the same point.

The common collector configuration shown in Figure 2.6(b) is very similar to the previous arrangement except for the common point. In this case, only six isolated power supplies are required to supply the gate drivers. It can be used only when the inductance between the devices sharing the power supply is low which is the case in matrix converter modules. The conduction losses are the same as that of the common emitter configuration. For this reason for high power applications common-emitter configuration is suitable.

Compared to the conventional 2-level VSIs, matrix converter has higher number of switches. This means that the stress per device for matrix converter is much less compared to a VSI.

2.3.4 Commutation of MC

Unlike conventional voltage source inverters, matrix converters do not have any free wheeling diodes. Hence it is difficult to commutate between two switches and the commutation has to be controlled at all times. This also has to be done without causing short circuits in the input lines and open circuits in the output lines. This problem was first solved by Burany [91] and Oyama [92] in 1989 by developing a multi-stage commutation technique that allows safe change of switching states. A standard multi-stage commutation requires four steps. Multi-step commutation can be based on the sign of the input voltage [92, 93, 94] or the direction of the output

current [94, 95]. This section gives a brief account of the simple and advanced current commutation techniques.

2.3.4.1 Simple current commutation

Simple commutation methods for MC can be divided into two categories: Overlap commutation and Dead-time commutation.

Overlap Commutation as the name suggests has an overlap region because the incoming switch is turned ON before the outgoing switch is turned OFF. Even though this method prevents open-circuiting the load current, it requires extra line inductance between the input capacitors and the switch to limit the current during the overlap period. During this period, these inductors conduct the input current and needs to be large. This is not desirable as it is not cost effective. Also, the output voltage waveform will be distorted during the overlap period [96].

Dead-time Commutation where the load current freewheels the anti-parallel diodes is normally employed in inverter circuits. This technique when applied to matrix converter causes open circuit of the inductive load resulting in the need for clamp protection circuits to provide a path for the load current and limit the resulting over voltage [97].

The methods mentioned above requires additional circuitry for protection even though the bidirectional switches can be used. A number of advanced commutation techniques have been studied [98, 99, 100, 101] for matrix converters that solves the matrix converter commutation problem.

2.3.4.2 Advanced commutation techniques

Four-step commutation [102] based on the output current direction is one of the advanced commutation techniques used with matrix converters. Commutation is

achieved with different unidirectional current paths to avoid the short circuits and open circuits mentioned before. This technique can be explained with the help of Figure 2.7 depicting two bidirectional switches undergoing commutation.

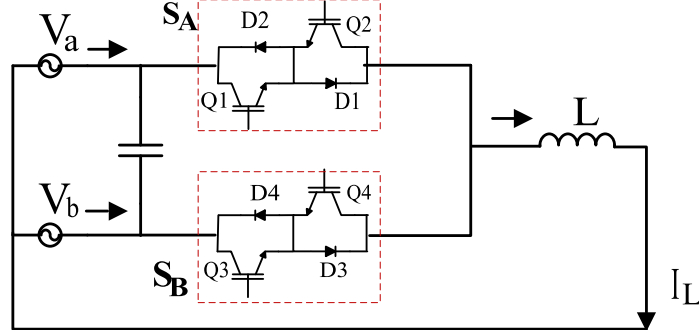


Figure 2.7: Bidirectional switches during commutation in a 2-phase to single phase MC

In this method, conducting and non-conducting halves of the outgoing switch are governed by the current direction. At steady state both the devices in active bidirectional switch are gated allowing current flow in both directions. It is assumed that the load current direction is as shown in Figure 2.7 and S_A is closed. The method is explained further based on this assumption. When S_B needs to be turned ON, the current direction is used to determine the non-conducting device in the active switch. This device; in this case Q2 is turned off. Then the device that will conduct the current in the incoming switch, in this case Q3 is gated. Depending on the polarity of the input voltage the load current transfers to the incoming device either at this point or when the outgoing device Q1 is turned off. Q4 is then turned ON to allow the current reversal. This process is shown in the timing diagram in Figure 2.8. The characteristics of the device determines the delay between each switching event. Inaccurate information about the output current direction can lead to open circuit of the load that could be fatal for the converter.

When only the conducting device is gated, it results in a **two-step commutation** method [99]. This leads to reduced commutation times and smaller minimum pulse width. This is usually advantageous as minimum pulse width leads to higher maximum voltage transfer ratio of the converter.

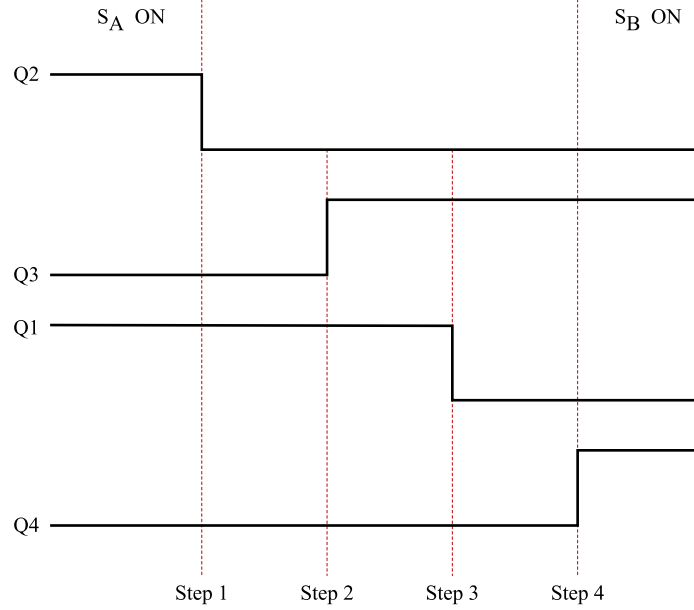


Figure 2.8: Timing diagram for four step current commutation

All the methods mentioned above requires the output line current direction which can be difficult to determine accurately in a switching power converter with conventional current sensors. A technique proposed by L.Empringham et.al [103] avoids this problem by using the voltage across the bidirectional switch to determine the current direction. This method makes use of intelligent gate drive circuit that controls the firing of the IGBTs and detects the direction of current flow in the bidirectional switch. This current direction information is then passed to all the gate drivers on the same output phase to ensure that they contribute to safe commutation.

Voltage Magnitude Based commutation is another technique based on the input voltage polarity between the commutating switches [16]. It uses the input voltage polarity to define a four step sequence combining the overlap and dead-time commutation. For example consider a commutation from S_A to S_B in Figure 2.7. If $V_A > V_B$, Q3 can be turned ON since D3 is reversed biased. When Q1 is turned off, it will commute to Q3 provided the current direction is same as shown in the Figure 2.8. Else the commutation will occur when Q4 is turned ON. The commutation is then completed by turning off Q2. In case of a mistake in the input voltage direction, the resulting commutation sequence will cause a line to line short circuit of the supply

which is difficult to protect.

2.4 Generation of Output voltages in a MC

The principle of output voltage generation with the mathematical model of the matrix converter will be discussed in this chapter based on a three phase to three phase matrix converter. As mentioned earlier, a matrix converter can be regarded as a voltage source inverter at the input and a current source inverter at the output. This is because the input of the matrix converter should never be short circuited and the output of the matrix converter should never be open circuited. To ensure this, matrix converter is commonly used in conjunction with an inductive load such as an induction machine and an input filter. The input filter is also required for the circulation of the high frequency switching harmonics. This ensures that the input side of the converter is voltage stiff and output side of the converter is current stiff. A schematic of the matrix converter showing the nine switches arranged in the form of a matrix is shown in the Figure 2.9.

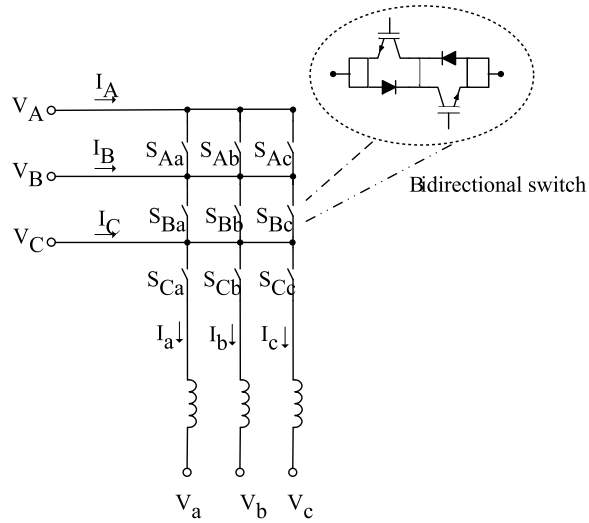


Figure 2.9: Matrix converter

Throughout this thesis, the input phases will be represented by the upper case suffices and the output phases will be represented by the lower case suffices. The switches

are represented by the input and output phase suffices connected to each of them. Assuming ideal switching, the mathematical model of the matrix converter shown in Figure 2.9 can be derived in terms of the switches. The switching function S_{ij} is defined as:

$$S_{ij} = \begin{cases} 1 & \text{switch } S_{ij} \text{ is closed} \\ 0 & \text{switch } S_{ij} \text{ is open} \end{cases} \quad (2.7)$$

where $i = A, B, C$ and $j = a, b, c$. This nomenclature for the input and output sides of the converter will be used throughout this thesis. The instantaneous values of the output voltage and the input currents of the MC can be obtained from the supply voltages and output currents respectively. The output voltage of the MC is obtained from the supply voltages and the switching functions as shown in 2.8.

$$\begin{bmatrix} V_a(t) \\ V_b(t) \\ V_c(t) \end{bmatrix} = \begin{bmatrix} S_{Aa}(t) & S_{Ba}(t) & S_{Ca}(t) \\ S_{Ab}(t) & S_{Bb}(t) & S_{Cb}(t) \\ S_{Ac}(t) & S_{Bc}(t) & S_{Cc}(t) \end{bmatrix} \cdot \begin{bmatrix} V_A(t) \\ V_B(t) \\ V_C(t) \end{bmatrix} \quad (2.8)$$

Similarly the input currents of the converter can be calculated from the output currents using the equation

$$\begin{bmatrix} I_A(t) \\ I_B(t) \\ I_C(t) \end{bmatrix} = \begin{bmatrix} S_{Aa}(t) & S_{Ba}(t) & S_{Ca}(t) \\ S_{Ab}(t) & S_{Bb}(t) & S_{Cb}(t) \\ S_{Ac}(t) & S_{Bc}(t) & S_{Cc}(t) \end{bmatrix}^T \cdot \begin{bmatrix} I_a(t) \\ I_b(t) \\ I_c(t) \end{bmatrix} \quad (2.9)$$

As mentioned earlier, there are restrictions to the possible combinations of input and output phases in a matrix converter. This is due to the absence of free-wheeling diodes in a matrix converter. To ensure that the input side is never short circuited and the output side is never open circuited, the following conditions should be satisfied.

$$\begin{aligned} S_{Aa}(t) + S_{Ba}(t) + S_{Ca}(t) &= 1 \\ S_{Ab}(t) + S_{Bb}(t) + S_{Cb}(t) &= 1 \\ S_{Ac}(t) + S_{Bc}(t) + S_{Cc}(t) &= 1 \end{aligned} \quad (2.10)$$

So the modulation problem can be identified as determining the switching states that satisfies the input-output relationship and meet the constraints in 2.10 for each cycle period. The output voltage is produced by a combination of the input voltages and the switching states applied for their respective duty cycles. A typical output voltage waveform of a matrix converter at 50Hz is shown in Figure 2.10.

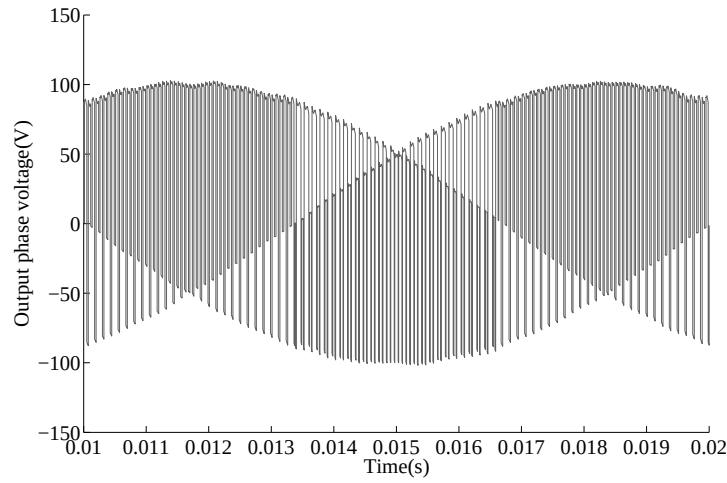


Figure 2.10: Output line voltage of a matrix converter

The input current of the matrix converter shown in Figure 2.11 is composed of the three phase output currents together with the blank periods during which the output current freewheels through the switch matrix. Figure 2.12 shows the input current after the filtering action.

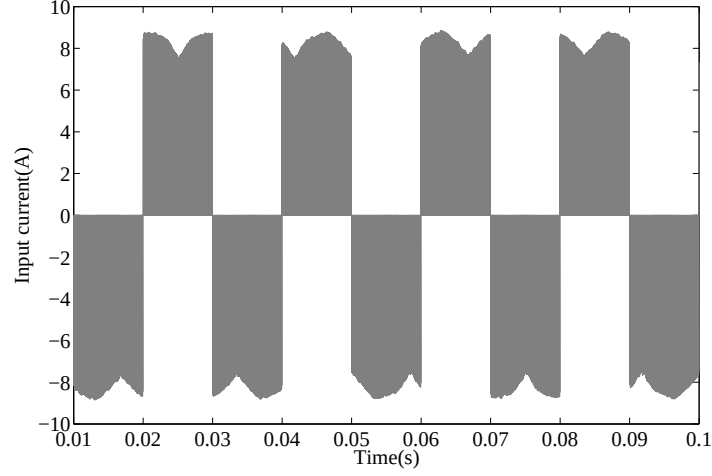


Figure 2.11: Input current of a matrix converter without input filter

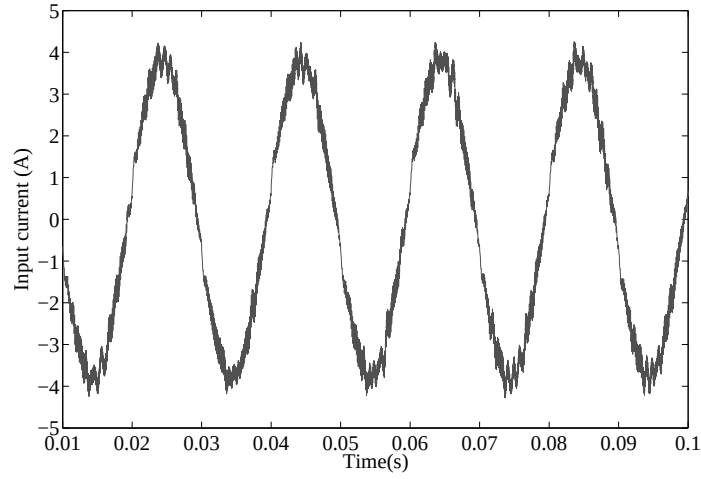


Figure 2.12: Input current of a matrix converter with input filter

The switching period is fixed and is similar to the period of the carrier wave in conventional pulse width modulation technique. The behavior of the converter at lower frequencies can be determined by the modulation duty cycle. The modulation indices for each switch can be defined such that $m_{Aa}(t) = t_{Aa}/T_s$ where t_{Aa} is the time for which the switch S_{Aa} is ON and T_s is the sampling time.

2.5 Modulation strategies for MC

The main objective of a modulation method is to synthesize the required output voltage according to the reference voltage demand and to obtain a sinusoidal input current with near to unity power factor. It is desirable to obtain least distortion in the input currents and the output voltage with lower power losses in the system.

A number of control and modulation techniques have been proposed and studied for a matrix converter. The classification of these methods can be divided into different categories based on the approach to solve the modulation problem [90]. One such classification is **direct** and **indirect** methods [4] based on how the switching state is selected. In direct methods the selection is made based on the fact that at a given period of time, each input phase will be connected to an output phase. The indirect methods select the switching states based on the assumption of a fictitious DC link which results in a rectification and an inversion stage. This enables to consider the effects of the input and output separately and then combine these results to generate the switching pattern.

Another way is to categorize the modulation techniques into **scalar** and **space vector** strategies. Scalar methods solve the modulation problem based on scalar variables and the space vector methods select the switching states based on the space vector representation. In this thesis, two modulation techniques which are the most popular and well researched will be discussed in detail: the Venturini modulation and space vector modulation (SVM). These methods also represent the categories mentioned above. Venturini modulation is a scalar and direct modulation technique and SVM is a space vector strategy.

2.5.1 Venturini Modulation

Venturini modulation, introduced in 1980s [104] addressed the modulation problem of the MC. In this method, the target output voltage and input current waveforms are

generated by the sequential combination of each input phase with output phase during a particular time period. Although this method solved the modulation problem of the MC, the maximum voltage transfer ratio it could achieve was only 50%. The method was later optimized to increase the voltage transfer ratio to 86% [16] and was called the optimum Venturini method. The instantaneous model of the converter can be represented in the form of the following equations.

$$\begin{bmatrix} \bar{v}_a \\ \bar{v}_b \\ \bar{v}_c \end{bmatrix} = \underbrace{\begin{bmatrix} m_{Aa} & m_{Ba} & m_{Ca} \\ m_{Ab} & m_{Bb} & m_{Cb} \\ m_{Ac} & m_{Bc} & m_{Cc} \end{bmatrix}}_M \cdot \begin{bmatrix} v_A \\ v_B \\ v_C \end{bmatrix} \quad (2.11)$$

$$\begin{bmatrix} \bar{i}_A \\ \bar{i}_B \\ \bar{i}_C \end{bmatrix} = \underbrace{\begin{bmatrix} m_{Aa} & m_{Ab} & m_{Ac} \\ m_{Ba} & m_{Bb} & m_{Bc} \\ m_{Ca} & m_{Cb} & m_{Cc} \end{bmatrix}}_{M^T} \cdot \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \quad (2.12)$$

where $\bar{\mathbf{v}}$ and $\bar{\mathbf{i}}$ are the low frequency output voltage and input current vectors respectively. M is the transfer matrix consisting of the duty cycles for all the nine bidirectional switches. Assuming sinusoidal supply, the input voltages to the MC can be given by

$$\begin{bmatrix} v_A \\ v_B \\ v_C \end{bmatrix} = \begin{bmatrix} V_i \cos(w_i t) \\ V_i \cos(w_i t + 2\pi/3) \\ V_i \cos(w_i t + 4\pi/3) \end{bmatrix} \quad (2.13)$$

Due to the inductive nature of the load, the output currents can be correctly assumed to be sinusoidal in nature. This results in the output current equations given below.

$$\begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} = \begin{bmatrix} I_o \cos(w_o t + \phi_o) \\ I_o \cos(w_o t + \phi_o + 2\pi/3) \\ I_o \cos(w_o t + \phi_o + 4\pi/3) \end{bmatrix} \quad (2.14)$$

where I_o is the amplitude, w_o is the desired output frequency and ϕ_o is the load angle. The reference voltage can be represented in the form of the input voltages as shown

in 2.15.

$$\begin{bmatrix} \bar{v}_a \\ \bar{v}_b \\ \bar{v}_c \end{bmatrix} = \begin{bmatrix} V_{ref} \cos(w_o t) \\ V_{ref} \cos(w_o t + 2\pi/3) \\ V_{ref} \cos(w_o t + 4\pi/3) \end{bmatrix} \quad (2.15)$$

where $V_{ref} = qV_i$ and q is the voltage ratio. Once the reference voltage and the load current is known, the desired input current can be expressed as shown in 2.16.

$$\begin{bmatrix} \bar{i}_A \\ \bar{i}_B \\ \bar{i}_C \end{bmatrix} = \begin{bmatrix} I_i \cos(w_i t + \phi_i) \\ I_i \cos(w_i t + \phi_i + 2\pi/3) \\ I_i \cos(w_i t + \phi_i + 4\pi/3) \end{bmatrix} \quad (2.16)$$

where I_i is the amplitude of the input current and ϕ_i the input displacement angle caused by the input filter. The modulation problem can be solved by considering the power balance criterion of the system. The active input power and output power must be equal due to the absence of DC link capacitor or other energy storage devices. As a result the power balance condition shown in 2.17 must hold true.

$$P_i = \frac{3V_i I_i \cos(\phi_i)}{2} = \frac{3qV_i I_o \cos(\phi_o)}{2} = P_o \quad (2.17)$$

From the above condition, the maximum input current amplitude is limited to $I_i = q \frac{\cos(\phi_o)}{\cos(\phi)}$. To solve the modulation problem, the transfer matrix M should be determined. This is possible by solving equations 2.13 to 2.16 to satisfy the conditions imposed by 2.11 and 2.12. The third set of conditions to be able to solve for the transfer matrix is the relation between the duty cycles of each of the output phases of MC.

$$\sum_{x=A,B,C} m_{xa} = \sum_{x=A,B,C} m_{xb} = \sum_{x=A,B,C} m_{xc} \quad (2.18)$$

The solution for the modulation problem can thus be solved resulting in two valid solutions as shown in 2.19 and 2.20.

$$M_1 = \begin{bmatrix} 1 + 2q \cos(w_m t) & 1 + 2q \cos(w_m t - 2\pi/3) & 1 + 2q \cos(w_m t - 4\pi/3) \\ 1 + 2q \cos(w_m t - 4\pi/3) & 1 + 2q \cos(w_m t) & 1 + 2q \cos(w_m t - 2\pi/3) \\ 1 + 2q \cos(w_m t - 2\pi/3) & 1 + 2q \cos(w_m t - 4\pi/3) & 1 + 2q \cos(w_m t) \end{bmatrix} \quad (2.19)$$

with $w_m = w_o - w_i$ and

$$M_2 = \begin{bmatrix} 1 + 2q \cos(w_m t) & 1 + 2q \cos(w_m t - 2\pi/3) & 1 + 2q \cos(w_m t - 4\pi/3) \\ 1 + 2q \cos(w_m t - 2\pi/3) & 1 + 2q \cos(w_m t - 4\pi/3) & 1 + 2q \cos(w_m t) \\ 1 + 2q \cos(w_m t - 4\pi/3) & 1 + 2q \cos(w_m t) & 1 + 2q \cos(w_m t - 2\pi/3) \end{bmatrix} \quad (2.20)$$

with $w_m = -(w_o + w_i)$ The solutions vary in the range of control of the input displacement angle. The first solution M_1 results in an input displacement angle equal to the load angle: $\phi_i = \phi_o$. On the other hand, M_2 results in an input displacement angle equal to the reverse of load angle: $\phi_i = -\phi_o$. Combining these both solutions will enable the control of input displacement angle within the range shown in 2.21.

$$\phi_o \leq \phi_i \leq \phi_o \quad (2.21)$$

The control is attained by adding weighting factors to the combination of solutions such that the final transfer matrix can be given by

$$M = \alpha_1 M_1 + \alpha_2 M_2 \quad (2.22)$$

where $\alpha_1 + \alpha_2 = 1$. The system will work at unity pf when $\alpha_1 = \alpha_2$ which means $\phi_i = 0$. In the case of a unity pf system the determination of M is simplified and is given by 2.23.

$$m_{xy} = \frac{1}{3} \left[1 + \frac{2v_x \bar{v}_y}{V_i} \right] \quad \text{for } x \in [A, B, C] \quad \text{and } y \in [a, b, c] \quad (2.23)$$

The drawback of this solution is the low voltage transfer ratio of 0.5 which is attributed to the fact that the target output voltage must fit within the input voltage envelope

at all times as shown in Figure 2.13.

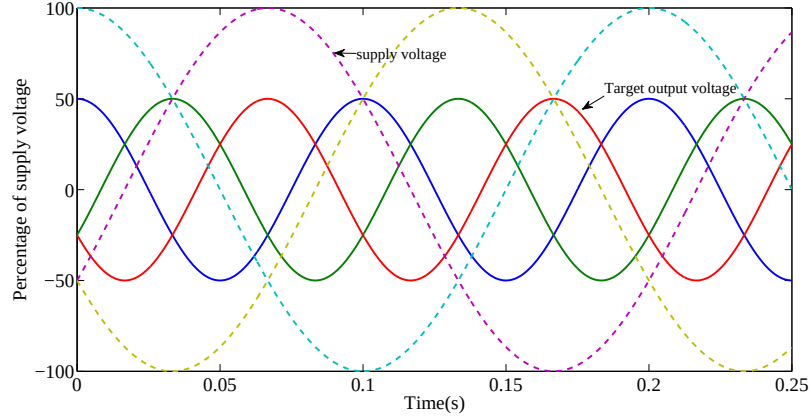


Figure 2.13: Target output voltages at 50% voltage ratio limit

Optimum Venturini method has successfully improved the voltage ratio by third harmonic injection of the input and output fundamental frequencies [16]. In the case of optimum Venturini method, the target output voltage is given by

$$\begin{bmatrix} \bar{v}_a \\ \bar{v}_b \\ \bar{v}_c \end{bmatrix} = V_{ref} \begin{bmatrix} \cos(w_o t) - \frac{1}{6} \cos(3w_o t) + \frac{1}{2\sqrt{3}} \cos(3w_i t) \\ \cos(w_o t + 2\pi/3) - \frac{1}{6} \cos(3w_o t) + \frac{1}{2\sqrt{3}} \cos(3w_i t) \\ \cos(w_o t + 4\pi/3) - \frac{1}{6} \cos(3w_o t) + \frac{1}{2\sqrt{3}} \cos(3w_i t) \end{bmatrix} \quad (2.24)$$

The triple harmonics injected into the reference voltage will be canceled in the output line voltage. The transfer matrix for this case can then be calculated. For unity pf, the solution for M is given by 2.25.

$$m_{xy} = \frac{1}{3} \left[1 + \frac{2v_x \bar{v}_y}{V_i} + \frac{4q}{3\sqrt{3}} \sin(w_i t + \beta_x) \sin(3w_i t) \right] \quad (2.25)$$

for $x \in [A, B, C]$ and $y \in [a, b, c]$.

where $\beta_A = 0, \beta_B = 2\pi/3$ and $\beta_C = 4\pi/3$. The maximum voltage transfer ratio that can be achieved in this case is $\sqrt{3}/2$. A graphical representation of this is shown in Figure 2.14. The output voltage target waveforms now utilize the input voltage envelope more effectively and hence this method has more practical significance.

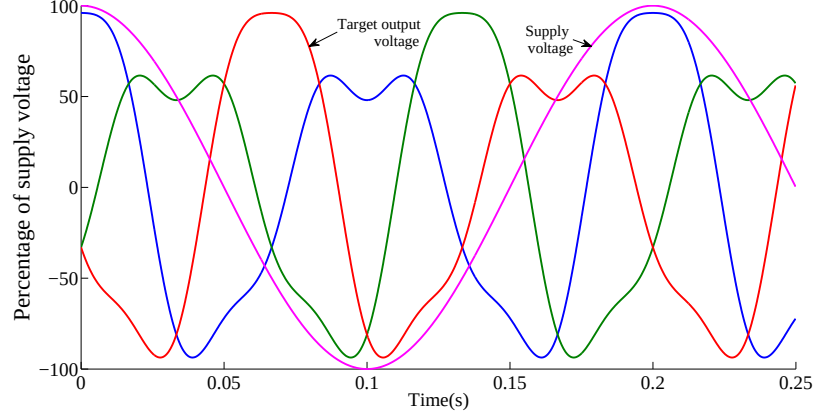


Figure 2.14: Target output voltages at 86% voltage ratio limit

The time for which each bidirectional switch is ON during a cycle T_{seq} is then obtained from the duty cycles.

$$t_{xy} = m_{xy}T_{seq} \quad \text{for } x \in [A, B, C] \quad \text{and } y \in [a, b, c] \quad (2.26)$$

The sum of the on-times for each of the three bidirectional switches of a given output phase should be equal to T_{seq} . The average switching frequency of IGBT can then be calculated as shown in 2.27.

$$f_s = \frac{1}{T_{seq}} \quad (2.27)$$

2.5.2 Space Vector Modulation

At the end of 1980s L.Huber and D.Borojevic [26] proposed the Space Vector Modulation (SVM) for MC extending the modulation method used on inverters and rectifiers. SVM attains the theoretical maximum output voltage ratio of 0.8666 without the addition of third harmonic components of the input and output. From Figure 2.2 the output voltage and input current space vectors can be defined as

$$\mathbf{v}_o = \frac{2}{3}(v_a + \mathbf{a}v_b + \mathbf{a}^2v_c) \quad (2.28)$$

$$\mathbf{i}_i = \frac{2}{3}(i_A + \mathbf{a}i_B + \mathbf{a}^2i_C) \quad (2.29)$$

where $\mathbf{a}=e^{j\frac{2\pi}{3}}$.

For a conventional 2-level three phase inverter, there will be eight switching states resulting in eight voltage space vectors. Six of these vectors will form the vertices of the a regular hexagon and are of equal magnitudes. These vectors are called the *active vectors*. Two of the remaining vectors will have zero magnitude and are called *zero vectors*. In an inverter, the desired voltage vector is created by the combination of two adjacent active vectors and one of the zero vectors. This technique is well established in the case of a 2-level inverter, however applying this method to MC can become cumbersome as there are 27 possible switching states with time varying input voltages. Firstly SVM was applied to MC to control the output voltage by using two active vectors for a sampling period [26]. It was later used to control the output voltage and input current angle by using four active vectors[105, 33]. This method is explained in detail in this section.

To elaborate on the SVM for MC, it is essential to consider the space vectors for output voltage and input current for every switching state. The 27 switching states of a MC can be divided into three categories or groups depending on the type of output voltage or input current vectors that results from the various switching combinations. They are

- I **Active Vectors:** These vectors are produced when two output phases are connected to the same input phase and the third connected to a different input phase. There are 18 of these states and the voltage or current vectors resulting from these states will have fixed direction spaced 60° apart with varying magnitudes. The amplitude of these vectors depends on the instantaneous values of the input line voltages and output currents.
- II **Rotating Vectors:** When each output phase is connected to a different input phase, the resulting vectors will have a constant amplitude. However the direction of the output voltage and input current vectors resulting from these 6

states varies with the angular frequencies of the output and input respectively.

III **Zero Vectors:** Three remaining states are when all the three output phases are connected to the same input phase. The resulting vectors will have zero amplitude.

Table 2.1 shows the magnitude and angle of the output voltage and input current vectors produced by all the 27 switching states. It also shows the vectors belonging to each of the groups mentioned earlier.

Table 2.1: Space vectors for all the switching combinations for a MC

Group	#	Vector Name	SwitchesON	Output Voltage	Input Current	Voltage Vector		Current Vector	
						Magnitude	Angle	Magnitude	Angle
I	1	+1	$S_{Aa} S_{Bb} S_{Bc}$	$v_A v_B v_B$	$i_a -i_a 0$	$\frac{2}{3}v_{AB}$	0	$\frac{2}{\sqrt{3}}i_a$	$-\frac{\pi}{6}$
	2	-1	$S_{Ba} S_{Ab} S_{Ac}$	$v_B v_A v_A$	$-i_a i_a 0$	$-\frac{2}{3}v_{AB}$	0	$-\frac{2}{\sqrt{3}}i_a$	$-\frac{\pi}{6}$
	3	+2	$S_{Ba} S_{Cb} S_{Cc}$	$v_B v_C v_C$	$0 i_a -i_a$	$\frac{2}{3}v_{BC}$	0	$\frac{2}{\sqrt{3}}i_a$	$\frac{\pi}{2}$
	4	-2	$S_{Ca} S_{Bb} S_{Bc}$	$v_C v_B v_B$	$0 -i_a i_a$	$-\frac{2}{3}v_{BC}$	0	$-\frac{2}{\sqrt{3}}i_a$	$\frac{\pi}{2}$
	5	+3	$S_{Ca} S_{Ab} S_{Ac}$	$v_C v_A v_A$	$-i_a 0 i_a$	$\frac{2}{3}v_{CA}$	0	$\frac{2}{\sqrt{3}}i_a$	$\frac{7\pi}{6}$
	6	-3	$S_{Aa} S_{Cb} S_{Cc}$	$v_A v_C v_C$	$i_a 0 -i_a$	$-\frac{2}{3}v_{CA}$	0	$-\frac{2}{\sqrt{3}}i_a$	$\frac{7\pi}{6}$
	7	+4	$S_{Ba} S_{Ab} S_{Bc}$	$v_B v_A v_B$	$i_b -i_b 0$	$\frac{2}{3}v_{AB}$	$\frac{2\pi}{3}$	$\frac{2}{\sqrt{3}}i_b$	$-\frac{\pi}{6}$
	8	-4	$S_{Aa} S_{Bb} S_{Ac}$	$v_A v_B v_A$	$-i_b i_b 0$	$-\frac{2}{3}v_{AB}$	$\frac{2\pi}{3}$	$-\frac{2}{\sqrt{3}}i_b$	$-\frac{\pi}{6}$
	9	+5	$S_{Ca} S_{Bb} S_{Cc}$	$v_C v_B v_C$	$0 i_b -i_b$	$\frac{2}{3}v_{BC}$	$\frac{2\pi}{3}$	$\frac{2}{\sqrt{3}}i_b$	$\frac{\pi}{2}$
	10	-5	$S_{Ba} S_{Cb} S_{Bc}$	$v_B v_C v_B$	$0 -i_b i_b$	$-\frac{2}{3}v_{BC}$	$\frac{2\pi}{3}$	$-\frac{2}{\sqrt{3}}i_b$	$\frac{\pi}{2}$
	11	+6	$S_{Aa} S_{Cb} S_{Ac}$	$v_A v_C v_A$	$-i_b 0 i_b$	$\frac{2}{3}v_{CA}$	$\frac{2\pi}{3}$	$\frac{2}{\sqrt{3}}i_b$	$\frac{7\pi}{6}$
	12	-6	$S_{Ca} S_{Ab} S_{Cc}$	$v_C v_A v_C$	$i_b 0 -i_b$	$-\frac{2}{3}v_{CA}$	$\frac{2\pi}{3}$	$-\frac{2}{\sqrt{3}}i_b$	$\frac{7\pi}{6}$
	13	+7	$S_{Ba} S_{Bb} S_{Ac}$	$v_B v_B v_A$	$i_c -i_c 0$	$\frac{2}{3}v_{AB}$	$\frac{4\pi}{3}$	$\frac{2}{\sqrt{3}}i_c$	$-\frac{\pi}{6}$
	14	-7	$S_{Aa} S_{Ab} S_{Bc}$	$v_A v_A v_B$	$-i_c i_c 0$	$-\frac{2}{3}v_{AB}$	$\frac{4\pi}{3}$	$-\frac{2}{\sqrt{3}}i_c$	$-\frac{\pi}{6}$
	15	+8	$S_{Ca} S_{Cb} S_{Bc}$	$v_C v_C v_B$	$0 i_c -i_c$	$\frac{2}{3}v_{BC}$	$\frac{4\pi}{3}$	$\frac{2}{\sqrt{3}}i_c$	$\frac{\pi}{2}$
	16	-8	$S_{Ba} S_{Bb} S_{Cc}$	$v_B v_B v_C$	$0 -i_c i_c$	$-\frac{2}{3}v_{BC}$	$\frac{4\pi}{3}$	$-\frac{2}{\sqrt{3}}i_c$	$\frac{\pi}{2}$
	17	+9	$S_{Aa} S_{Ab} S_{Cc}$	$v_A v_A v_C$	$-i_c 0 i_c$	$\frac{2}{3}v_{CA}$	$\frac{4\pi}{3}$	$\frac{2}{\sqrt{3}}i_c$	$\frac{7\pi}{6}$
	18	-9	$S_{Ca} S_{Cb} S_{Ac}$	$v_C v_C v_A$	$i_c 0 -i_c$	$-\frac{2}{3}v_{CA}$	$\frac{4\pi}{3}$	$-\frac{2}{\sqrt{3}}i_c$	$\frac{7\pi}{6}$
II	19	R_1	$S_{Aa} S_{Bb} S_{Cc}$	$v_A v_B v_C$	$i_a i_b i_c$	V_{im}	$\omega_i t$	I_{om}	$\omega_o t + \phi_o$
	20	R_2	$S_{Ba} S_{Cb} S_{Ac}$	$v_B v_C v_A$	$i_c i_a i_b$	V_{im}	$\omega_i t - \frac{2\pi}{3}$	I_{om}	$\omega_o t + \phi_o - \frac{4\pi}{3}$
	21	R_3	$S_{Ca} S_{Ab} S_{Bc}$	$v_C v_A v_B$	$i_b i_c i_a$	V_{im}	$\omega_i t - \frac{4\pi}{3}$	I_{om}	$\omega_o t + \phi_o - \frac{2\pi}{3}$
	22	R_4	$S_{Aa} S_{Cb} S_{Bc}$	$v_A v_C v_B$	$i_a i_c i_b$	V_{im}	$-\omega_i t$	I_{om}	$-\omega_o t - \phi_o$
	23	R_5	$S_{Ba} S_{Ab} S_{Cc}$	$v_B v_A v_C$	$i_b i_a i_c$	V_{im}	$-\omega_i t - \frac{4\pi}{3}$	I_{om}	$-\omega_o t - \phi_o - \frac{4\pi}{3}$
	24	R_6	$S_{Ca} S_{Bb} S_{Ac}$	$v_C v_B v_A$	$i_c i_b i_a$	V_{im}	$-\omega_i t - \frac{2\pi}{3}$	I_{om}	$-\omega_o t - \phi_o - \frac{2\pi}{3}$
III	25	0_1	$S_{Aa} S_{Ab} S_{Ac}$	$v_A v_A v_A$	0	0	0	0-	0-
	26	0_2	$S_{Ba} S_{Bb} S_{Bc}$	$v_B v_B v_B$	0	0	0	0-	0-
	27	0_3	$S_{Ca} S_{Cb} S_{Cc}$	$v_C v_C v_C$	0	0	0	0-	0-

The output reference voltage vector is applied externally such that the instantaneous values of the reference voltage is known and time invariant during the modulation interval. The input current reference angle can be obtained from the supply voltages usually assuming unity pf. The modulation problem is to generate the desired reference vector from the output voltage and input current vectors at every cycle period. For this, the magnitude and angle of the vectors produced by all 27 switching states that will generate the reference vector should be known. Therefore the rotating vectors in group II cannot be used in this modulation approach. The group I (active) and group III (zero) vectors are used to carry out the modulation and synthesize the reference voltage and current vector.

It is of utmost importance to note that it is not possible to control both the magnitude and angle of the output voltage and input current to meet the power balance between the input and output powers as stated in 2.17. Therefore the magnitude of the input current is determined by the power balance. As a result the method aims to control the amplitude and angle of the output voltage and the angle of the input current to generate the desired power factor.

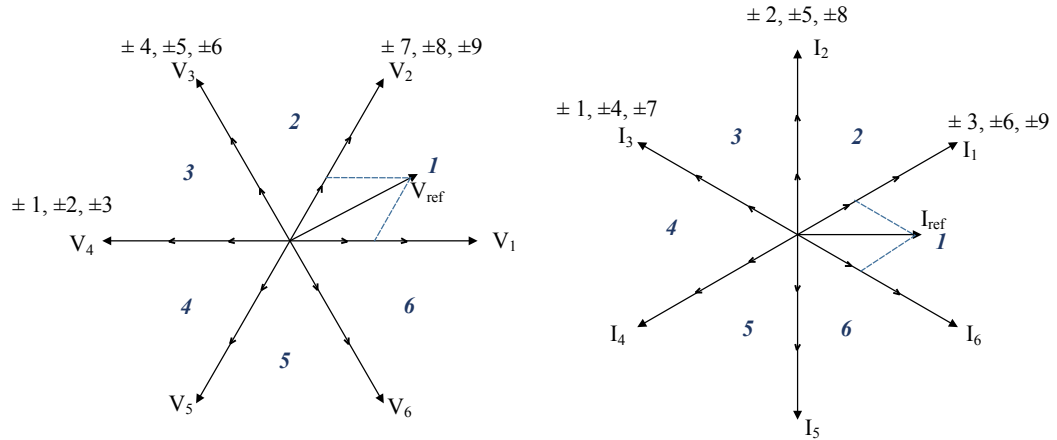


Figure 2.15: Output voltage(left) and input current(right)space vectors used in SVM

The output voltage and input current vectors for group I and III are shown in Figure 2.15. The figure shows the output voltage sector $K_v = 1to6$ and input current sector $K_i = 1to6$ and the vectors that can produce the voltage vectors within each sector. An arbitrary reference voltage $\mathbf{V}_{ref}$ and the resulting input current reference $\mathbf{I}_{ref}$ is

also shown in Figure 2.15.

As seen from Figure 2.15, every available voltage vector can be generated by three switching states. In order to generate V_1 or V_4 , $\pm 1, \pm 2$ or ± 3 can be used. The states 1, 2 and 3 will generate V_1 and -1, -2 and -3 will generate V_4 . Since two active vectors and zero vectors are required to modulate, there will be six switching states available (three for each active vector) to select from. In Figure 2.15 the reference voltage vector is situated in $K_v = 1$ which means V_1 and V_2 can be used to generate V_{ref} . In that case, the possible switching combinations are $\pm 1, \pm 2, \pm 3, \pm 7, \pm 8$ and ± 9 .

However, to generate the input current reference I_{ref} in sector $K_i = 1$, states $\pm 1, \pm 3, \pm 4, \pm 6, \pm 7$ and ± 9 can be considered. So the active vectors that can be used to obtain the reference current are I_1 and I_6 . These four active vectors V_1, V_2, I_1 and I_6 are then used to modulate and generate the desired output voltage. This is done by choosing the switching states that both sets of vectors V_1, V_2 and I_1, I_6 have in common which in this case are $\pm 1, \pm 3, \pm 7$ and ± 9 . The sign of the selected set of states (+ or -) depends on the sign that generates the active vectors, depending on the instantaneous values of the supply voltages (see Table 2.2). Based on these 4 switching states and the zero vector, the output reference voltage vector and angle of the input current reference can be synthesized.

Table 2.2: Selection of switching states in SVM

		K_v		
		$1 \text{ or } 4$	$2 \text{ or } 5$	$3 \text{ or } 6$
K_i	$1 \text{ or } 4$	1, 3, 7, 9	4, 6, 7, 9	1, 3, 4, 6
	$2 \text{ or } 5$	2, 3, 8, 9	5, 6, 8, 9	2, 3, 5, 6
	$3 \text{ or } 6$	1, 2, 7, 8	4, 5, 7, 8	1, 2, 4, 5

Similarly a set of four vectors can be selected if the reference output voltage and the input reference current sectors are known. The set of switching states for different combinations of K_v and K_i are shown in Table 2.2.

For any sampling period the four switching states selected in this manner will produce

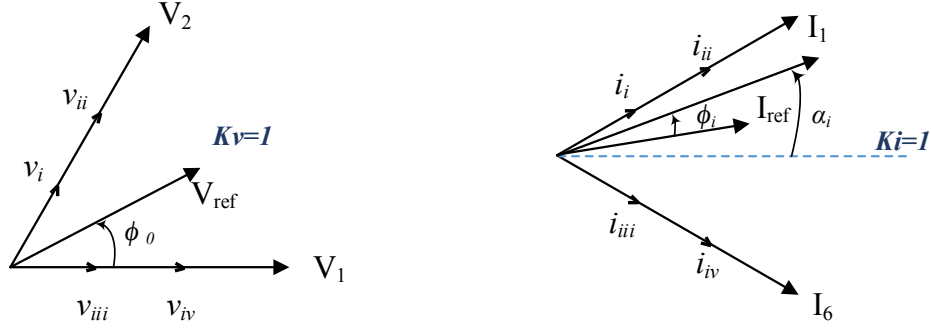


Figure 2.16: Active vector states to produce the target output voltage vector and input current direction for $K_v = 1$ and $K_i = 1$

voltage vectors $\bar{v}_i, \bar{v}_{ii}, \bar{v}_{iii}$ and $\bar{v}_{iv}$ and current vectors $\bar{i}_i, \bar{i}_{ii}, \bar{i}_{iii}$ and $\bar{i}_{iv}$ as shown in Figure 2.16. Φ_o and α_i are the angles of output and input voltages respectively and ϕ_i is the input phase displacement angle. The duty cycles for these four active vectors can then be calculated [33] using equations 2.30 to 2.33.

$$\delta_i = (-1)^{K_v+K_i+1} \frac{2\sqrt{3}}{3} q \frac{\cos(\phi_o - \frac{\pi}{2}) \cos(\phi_i - \frac{\pi}{2})}{\cos(\phi_i)} \quad (2.30)$$

$$\delta_{ii} = (-1)^{K_v+K_i} \frac{2\sqrt{3}}{3} q \frac{\cos(\phi_o - \frac{\pi}{2}) \cos(\phi_i + \frac{\pi}{6})}{\cos(\phi_i)} \quad (2.31)$$

$$\delta_{iii} = (-1)^{K_v+K_i} \frac{2\sqrt{3}}{3} q \frac{\cos(\phi_o + \frac{\pi}{6}) \cos(\phi_i - \frac{\pi}{2})}{\cos(\phi_i)} \quad (2.32)$$

$$\delta_{iv} = (-1)^{K_v+K_i+1} \frac{2\sqrt{3}}{3} q \frac{\cos(\phi_o + \frac{\pi}{6}) \cos(\phi_i + \frac{\pi}{6})}{\cos(\phi_i)} \quad (2.33)$$

where ϕ_o and ϕ_i are the angles of output voltage and input current respectively and q the voltage ratio. Note that the duty cycles will always be positive. A negative duty cycle means that the corresponding negative switching configuration should be selected. The zero vectors are applied for the time remaining after the application of the active vectors and is given by

$$\delta_0 = 1 - (\delta_i + \delta_{ii} + \delta_{iii} + \delta_{iv}) \quad (2.34)$$

The sum of the duty cycles of the active vectors should be less than 1 for each sampling

interval. This results in a maximum voltage transfer ratio, q_{max} for linear operation of MC as shown below.

$$q_{max} = \frac{\sqrt{3}}{2} \cos(\phi_i) \quad (2.35)$$

For unity power factor, the maximum voltage transfer ratio is $\frac{\sqrt{3}}{2}$ or 0.8666 which is the limit that can be attained by the Optimum Venturini method.

Once the duty cycle are calculated the active and zero vectors are arranged in a predefined sequence or pattern. The placement of zero vectors in the sequence affects the ripple content in the MC waveforms and the number of switching transitions/changes in a switching sequence. The sequence is normally defined such that only one output phase undergoes commutation for each change of state. There is no predefined pattern for the placement of vectors to constitute a sequence. One method is to use two zero vectors and four active vectors as shown in Figure 2.17. In this case, there will be ten switching changes within one sequence.

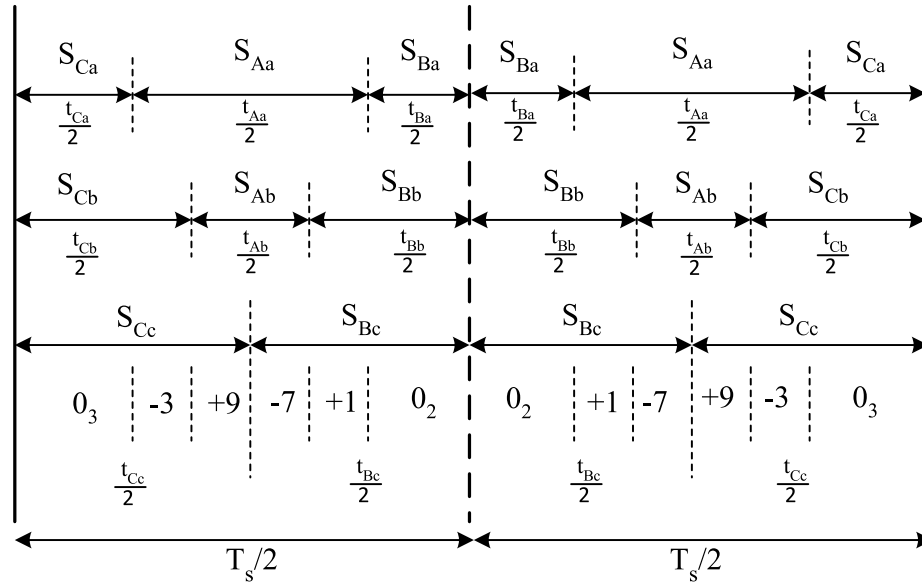


Figure 2.17: SVM sequence with two zeros; for $K_v = 1$ and $K_i = 1$

The most common practice is to use a symmetrical arrangement in which the switching period is divided into two halves and the four active vectors are applied with zero vector in the middle and at both ends [33]. A typical double-sided symmetrical switching pattern is given in Figure 2.18 which uses all the three zero vector with

equal duty cycles. This arrangement will results in twelve switching changes within one modulation sequence.

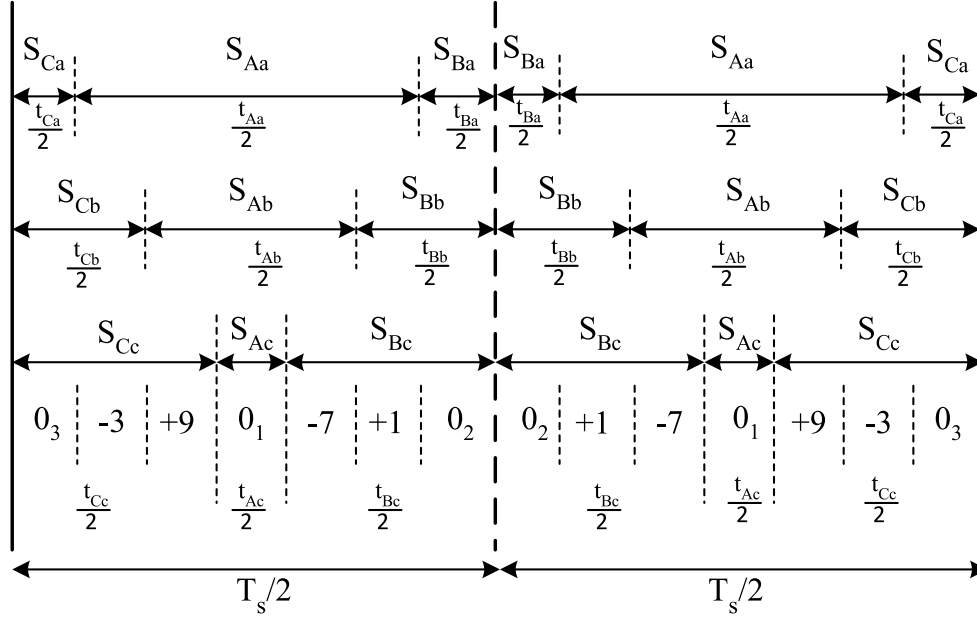


Figure 2.18: Symmetrical sequence for SVM for $K_v = 1$ and $K_i = 1$

2.6 Summary

The basic concept of matrix converter and its operation as a AC-AC converter is discussed in this chapter. A MC lacks a DC link capacitor which makes high power density and high operating temperatures possible. Another feature of this converter is the bidirectional power flow and absence of low frequency harmonics. The input displacement factor can be controlled depending on the modulation method used.

The limited voltage ratio of 86.666% is beneficial in certain applications like the integrated drive. MC is also the ideal candidate for applications where the size of converter is critical. The high temperature endurance is also helpful when the converter is inside the machine as in the case of integrated drives. Compared to other power converters, MC also has a higher number of switches compared to conventional converters which translates to lower stress per device.

SVM is a very popular modulation technique for direct matrix converter as it provides sinusoidal input and output currents. The desired reference voltage vector can be generated by four active vectors and three zero vectors applied for the correct duty cycles. The different methods to arrange the vectors within a modulation sequence was also discussed in this chapter.

Chapter 3

Control of Matrix Converter with Proportional-Integral Controller

3.1 Introduction

Before discussing predictive control strategies, it is worthwhile to study a basic control scheme like the proportional-integral (PI) controllers for the control of a direct matrix converter. Conventional PI controllers can only control one variable when the reference is known. These controllers are also known as SISO (Single input single output) controllers. Therefore in order to control the current and speed of an electrical drive simultaneously, PI controllers in a cascaded structure are required. A PI controller together with the SVM modulation scheme is able to control the matrix converter for different type of loads including induction machines. This chapter focuses on the cascaded closed loop control of a matrix converter feeding a RL load and an induction machine with PI controllers. The design of the current and speed controllers from the system parameters are discussed in detail in this chapter. Simulation results to show the performance of the controller for a matrix converter feeding an RL load and an induction machine are also included in this chapter.

3.2 Control of matrix converter with a RL load

A typical matrix converter system is controlled by PI controllers in conjunction with a modulation strategy like SVM to control the load current. The input current control is inherently achieved by the SVM technique as explained in chapter 2. A control block diagram of the load current control for a matrix converter in rotating reference frame is shown in Figure 3.1.

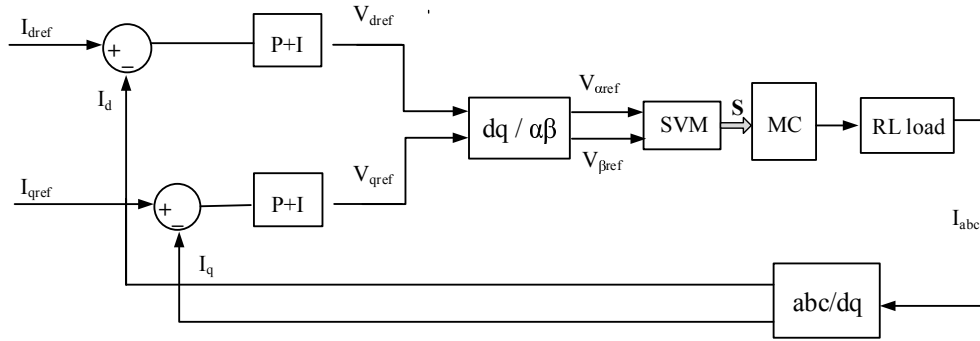


Figure 3.1: Control block diagram of matrix converter with RL load

The 3-phase load currents are converted into components in the rotating reference frame which rotates with the same frequency as that of the supply. The three phase load currents are converted to dq reference frame using Parks transformation (see section 3.3.1). The load currents thus obtained are controlled by two PI controllers to obtain the reference voltages to be imposed on the system. Figure 3.1 depicts the cascaded control structure for load current control of MC. The output of the PI controllers will be dq voltage references which are then converted to alpha-beta reference voltages to generate the PWM signal. The SVM block includes the calculation of duty cycles and selection of switching states to attain the reference voltages.

3.2.1 Plant Model

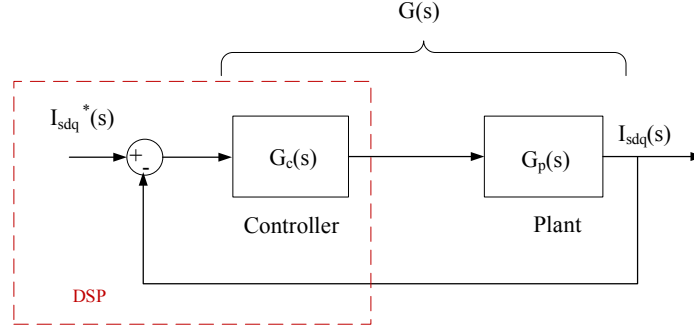


Figure 3.2: PI Controller design block diagram

The block diagram of a typical PI controller design is shown in Figure 3.2. The open loop transfer function of the whole system, $G(s)$ includes the open loop controller transfer function, $G_c(s)$ and the plant transfer function, $G_p(s)$. A typical PI controller transfer function is given by

$$G_c(s) = \frac{k_c(s + a_c)}{s} \quad (3.1)$$

where k_c is the proportional gain and $k_c * a_c$ is the integral gain of the controller. The plant transfer function represents the load behaviour and has a structure as shown in 3.2.

$$G_p(s) = \frac{k_p}{s\tau_p + 1} \quad (3.2)$$

where τ_p is the plant time constant and k_p the plant gain. The time constant is the time taken by the step response of the transfer function to reach 63 % of its steady state value and the plant gain is the steady state value. The plant model for RL load can be derived as shown in 3.3.

$$V_s = R.I(t) + L \frac{dI(t)}{dt} \quad (3.3)$$

The plant transfer function in frequency domain, $G_p(s)$ for the current controller is

then obtained by taking the Laplace transform of 3.3 as shown below.

$$\begin{aligned} V_s &= R.I(s) + sLI(s) \\ G_p(s) &= \frac{I(s)}{V_s} = \frac{1}{R + sL} \end{aligned} \tag{3.4}$$

$$G_p(s) = \frac{1/R}{s\tau + 1}$$

where $\tau = \frac{L}{R}$.

The closed loop transfer function with unity feedback, $G_{cl}(s)$ is given by

$$G_{cl} = \frac{G(s)}{1 + G(s)} \tag{3.5}$$

where $G(s)$ is the forward loop transfer function and $G(s) = G_c(s).G_p(s)$. Once the natural frequency and damping factor are selected, the controller gains k_c and a_c can be calculated from 3.5.

3.2.2 Simulation Results

Simulation of control of a matrix converter with an RL load using PI controllers is done in the Matlab Simulink environment. The control algorithm is implemented in Simulink and the switching model of the converter is implemented using Matlab Plects blockset. The load current of the converter is controlled using PI controllers to produce the voltage references required for the space vector modulation. Since SVM also takes into consideration the angle of the supply currents, the input currents will be controlled. The system parameters used in the simulation are shown in Table 3.1. The controller gain calculations are shown in Appendix.C.

Table 3.1: System Parameters

Parameter	Value	Unit
<i>Load resistance, R</i>	10	Ω
<i>Load inductance, L</i>	3.75	mH
<i>Input filter capacitance, $C_f(\text{delta})$</i>	8.3	μF
<i>Input filter inductance, L_f</i>	0.7	mH
<i>Damping resistor, R_d</i>	10	Ω
<i>Supply frequency</i>	50	Hz
<i>Output frequency</i>	30	Hz
<i>Switching frequency</i>	12.5	kHz

A load current reference of 5A at 30Hz is demanded from the system. The performance of dq-current controllers and reference tracking is depicted in Figure 3.3.

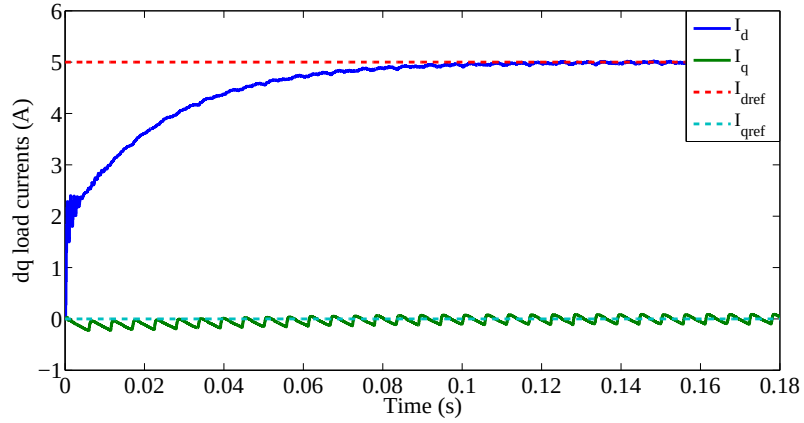


Figure 3.3: dq load currents with references

As seen from the Figure 3.3, the dq-load currents track the references with zero steady state error and no overshoots. The controlled three phase load currents and the corresponding MC output line voltage at steady state is shown in Figure 3.4. THD of the controlled load currents is approximately 1.89%.

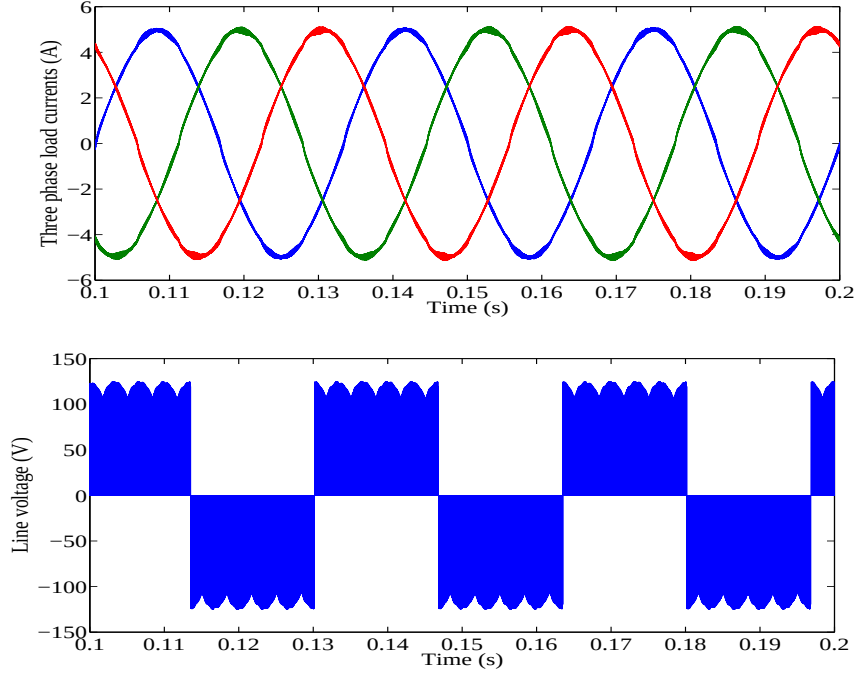


Figure 3.4: (a) Controlled three phase load current of MC (b) MC output line voltage

Since SVM also takes into account the input current angle for duty cycle calculation, the resulting input currents will be sinusoidal. Figure 3.5 shows the three phase supply currents of the MC.

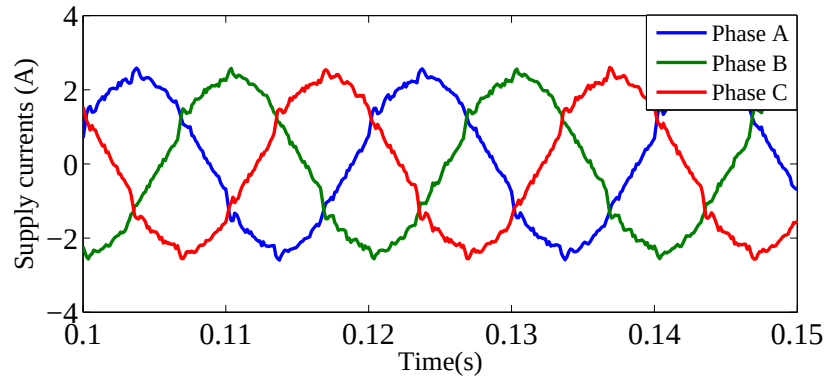


Figure 3.5: Three phase supply currents of the MC

To see the dynamics and performance of a PI controller, a step change in the output reference magnitude and frequency is introduced at $t=0.2s$. The resulting waveforms are shown in Figure 3.6 and Figure 3.7. Figure 3.6 shows the alpha-beta load currents

and their references when the reference magnitude is changed from 2A to 4A. It is evident from the graph that the controller responds to the change in the magnitude and reaches steady state within one cycle without any overshoots. One of the advantages of matrix converter is that it can provide variable frequency output waveforms. The controller performance during a step change in the output reference frequency from 20Hz-40Hz of the current demand is demonstrated by Figure 3.7. The controller responds to the change in frequency without any delay and it can be seen that the currents follows the reference during transient and steady state.

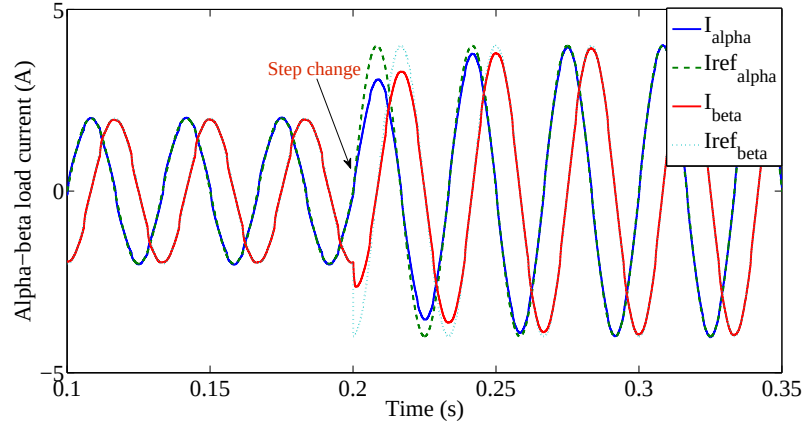


Figure 3.6: Step change in load current from 2A to 4A

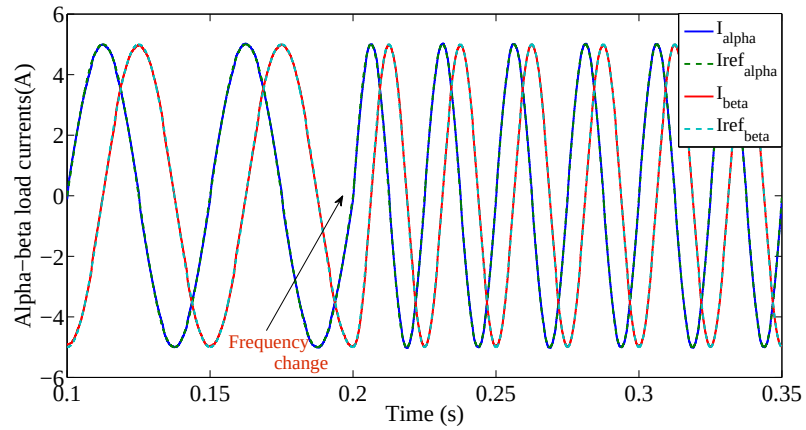


Figure 3.7: Step change in the output frequency of the converter from 20Hz to 40Hz

3.3 Control of a Matrix Converter with an Induction Machine Load

DC drives were initially used for variable speed control as it was able to attain torque and speed response accurately. Torque in DC motor is controlled using the armature and field current. A mechanical commutator with brushes is responsible for field orientation of the motor. However, the brushes and commutators are susceptible to wear and tear thus requiring regular servicing and this reduces the reliability of the DC motor [106].

AC drives are becoming popular for high performance applications as induction motors (IM) are cheaper, low maintenance (no brushes) and has faster speed and torque responses. This is driven by the desire to imitate the performance of DC drive while using AC motors which are easy to maintain.

Induction motors are typically controlled by *scalar control* or *vector control*. In V/f control which is a scalar control method, the desired voltage and frequency are applied to the stator windings and there is no speed feedback. This type of control is suitable for applications like pumps and fans where high levels of accuracy are not needed. The concept of vector control is explained in section 3.3.1.

3.3.1 Vector control of IM

Vector control was developed in 1970 by Blaschke to vary the speed of an induction motor over a wide range [107]. In this scheme the current is synthesized from two quadrature components, one of which is responsible for the flux level in the motor and another which produces torque in the motor.

The vector control strategy is based on two fundamental ideas such as:

Flux and torque producing currents : It is simpler to model an IM using the

flux and torque producing currents rather than considering the classical three phase model of the machine. The direct current, I_{sd} which is the flux producing current and the quadrature current, I_{sq} which is the torque producing current are at right angles with each other. It has to be noted that the actual voltages applied to the motor and the resulting currents are in the three phase system.

Referene frames : The basic idea of reference frame is to transform a sinusoidal quantity in one reference frame to a constant value in another reference frame which is rotating at the same frequency. After the conversion of a sinusoidal quantity to a constant value by carefully choosing the reference frame, it is possible to control that quantity with conventional PI controllers. The most commonly used vector transforms are the Clarke and Park transforms.

1) Clarke transform

A magnitude invariant transformation from three phase system to two orthogonal components is achieved by Clarke transform. The resulting components are referred to as alpha-beta components.

The Clarke transform for the three phase stator currents of an IM is given as

$$\begin{aligned} I_{s\alpha} &= \frac{2}{3}(I_{sa} - \frac{1}{2}(I_{sb} - I_{sc})) \\ I_{s\beta} &= \frac{1}{\sqrt{3}}(I_{sb} - I_{sc}) \end{aligned} \tag{3.6}$$

where I_{sa} I_{sb} I_{sc} are the three phase stator currents.

2)Park transform

Unlike Clarke transform, Park transform involves a vector rotation. It rotates a vector that is defined by its quadrature components through a specific angle.

The alpha-beta currents obtained from 3.6 can be converted to dq currents by using

equation 3.7:

$$\begin{aligned} I_d &= I_{s\alpha} \cos \lambda + I_{s\beta} \sin \lambda \\ I_q &= -I_{s\alpha} \sin \lambda + I_{s\beta} \cos \lambda \end{aligned} \quad (3.7)$$

where λ is angle between the rotating and fixed coordinate system and I_d and I_q are the currents in the rotating reference frame.

The control block diagram of vector control of an IM is shown in Figure 3.8.

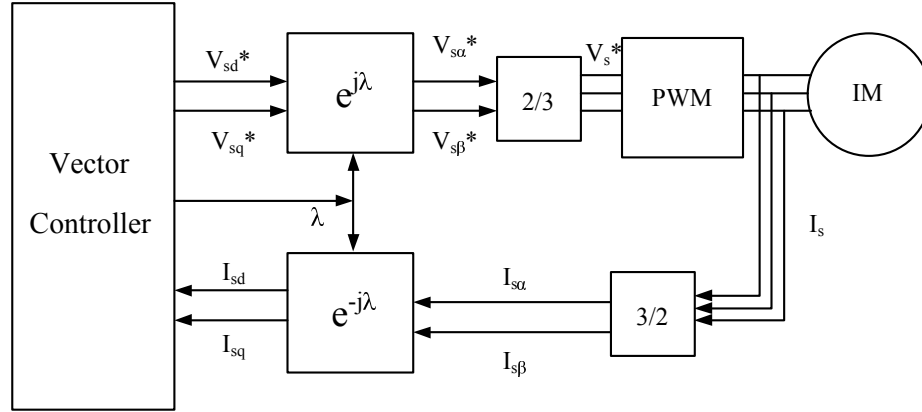


Figure 3.8: Block diagram of vector control

The basic steps involved in vector control of an IM can be summarised as follows:

- Transformation of the measured currents and voltages to dq domain.
- The currents are controlled in dq domain and with dq voltage demands as output of the controller.
- Voltage demands are then inverse transformed into 3-phase demand voltages for PWM.
- The transformations need the angle θ at every point in time

Vector control can be divided into (a) Direct and indirect field orientation control (FOC) and (b) Direct torque and stator flux control [108]. This chapter focuses on field orientation control of an induction machine.

3.3.2 Field Oriented Control

In FOC, mathematical modelling of the machine and the rotor speed measured using an encoder are used to attain field orientation [109]. Even though this requires an encoder which is relatively expensive, the resulting drive will be stable and capable of fast torque response and accurate speed control. Compared to armature and field current in DC drives, the control variables in an AC drive are the voltage and frequency which are controlled by the modulator.

In vector control, space phasors or vectors are used to represent the machine dynamics. The voltage vector V , the current vector I and flux vector Ψ are associated with a stator winding. The instantaneous magnitude of the voltage/current/flux vector is the instantaneous size of the V , I , Ψ in the winding. The direction of the V , I , Ψ is that of the orientation of the winding; or the direction of magnetic motive force (mmf) set up by the current in the winding.

For simplicity, the machine is modelled in dq frame where the dq-axis is aligned with the rotor flux as shown in Figure 3.9. The dq axis exists as a concept for the implementation of the control. I_{sq} is called the torque producing current and I_{sd} is known as the field or magnetizing current. In rotor flux orientation, all the vectors rotate at a speed of $w_e = \frac{d\lambda}{dt}$ with respect to the stator. Therefore the dq axis fixed at the rotor flux will also rotate at w_e . The angle λ is the rotor flux angle.

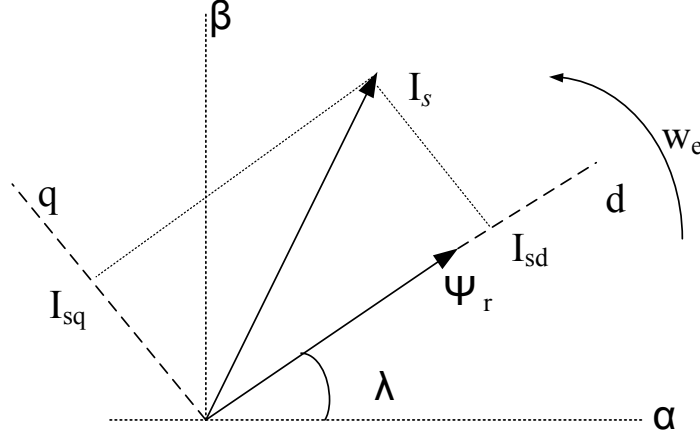


Figure 3.9: dq reference frame

Once the rotor flux angle, λ is obtained, the dq stator currents can be calculated using Park transformation as shown in 3.8.

$$\begin{aligned} I_{sd} &= I_{s\alpha} \cos \lambda + I_{s\beta} \sin \lambda \\ I_{sq} &= -I_{s\alpha} \sin \lambda + I_{s\beta} \cos \lambda \end{aligned} \quad (3.8)$$

Similarly the dq stator currents can be transformed to alpha-beta currents using inverse Park transform as shown in 3.9.

$$\begin{aligned} I_{s\alpha} &= I_{sd} \cos \lambda - I_{sq} \sin \lambda \\ I_{s\beta} &= I_{sd} \sin \lambda + I_{sq} \cos \lambda \end{aligned} \quad (3.9)$$

These transformations can then be applied to voltages and fluxes.

The transformation requires the value of λ at every point in time. It can be calculated using the measured stator voltages and currents (Direct FOC) or can be calculated from the vector controlled constraint equations (Indirect FOC). Direct FOC attempts to directly get the value of machine flux to calculate the transformation angle. As it is difficult to physically measure the flux, an estimation using the stator voltage and current equations are used. The flux estimated this way will not be accurate at lower speeds and hence this method is not reliable. Indirect FOC on the other hand obtains the flux angle by summation of the rotor speed and slip frequency [110].

Indirect vector control being more reliable is mostly used for AC drive applications.

3.3.3 Induction Machine Model

Mathematical model of an IM in a stationary reference frame is explained in Appendix.A. This section briefly describes the modelling of an IM in the dq axis in a synchronous rotating reference frame. If the reference frame is rotating at an angular velocity, ω_k then the stator and rotor voltage vectors of the machine are given by 3.10

$$\begin{aligned} V_{sd} &= R_s I_{sd} + \frac{d\Psi_{sd}}{dt} - \omega_k \Psi_{sq} \\ V_{sq} &= R_s I_{sq} + \frac{d\Psi_{sq}}{dt} + \omega_k \Psi_{sd} \\ V_{rd} &= R_r I_{rd} + \frac{d\Psi_{rd}}{dt} - (\omega_k - \omega_r) \Psi_{rq} \\ V_{rq} &= R_r I_{rq} + \frac{d\Psi_{rq}}{dt} + (\omega_k - \omega_r) \Psi_{rd} \end{aligned} \quad (3.10)$$

where Ψ_{sdq} and Ψ_{rdq} are the stator and rotor flux linkages respectively, ω_r is the electric rotor angular speed. For rotor flux orientation, $\Psi_{rq} = 0$.

The stator and rotor flux linkages depends on the magnetizing inductance and the stator and rotor self inductances and can be expressed as shown in 3.11.

$$\begin{aligned} \Psi_{sd} &= L_s I_{sd} + L_m I_{rd} \\ \Psi_{sq} &= L_s I_{sq} + L_m I_{rq} \\ \Psi_{rd} &= L_r I_{rd} + L_m I_{sd} \\ \Psi_{rq} &= L_r I_{rq} + L_m I_{sq} \end{aligned} \quad (3.11)$$

Figure 3.10 shows the equivalent circuit of induction motor at a synchronously rotating reference frame.

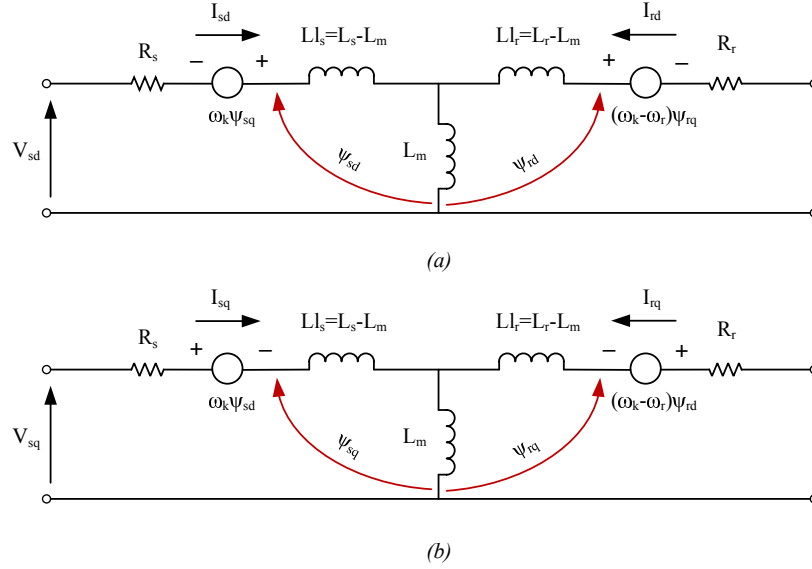


Figure 3.10: dq equivalent circuit at synchronously rotating reference frame (a) d-axis circuit (b) q-axis circuit

Once the mathematical model of the induction machine is derived, the reference frame is selected depending on the control strategy. In this chapter, the control of IM using indirect rotor flux orientation is considered which means the reference frame will rotate at an angular frequency of the rotor flux.

3.3.4 Indirect Rotor Flux Orientation control

Indirect rotor flux oriented control (IRFO) requires the measurement of speed to calculate the rotor flux angle. The rotor flux angle from the rotor speed and slip frequency is estimated using the vector control constraint equations.

The d-axis vector control (constraint) equation is given by

$$\frac{L_r}{R_r} \frac{dI_{mrd}}{dt} + I_{mrd} = I_{sd} \quad (3.12)$$

I_{mrd} is the magnetizing current and can be obtained from I_{sd} using Laplace transform as shown in 3.13. The above equation is obtained by assuming that the magnetizing

current at steady state is constant. Therefore for the field weakening operating range where I_{mrd} will decrease, this relation will not be valid. In that case, a more detailed model of the machine need to be used.

$$\begin{aligned} I_{sd} &= \tau_r s I_{mrd} + I_{mrd} \\ I_{sd} &= (\tau_r s + 1) I_{mrd} \\ I_{mrd} &= \frac{I_{sd}}{(\tau_r s + 1)} \end{aligned} \quad (3.13)$$

The q-axis vector control equation is given by

$$w_{sl} = \frac{I_{sq}}{\tau_r I_{mrd}} \quad (3.14)$$

where $\tau_r = \frac{L_r}{R_r}$ From 3.13 and 3.14 the rotor flux angle, λ can be calculated as follows.

$$\lambda = \int w_e dt = \int (w_r + w_{sl}) = \int \left(w_r + \frac{I_{sq}}{\tau_r I_{mrd}} \right) \quad (3.15)$$

A block diagram showing the calculation of the rotor angle, λ is shown in Figure 3.11.

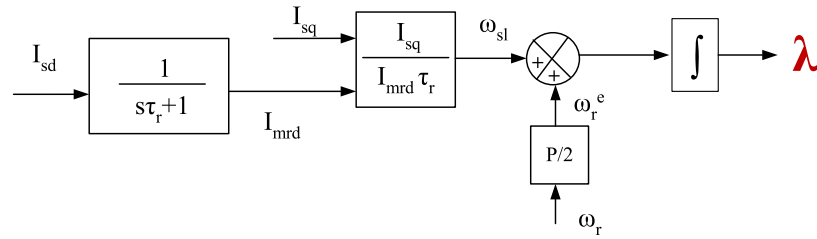


Figure 3.11: Rotor angle calculation

Once the rotor angle is calculated, the stator currents and voltages can be transformed to the synchronous reference frame to design the controller.

3.3.5 Controller Design

A typical structure of the field-oriented control with cascaded PI controllers is shown in Figure 3.12. The inner control loop comprises of current controllers for both field

and torque producing stator currents. The speed controller forms the outer control loop in this structure. The design and implementation of the current and speed control loops are discussed in detail in section 3.3.5.1 and 3.3.5.2. The design procedure to develop the PI controller for speed and current control is explained in Appendix.C.

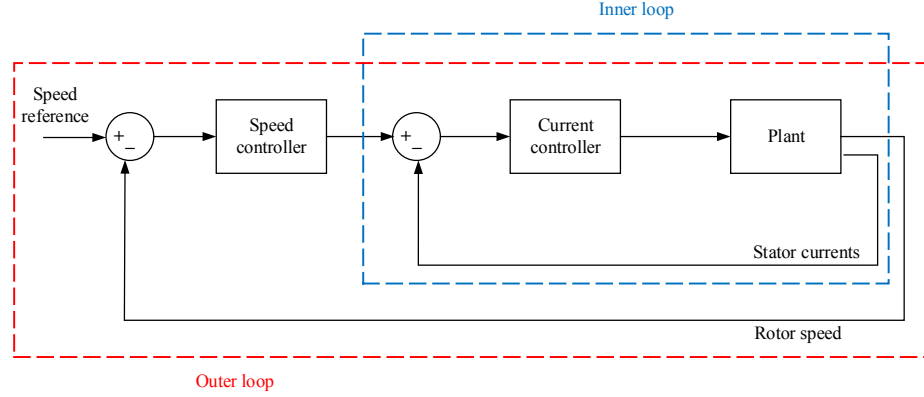


Figure 3.12: Block diagram of a cascaded closed loop control for IM

3.3.5.1 Current Controller

As mentioned in section 3.3.5, the control loops for the field producing current I_{sd} and the torque producing current I_{sq} will be identical which eliminates the need to design two different current controllers. The basic block diagram of a current controller with the converter and the plant is depicted in Figure 3.13.

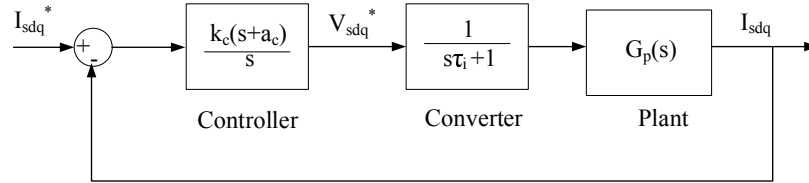


Figure 3.13: Current controller block diagram

To implement the current controller, the plant transfer function, $G_p(s)$ has to be derived. In this case, this is achieved by employing the stator equations in dq frame

given by 3.16.

$$\begin{aligned} V_{sd} &= R_s I_{sd} + \sigma L_s \frac{dI_{sd}}{dt} - \omega_e \sigma L_s I_{sq} + \frac{L_m}{L_r} \frac{d\Psi_{rd}}{dt} \\ V_{sq} &= R_s I_{sq} + \sigma L_s \frac{dI_{sq}}{dt} + \omega_e \sigma L_s I_{sd} + w_e \frac{L_m}{L_r} \Psi_{rd} \end{aligned} \quad (3.16)$$

Considering the first two terms of V_{sd} and V_{sq} of 3.16, the first order plant model for the current controllers can be obtained from 3.17. The last two terms in 3.16 are the voltage compensation terms to overcome the changing flux and back emf of the machine. These are added to the controller as compensation terms.

$$\begin{aligned} I_{sd} &= \frac{V_{sd}}{(s\sigma L_s + R_s)} \\ I_{sq} &= \frac{V_{sq}}{(s\sigma L_s + R_s)} \end{aligned} \quad (3.17)$$

From 3.17 the plant transfer function for both the controllers can be represented as in 3.18.

$$G_p(s) = \frac{1/R_s}{s\tau_s + 1} \quad (3.18)$$

where $\tau_s = \frac{\sigma L_s}{R_s}$ is the stator time constant and is normally in the range of 10-25ms.

Current controllers are typically very fast and the bandwidth is normally in the range of 100-200Hz. When finding the plant transfer function it is good practice to take into account the delays in the system including the converter delay τ_i as shown in Figure 3.13. Usually this is considered to be $1/f_{samp} = 1/12.5k = 80\mu s$.

3.3.5.2 Speed Controller

The machine will develop a torque, T_e which will generate a speed response. This response is defined by

$$T_e = T_{load} + J \frac{dw_r}{dt} + B_v w_r \quad (3.19)$$

T_{load} is the load torque, J is the moment of inertia of the motor and B_v is the frictional coefficient. The design of the outer speed loop requires the approximation of the behaviour of the inner current loop. The plant for the speed controller transfer function can be obtained from 3.19 using Laplace transform as shown in 3.20.

$$w_r = (T_e - T_{load}) \frac{1/J}{s + B_v/J} \quad (3.20)$$

A control block diagram for speed controller is shown in Figure 3.20. The electromagnetic torque can be expressed as :

$$T_e = \frac{3}{2} \frac{P}{2} \frac{L_0^2}{L_r} I_{mrd} I_{sq} = k_t I_{sq} \quad (3.21)$$

where k_t is called the torque constant. Since there is a direct co-relation between the torque and I_{sq} , it can be considered as the torque producing current. So when a torque demand of I_{sq}^* is put on the machine, it will develop the torque promptly due to the fast current controllers. It can be noted that this relation is accurate only during the steady state operation of the machine when the magnetizing current remains constant.

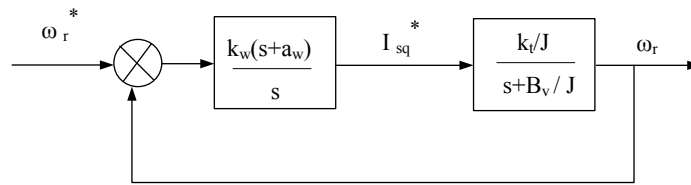


Figure 3.14: Speed controller block diagram

The speed controller forms the outer loop in the cascaded loop structure of a vector controlled drive and hence is much slower than the inner current loops. Normally the bandwidth is in the range of 5-10Hz.

To summarise, a combined block diagram for IRFO including the current and speed controllers and rotor angle calculator is shown in Figure 3.15. This method requires accurate measurement of speed as the rotor angle calculation depends on the speed of the machine at every instant. The dynamics of the inner and outer loops must be different for the correct performance of the control. The inner current loops must

Table 3.2: Parameters of the induction machine

Parameter	Value	Unit
<i>Rated Power</i>	3	<i>kW</i>
<i>Rated RMS voltage(star)</i>	380	<i>V</i>
<i>Rated current</i>	6.3	<i>A</i>
<i>Rated speed</i>	1445	<i>rpm</i>
<i>Stator leakage inductance, Ll_s</i>	10.569	<i>mH</i>
<i>Rotor leakage inductance, Ll_r</i>	7.3916	<i>mH</i>
<i>Mutual inductance, L_0</i>	0.2388	<i>H</i>
<i>Stator resistance, R_s</i>	1.9	Ω
<i>Rotor resistance, R_r</i>	1.33	Ω
<i>Switching frequency, f_{sw}</i>	12.5	<i>kHz</i>
<i>Speedcontroller gain, k_{pw}</i>	0.2	
<i>Speedcontroller gain, k_{iw}</i>	0.6	
<i>Currentcontroller gain, k_{pc}</i>	17.1	
<i>Currentcontroller gain, k_{ic}</i>	2000	

A speed demand of 500rpm is applied to the machine at no load condition and the speed controller converts this into the q-axis current demand for the current controllers to produce the torque required to achieve that speed. The speed PI controllers for current includes saturation and an antiwind-up loop that uses conditional integration. The current is saturated at rated I_{sq} and to eliminate the accumulation of error when saturation occurs, the integrator is turned off. The current PI controllers includes saturation that limits the stator voltages. The performance of the current controller can be seen from Figure 3.16. The currents follow the references without any steady state error during the transient and steady state. The performance of the speed controller together with the three phase stator currents and the electromagnetic torque is shown in Figure 3.17. As seen from Figure 3.17, the speed of the machine reaches steady state without any overshoot. THD of the controlled stator currents is approximately 2.2 %.

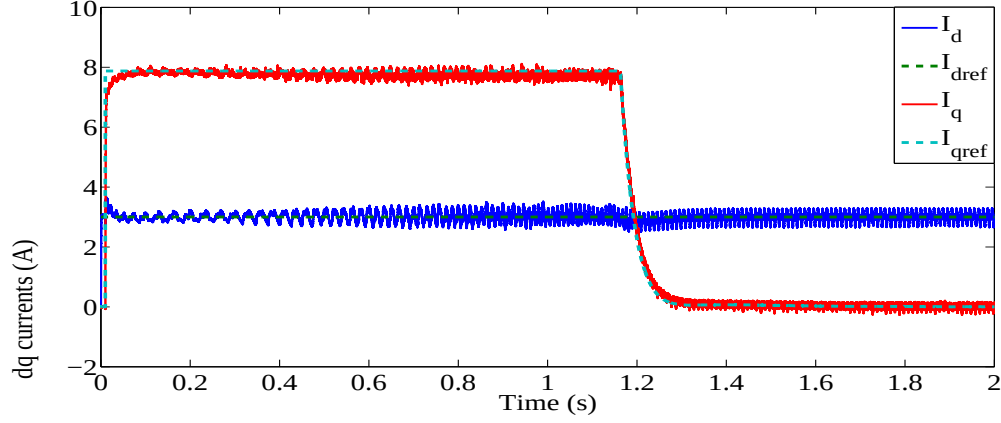


Figure 3.16: Reference tracking of dq currents

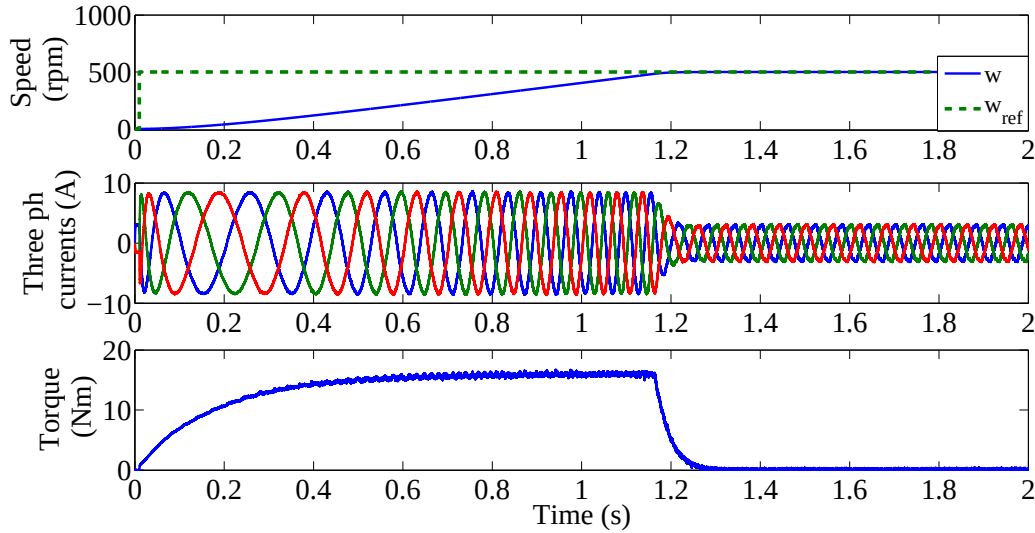


Figure 3.17: Speed, three phase currents and torque of the drive during transient and steady state operation at no load

Figure 3.18 demonstrates the performance of the drive when the speed demand is changed from 200rpm to 500rpm at $t=1.2\text{sec}$. This test is to determine the response of the controller during transients. It can be noted that the torque response at the start of the simulation is slower than the response during a step change. This is due to the fact that the speed demand at the start was applied to the machine before it is completely magnetised. From Figure 3.18, it is evident that the controller responds to the step change in speed with fast transients. The difference in dynamics between

the current and speed controllers is also evident from the above figures.

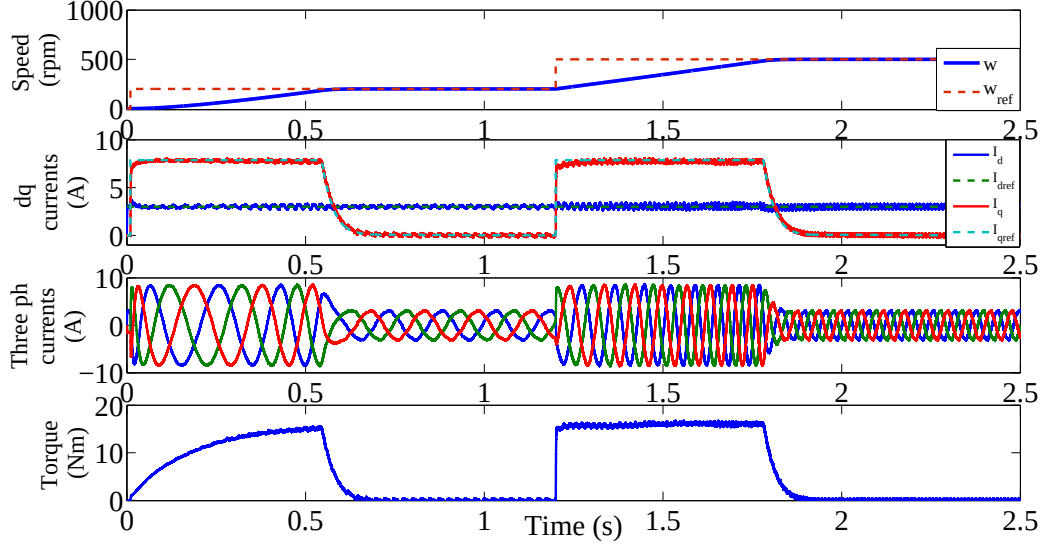


Figure 3.18: Speed, three phase currents and torque of the drive during transient and steady state operation at no load

3.4 Summary

The control of a direct matrix converter using conventional PI controllers together with SVM modulation is discussed in this chapter and the simulation results for the same are also included. It is inferred that in a cascaded control structure, the dynamics of the outer loop will be slower than that of the inner loop and this needs to be considered when selecting the bandwidth. It is not possible to control multiple variables with a PI controller without employing a cascaded structure.

Controllers like MIMO (Multiple input multiple output) are used to control multiple control variables at the same time. In this case, there is no need to have different dynamics between the two controller like in the case of cascaded PI. Predictive control is a MIMO controller and its application in matrix converter will be discussed in chapter 4.

Chapter 4

Model Predictive Control

4.1 Introduction

Predictive control has gained popularity in the recent years in many different engineering applications as well as in power electronics. A number of different control strategies belong to the predictive control family. Researchers have attempted to implement them for power converter control and some of them will be discussed in this chapter. Model Predictive Control (MPC) is one of these strategies and it is the most promising one in power electronics for reasons explained in the next sections. The basic concepts of MPC to control power converters and advantages and disadvantages compared to conventional control strategies are discussed in this chapter. Finally the working principle of MPC and its application on three phase two level inverter and a direct matrix converter is also discussed.

4.2 Predictive control for power converters

The traditional approach to control of power converters is based on linear PID controllers, which use the deviation in control signal from the reference to generate the actuating signal. This type of controller requires information about the plant only at the design stage. With the advancement in digital electronics, the idea to pre-calculate the plant behaviour using mathematical model of the plant and to find out the optimum values for the actuating variables has become popular. This was somewhat the origin of predictive control.

As explained in chapter 3, linear PI controllers in cascaded form that are commonly used in drive control applications suffer from some drawbacks. Predictive control eliminates the need for cascaded control loops since it has the capability of controlling multiple variables simultaneously in a single control loop.

Predictive control represents a family of controllers namely: Hysteresis-based control, Dead-beat control, Trajectory-based control and Model Predictive control as depicted in Figure 4.1. All these strategies will be explained briefly with emphasis on MPC in their discrete implementation in this chapter.

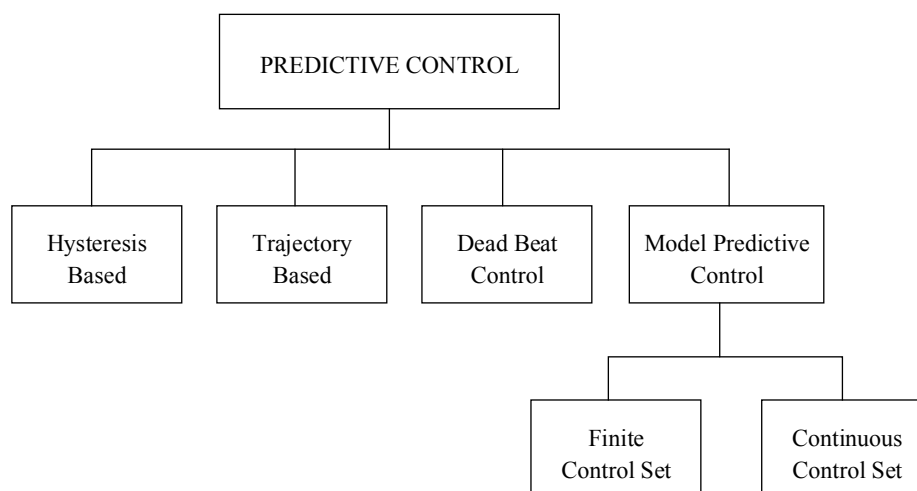


Figure 4.1: Classification of Predictive control

4.2.1 Hysteresis-based control

The objective of hysteresis-based predictive control is to maintain the controlled variables within the bounds of a hysteresis area [111]. An example of this type of controller is the hysteresis or bang-bang controller. Even though they are not explicitly considered to be predictive based controllers, they exhibit their behaviour. The basic idea behind this control can be explained with the help of Figure 4.2. The circle shown in Figure 4.2 is the hysteresis curve/boundary for the control variable, x .

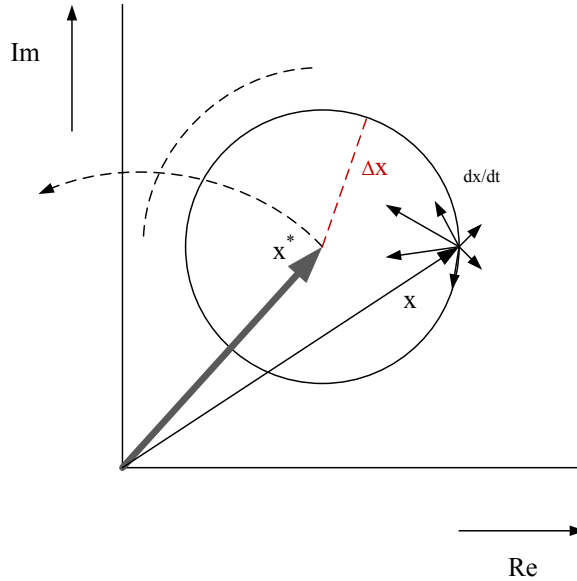


Figure 4.2: Hysteresis-based Predictive controller

The reference control variable vector, x^* dictates the hysteresis boundary which is a circle, the radius, Δx of which is defined by the permissible error band that can be obtained by minimizing the desired cost function [112]. Initially the future trajectories of the control variable vector, x for all the possible switching states are computed. When the control variable reaches the boundary of the curve, the switching state that will keep the control variable within the circle is selected. This results in a variable switching frequency operation. It can be noted that the circular tolerance boundary moves along with the reference control variable vector within the complex plane. This is shown by the dotted line in Figure 4.2.

One of the advantages of hysteresis-based strategies is that an accurate model of the system is not essential. The controller is able to keep the control error within specified limit bands in this case. This is achieved by ensuring the controller reacts quickly if the control variable goes outside the hysteresis boundary/band. Hysteresis controller implemented using micro controllers will not be efficient as the detection of control signal crossing the hysteresis band happens only in the next sampling instant. It is likely that the error has increased to a large value in the next sampling instant. Hence this type of controller is more suitable for analog implementation using operational amplifiers.

To summarise, a hysteresis based predictive controller directly generates the switching signals for the converter without the need for a modulator. As a result, this controlled will result in a variable switching frequency operation.

4.2.2 Trajectory based predictive control

Trajectory based predictive control is based on forcing the control variables to follow a pre-calculated trajectory. Once the system is pushed onto a trajectory, it continues to remain in the trajectory owing to its own properties till an external change is applied. It does not require a modulator in order to generate the switching signals for the converter. This type of controller requires accurate prediction of the future control system behaviour and hence requires an accurate model of the system, unlike hysteresis-based controller. Trajectory-based controllers are suitable for implementation in microprocessors since complex calculations for predictions are required. An advantage of this control strategy is that it does not require cascaded control loops to control more than one variable [111]. The absence of a modulation technique in this case results in variable switching frequency operation as the application time of the vectors is not fixed.

4.2.3 Dead beat control

Dead beat control is another type of control strategy belonging to the predictive control family. It requires a modulator like conventional converter control methods. It involves the calculation of the voltage reference vector for the pulse width modulator in each sampling instant using a model of the system. Then the modulator is used to generate the calculated voltage vector. This has been applied to power converters for a variety of applications like current control [61] and torque control of induction machines [113].

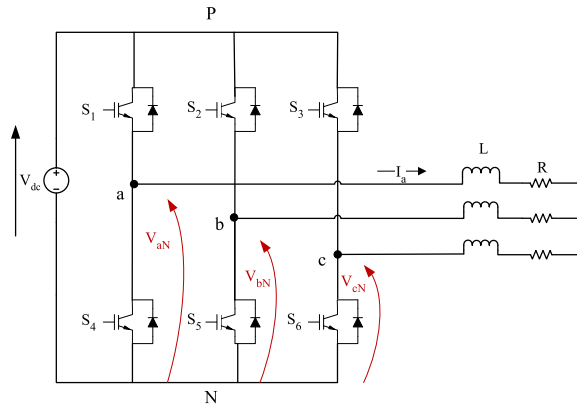


Figure 4.3: Power circuit of a three phase two level voltage source inverter

The basic idea of dead beat control can be explained by considering the example of a three phase two-level inverter shown in Figure 4.3 fed for simplicity by a DC voltage generator. The control block diagram for this particular case is shown in Figure 4.4 and demonstrates the dead-beat method to control the load current.

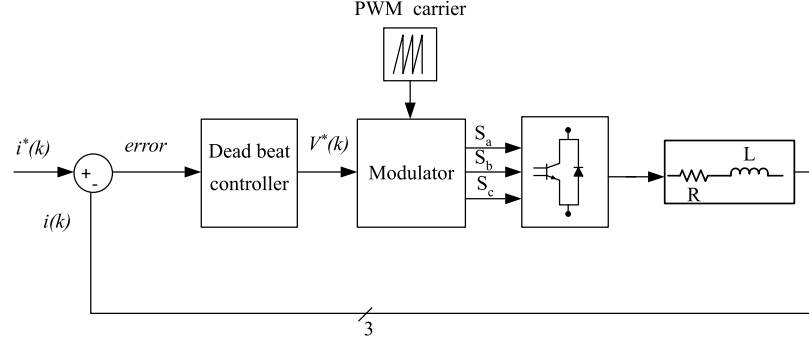


Figure 4.4: Control block diagram for dead beat control

Once a current reference signal is obtained, the load model and measurement of control variables at the current sampling time, k are used to calculate the value of load current in the next sampling interval, $(k+1)$. The difference between the reference value and the current obtained from the system model is the error in the control variable. This error is then used to generate the voltage necessary to produce the current calculated for the next sampling interval. The load model for a generic RL load is given by

$$V(t) = Ri(t) + L \frac{di(t)}{dt} \quad (4.1)$$

where L and R are the load inductance and resistance respectively and $V(t)$ is the converter voltage. To discretise 4.1, first order Euler approximation for the derivative given by 4.2 can be used.

$$\frac{dx}{dt} \approx \frac{x(k+1) - x(k)}{T_s} \quad (4.2)$$

where $x(k+1)$ is the predicted value of the control variable 'x' at instant $(k+1)$, $x(k)$ the value at instant k and T_s the sampling time. Applying 4.2 to 4.1, the voltage required to reach the load current demand is obtained as shown in 4.3.

$$V^*(k) = \left(R - \frac{L}{T_s}\right)i(k) + \frac{L}{T_s}i^*(k+1) \quad (4.3)$$

The reference voltage thus obtained is applied to the converter using a modulator.

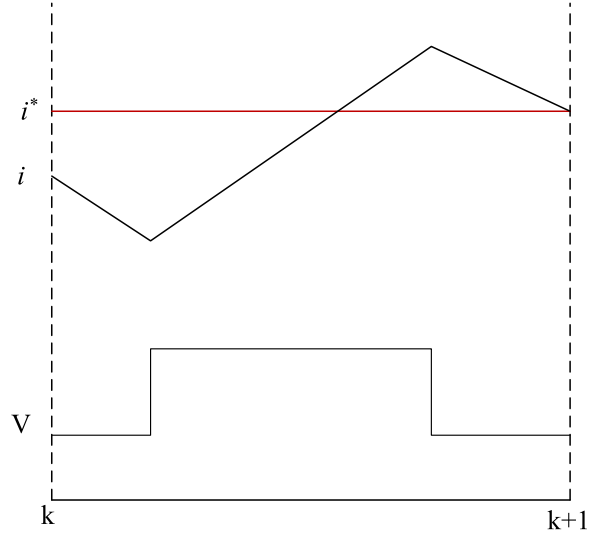


Figure 4.5: Basic principle of dead beat control

Figure 4.5 illustrates the basic principle of dead beat control. The load current i at instant k is not equal to the reference current i^* . The error between these two currents is used to calculate the reference voltage to the modulator in order to fulfil the current demand. In an ideal case, the load current will be equal to the reference current at $(k+1)$. For practical implementation, the voltage is predicted for $(k+1)$ instant to compensate for the computation delay.

The dead-beat control provides fast transient response compared to conventional PI controllers. The modulator makes sure the switching frequency is fixed which minimizes lower order harmonics. Even though this provides fast dynamic response, this method is vulnerable to performance deterioration and even instability if the model is not accurate or in the case of parameter mismatch. It is also not possible to achieve multi-objective control without the use of cascaded loops in this controller implementation.

4.2.4 Model Predictive Control

Model Predictive Control (MPC) was introduced in the late seventies [47] and in recent years it has been applied to power converters and has been used for different applications like current control [52], power control [114] for a VSI and current and torque control for induction motors and permanent magnet machines [55], [56], [57]. Due to the fast sampling times used in the control of power converters, solving the optimization problem of MPC online is not practical. One approach is to use an explicit solution of MPC, solving the optimization problem offline. The resulting controller is a search tree or a lookup table and can be implemented without big computational effort. Considering that power converters are systems with finite number of states, given by the possible combinations of the state of the switching devices, the MPC optimization problem can be simplified and reduced to the prediction of the behaviour of the system for each possible state. Then, each prediction is evaluated using the cost function and the state that minimizes it, is selected.

Hysteresis-based and trajectory-based controllers makes use of the present state variables to calculate the value of the control variables in the next sampling instant. These controllers do not use past variable states for the calculations. Model-based predictive control methods consider the value of control variable in the previous sampling instant to optimize the future values of control/actuating variables, not just for the next sampling instant, but up to a specified control horizon.

MPC strategy can be divided into two categories namely continuous and finite set. Continuous time MPC requires a modulator to control the converter as it generates continuous control signal whereas finite set MPC selects the control action among a finite number of switching states directly applied by the converter. It relies on the switching model of the power converter and has been widely used to control power converters. The focus of this chapter will be on finite set model predictive control (FS-MPC). Table 4.1 summarises the different predictive control methods and their characteristics.

Table 4.1: Predictive control methods in power electronics

Predictive control methods	Features
Deadbeat	Fixed switching frequency Uses Modulator Requires cascaded loops for multi-objective control
Hysteresis-based Control	Variable switching frequency No modulator
Trajectory-based	Variable switching frequency No modulator Does not require cascaded loops
Continuous-time MPC	Fixed switching frequency Requires a modulator Can include constraints Not easy to implement
Finite-set MPC	Variable switching frequency Does not require modulator Can include constraints Easy to implement Online optimization

4.3 FS-MPC-Overview

FS-MPC is an optimization problem where a model of the system is considered in order to predict its future behaviour over a horizon in time and a sequence of future actuations is then obtained by minimizing a cost function representing the desired behaviour of the system. The first element of the sequence is applied and all the calculation is repeated every sample period. The characteristics of power converters and modern day control systems is shown in Figure 4.6.

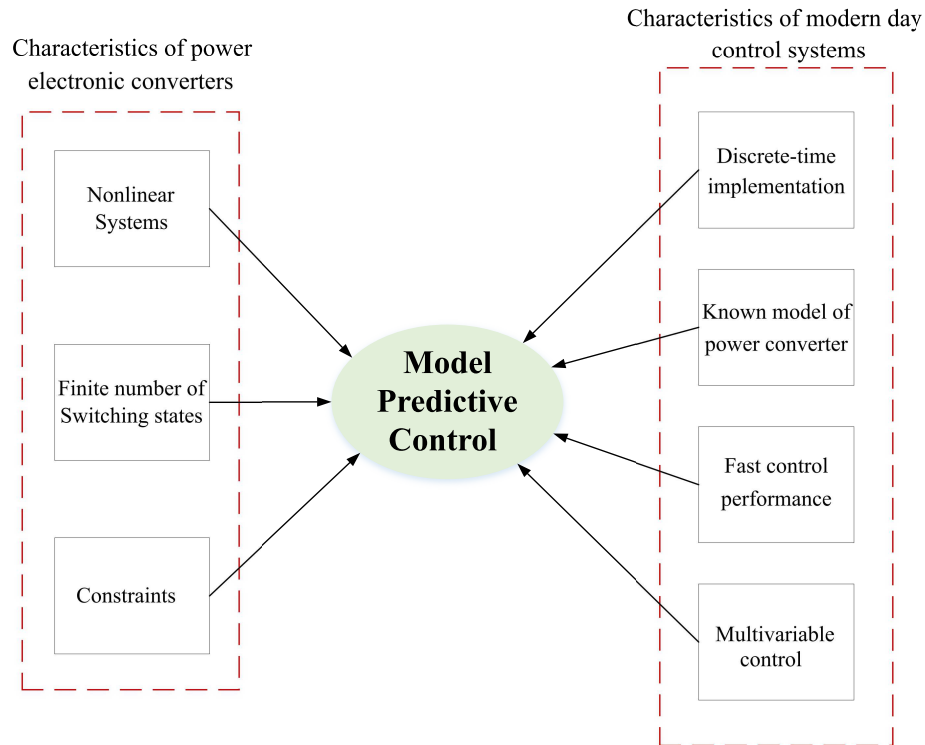


Figure 4.6: Model Predictive control for power converters

The basic idea of model predictive control strategy involves the following key steps:

- The explicit use of the model of the system to predict the output at future time instants (horizon).
- Calculation of a control sequence by minimizing a cost function.
- Receding horizon control, this means that at each instant the horizon is moved forwards and the first control signal of the sequence calculated at each step is applied at that instant.

MPC algorithms are used in a variety of applications including robotics, process industries in chemical plants, oil refineries, power system balancing and so on. A block diagram of the MPC scheme is shown in Figure 4.7.

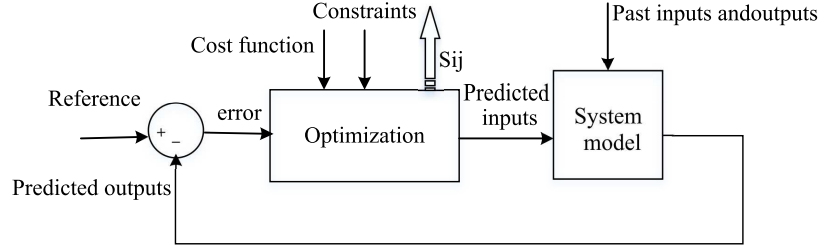


Figure 4.7: Model Predictive control scheme block diagram

MPC presents a series of advantages such as:

- Knowing the system model gives the control sequence.
- A variety of processes ranging from simple to complex dynamics can be controlled.
- Can obtain very fast transient response.
- Easy to control a multi-variable system.
- Easy to implement.
- Non-linearities can be easily included.
- Can include various constraints in the cost function as the application demands without any complex steps.
- It is possible to modify and extend the methodology depending on specific applications.

MPC has high computation demands due to the calculation of control variables for future sampling instants. The number of predictions made is referred to as the prediction horizon, N_p . By considering a limited number of future sampling instants called a control horizon, N_c the calculation complexity can be significantly reduced. After N_c instants, it is assumed that the system has reached steady state and hence the controller output remains constant. This is illustrated in Figure 4.8.

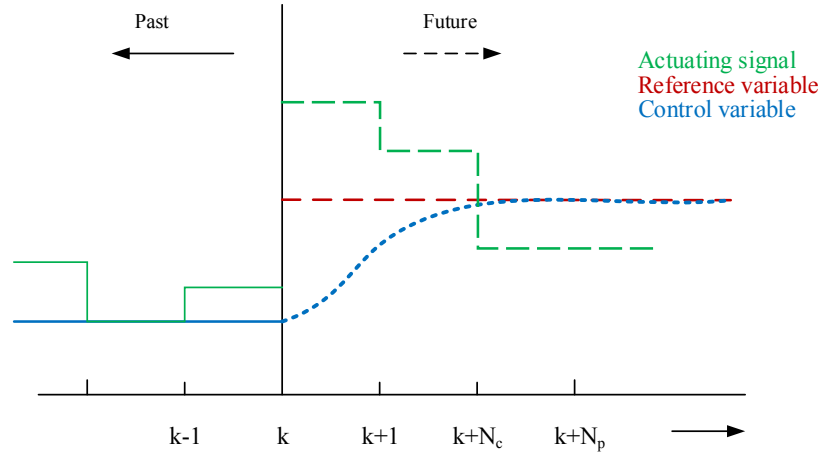


Figure 4.8: Control and prediction horizon in MPC

Like any other control strategy, MPC also has its drawbacks. Although the control law is easy to implement, the derivation of the control equations is more complex compared to classical PID controllers. Also the quality of the model of the system has a direct influence on the response of the controller. If the dynamics of the process does not change, the derivation of the controller can be done beforehand. However if the system changes in time, then some adaptation or estimation algorithm is required.

Since the application of vectors depends entirely on the cost function minimization algorithm, it is possible that the same switching state gives the minimum error for multiple sampling periods. Since the optimized switching state is applied over the entire sampling interval at all times, it is likely that the optimum switching state remains the same for more than one sampling period. This generates large ripples in the waveforms resulting in variable switching frequency. MPC also results in output waveforms with low order harmonics when compared to conventional control methods where the prominent harmonics are in the range of switching frequency and its multiples.

4.3.1 Control strategy

Finite-set MPC is an optimization problem where a cost function is minimized for a predefined interval of time also known as the control horizon. The model of the system that is used to predict the system behaviour is crucial to this control method. It is advantageous to obtain the system model and controller design in discrete-time rather than realizing continuous-time models in digital controllers. Hence the system model and controller description in this case is in discrete-time domain. A discrete-time representation of a system can be obtained from the *state space model*. A linear system can be represented in discrete time state space description as shown in equation 4.4.

$$\begin{aligned}\vec{x}(k+1) &= \mathbf{A}\vec{x}(k) + \mathbf{B}\vec{u}(k) \\ \vec{y}(k) &= \mathbf{C}\vec{x}(k) + \mathbf{D}\vec{u}(k)\end{aligned}\tag{4.4}$$

where $\mathbf{A}$ is the state matrix, $\mathbf{B}$ is the input matrix, $\mathbf{C}$ is the output matrix and $\mathbf{D}$ the feed-forward matrix. $\vec{x}(k)$ represents the state vector, $\vec{u}(k)$ the input vector and $\vec{y}(k)$ the output vector at instant k . $\vec{x}(k+1)$ is the predicted value of the state vector.

Once the model of the system is derived and predictions of control variables are made, a cost function is designed to obtain the desired behaviour of the system. This function mainly comprises of the references and future errors. A typical cost function will have the following format.

$$G = f(x(k), u(k), \dots, u(k + N_p))\tag{4.5}$$

where N_p is called the prediction horizon which is the number of sampling intervals considered.

The optimization problem is to find the minimum value of the cost function and the switching state that will give the minimum error. The entire process is then repeated at the next sampling instant.

4.4 Predictive current control of a Voltage Source Inverter

As a first step for the application of MPC, load current control of a three phase, two-level inverter is considered. This is also the most commonly used inverter for motor drive applications and therefore represents a valid practical example. The MPC strategy to control the load current for a two level inverter includes prediction of current errors for all the eight switching states, formulating a cost function describing the current error and then optimizing the cost function to get the voltage vector that will produce the minimum error in the next sampling instant.

The power circuit for a three phase, two-level inverter feeding an RL load is shown in Figure 4.9. As seen in the Figure 4.9, the converter has six switches and three switching states to control them. Two switches on each inverter phase operate in a complementary mode in order to avoid short circuiting the DC source. The power switches S_1 to S_6 can be connected in a particular manner which can be represented by the switching states S_a, S_b and S_c . V_{aN}, V_{bN} and V_{cN} are the phase neutral voltages at the output.

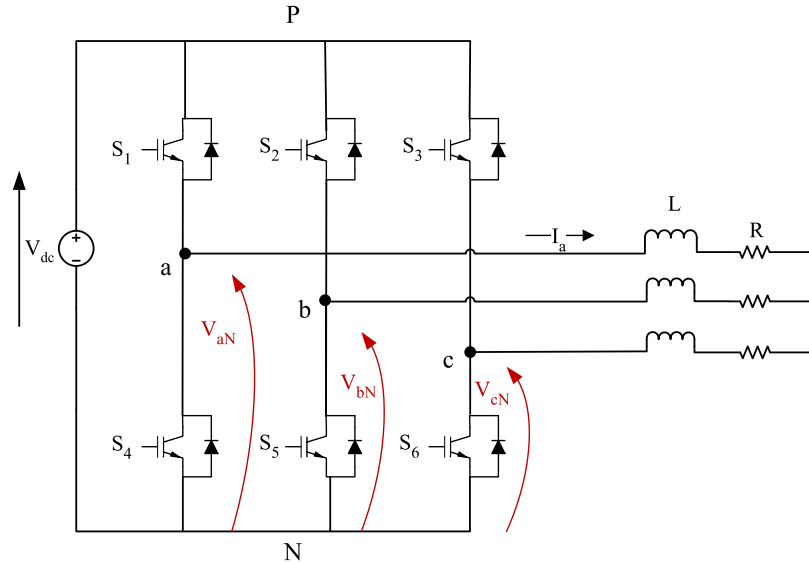


Figure 4.9: Power circuit of a 3-phase voltage source inverter

The switching states of the converter are defined as shown in 4.6-4.8.

$$S_a = \begin{cases} 1 & S_1 \text{ is ON and } S_4 \text{ is OFF} \\ 0 & S_1 \text{ is OFF and } S_4 \text{ is ON} \end{cases} \quad (4.6)$$

$$S_b = \begin{cases} 1 & S_2 \text{ is ON and } S_5 \text{ is OFF} \\ 0 & S_2 \text{ is OFF and } S_5 \text{ is ON} \end{cases} \quad (4.7)$$

$$S_c = \begin{cases} 1 & S_3 \text{ is ON and } S_6 \text{ is OFF} \\ 0 & S_3 \text{ is OFF and } S_6 \text{ is ON} \end{cases} \quad (4.8)$$

Different combinations of these switching states will determine the output voltages. The load voltage in terms of the input DC source voltage and the switching state is given by 4.9.

$$\begin{aligned} V_{aN} &= S_a V_{dc} \\ V_{bN} &= S_b V_{dc} \\ V_{cN} &= S_c V_{dc} \end{aligned} \quad (4.9)$$

where V_{dc} is the DC source voltage.

For unity power factor, output voltage space vector can be defined as in 4.10.

$$\mathbf{v} = \frac{2}{3}(V_{aN} + aV_{bN} + a^2V_{cN}) \quad (4.10)$$

where $a = e^{j2\pi/3}$.

The three switching states can be connected in eight different combinations to produce different voltage vectors. Out of the eight, six of the combinations will yield a non-zero/active voltage vector and the rest of the options result in zero vectors. The active

vectors represented by V_1 to V_6 will vary in magnitude and direction and $V_0 = V_7 = 0$.

The voltage vectors for all switching combinations can be calculated as follows:

For a zero vector where $(S_a, S_b, S_c) = (0, 0, 0)$,

$$V_0 = \frac{2}{3}(0 + a0 + a^20) \quad (4.11)$$

For an active vector, for example where $(S_a, S_b, S_c) = (1, 0, 0)$,

$$V_1 = \frac{2}{3}(V_{dc} + a0 + a^20) \quad (4.12)$$

Similarly the voltage vectors for all the eight combinations can be calculated and are shown in Table 4.2.

Table 4.2: Switching states and voltage vectors for a VSI

S_a	S_b	S_c	Voltage Vector $\mathbf{V}$
0	0	0	$V_0 = 0$
1	0	0	$V_1 = \frac{2}{3}V_{dc}$
1	1	0	$V_2 = \frac{1}{3}V_{dc} + j\frac{1}{\sqrt{3}}V_{dc}$
0	1	0	$V_3 = -\frac{1}{3}V_{dc} + j\frac{1}{\sqrt{3}}V_{dc}$
0	1	1	$V_4 = -\frac{2}{3}V_{dc}$
0	0	1	$V_5 = -\frac{1}{3}V_{dc} - j\frac{1}{\sqrt{3}}V_{dc}$
1	0	1	$V_6 = \frac{1}{3}V_{dc} - j\frac{1}{\sqrt{3}}V_{dc}$
1	1	1	$V_7 = 0$

The voltage vectors thus attained can be represented in the complex plane as shown in Figure 4.10.

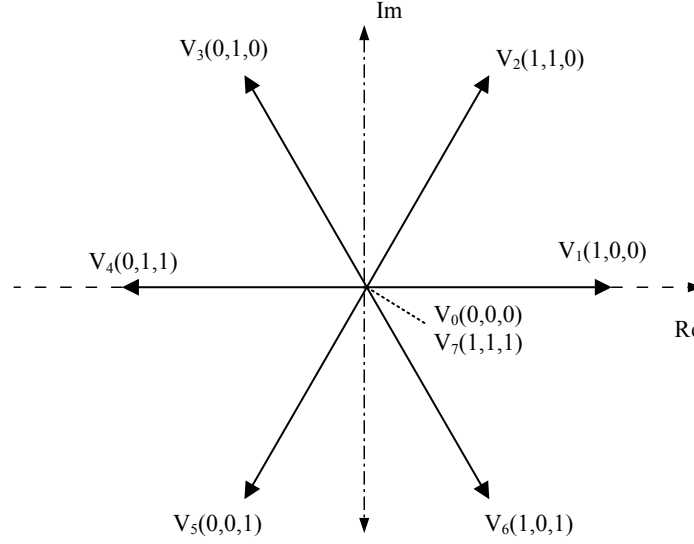


Figure 4.10: Voltage vectors generated by a 3-ph VSI

Once the switching states of the converter are known, predictive controller requires the load model in order to compute the load currents at $k+1$ instant.

From Figure 4.9, the equations for the load current dynamics for the three phases can be written as

$$V_{jN}(t) = L \frac{dI_j(t)}{dt} + RI_j(t) \quad (4.13)$$

where $j=a,b,c$. The three phase equations are then converted to alpha-beta reference frame to simplify the calculations. Discretising 4.13 in alpha-beta using forward Euler approximation (see 4.2) the following equations are obtained.

$$\begin{aligned} I_\alpha(k+1) &= \left(1 - \frac{RT_s}{L}\right) I_\alpha + \frac{T_s}{L} V_\alpha \\ I_\beta(k+1) &= \left(1 - \frac{RT_s}{L}\right) I_\beta + \frac{T_s}{L} V_\beta \end{aligned} \quad (4.14)$$

The predicted values of alpha-beta currents are then used to construct the cost function. The predictive control algorithm then chooses the optimum switching state which gives the minimum value of cost function.

To explain the working principle of MPC strategy, a detailed example is shown in Figure 4.11. Figure 4.11 shows the reference, present instant and future alpha load

currents and how the predictive controller chooses the correct switching state to minimize the error between the reference and predicted current. The predicted values of current $I_\alpha(k+1)$ are obtained from the measured current $I_\alpha(k)$ and the converter voltage. In Figure 4.11, $I_\alpha(k+1), V_{(0,7)}$ are produced by the voltage vectors V_0 and V_7 . It is evident from the figure that the current produced by V_2 and V_6 are closer to the reference value than any other voltage vectors which in turn will have the minimum error. The same procedure is done for beta load current to obtain the final cost function.

Figure 4.11 essentially shows the meaning of cost function as a measure of the error or distance between the reference and predicted currents. The cost function therefore is merely the error signal and can be represented as

$$G = | I_{\alpha ref}(k+1) - I_\alpha(k+1) | + | I_{\beta ref}(k+1) - I_\beta(k+1) | \quad (4.15)$$

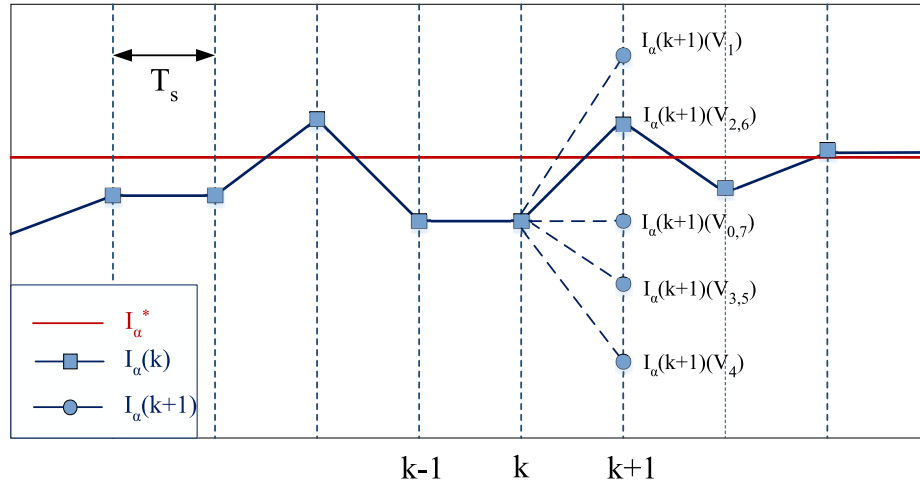


Figure 4.11: Working principle of MPC:Reference and predicted currents

The predictive control algorithm processes the cost function in 4.15 for all the available switching states and arrives at the optimum one based on the error minimization principle. The optimum switching state is then applied to the converter.

4.5 Model Predictive control of a Matrix Converter

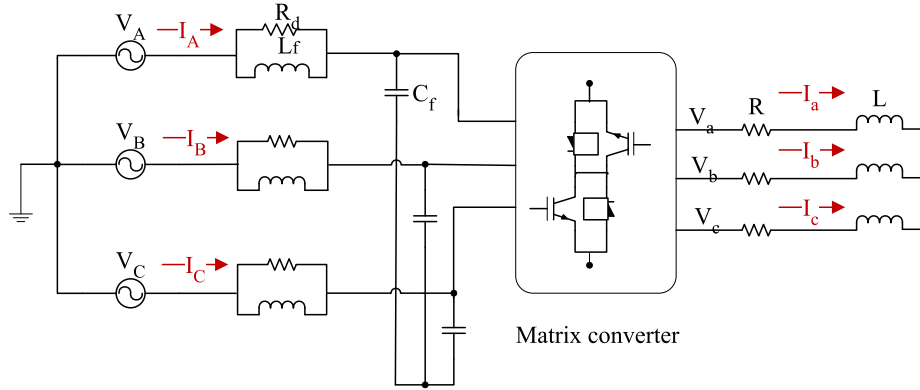


Figure 4.12: Schematic of matrix converter feeding an RL load

Control of a direct matrix converter using MPC can be achieved in the same way as a two-level inverter. However, for matrix converter the computational effort is more than that of a two-level inverter due to the presence of 27 switching states which corresponds to 27 iterations of the cost function minimization algorithm. MPC strategy is applied to a matrix converter to control the load current with different loads like RL load and induction machine. A schematic of a direct matrix converter feeding an RL load is shown in Figure 4.12. The control strategy where the load currents are controlled is also known as the predictive current control (PCC). To demonstrate the multi-objective control feature of MPC, control of load current and instantaneous input reactive power will also be discussed in this chapter.

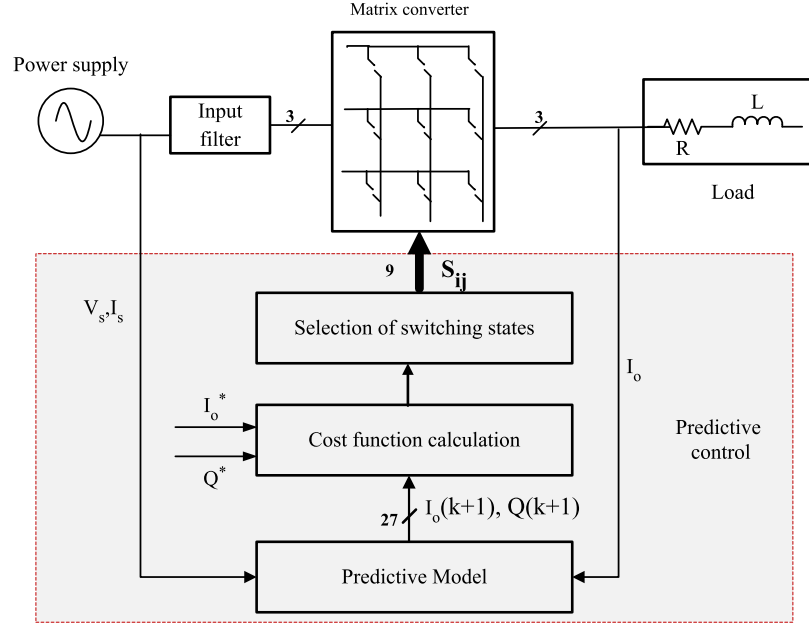


Figure 4.13: Control block diagram for PCC of matrix converter

The control block diagram of MPC applied to a direct matrix converter feeding an RL load in Figure 4.13 shows the different stages involved in this strategy. The basic steps for the MPC strategy applied to MC are determination of the system model (converter and load), prediction of control variables and Cost function calculation and optimization. These steps are discussed in detail in the following sections.

4.5.1 Converter model

The output voltages and input currents of the matrix converter can be obtained from the circuit schematic in Figure 4.12 as shown in 4.16.

$$\begin{bmatrix} v_a(t) \\ v_b(t) \\ v_c(t) \end{bmatrix} = \begin{bmatrix} m_{Aa} & m_{Ba} & m_{Ca} \\ m_{Ab} & m_{Bb} & m_{Cb} \\ m_{Ac} & m_{Bc} & m_{Cc} \end{bmatrix} \cdot \begin{bmatrix} v_A(t) \\ v_B(t) \\ v_C(t) \end{bmatrix} \quad (4.16)$$

where $v_a(t)$, $v_b(t)$ and $v_c(t)$ are the converter output voltages and $v_A(t)$, $v_B(t)$ and $v_C(t)$ the three phase supply voltages. Similarly the input currents of the converter can be

calculated from the output currents using equation 4.17.

$$\begin{bmatrix} i_A(t) \\ i_B(t) \\ i_C(t) \end{bmatrix} = \begin{bmatrix} m_{Aa} & m_{Ba} & m_{Ca} \\ m_{Ab} & m_{Bb} & m_{Cb} \\ m_{Ac} & m_{Bc} & m_{Cc} \end{bmatrix}^T \cdot \begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} \quad (4.17)$$

where $i_a(t)$, $i_b(t)$ and $i_c(t)$ are the three phase load currents and $i_A(t)$, $i_B(t)$ and $i_C(t)$ the supply currents. m_{ij} for $i=A,B,C$ and $j=a,b,c$ are the modulation indices where $m_{ij} = \frac{t_{ij}}{T_s}$. T_s is the sampling time and t_{ij} is the time for which the switch is ON.

If the control of the input reactive power or supply current is required, the input filter model should also be considered. For example, for phase A the schematic of the input filter is shown in Figure 4.14.

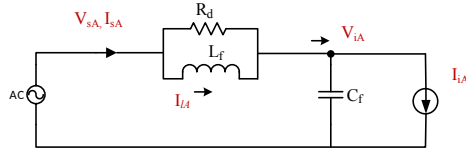


Figure 4.14: Input filter schematic for Phase A

An LC filter with a damping resistor is considered in this case. From Figure 4.14 the continuous time state space model of the filter can be derived and is given by 4.18.

$$\dot{x}(t) = \underbrace{\begin{bmatrix} -\frac{1}{R_d C_f} & \frac{1}{C_f} \\ -\frac{1}{L_f} & 0 \end{bmatrix}}_{A_c} x(t) + \underbrace{\begin{bmatrix} \frac{1}{R_d C_f} & -\frac{1}{C_f} \\ -\frac{1}{L_f} & 0 \end{bmatrix}}_{B_c} u(t) \quad (4.18)$$

where $x(t) = \begin{bmatrix} V_{iA}(t) \\ I_{iA}(t) \end{bmatrix}$ and $u(t) = \begin{bmatrix} V_{sA}(t) \\ I_{sA}(t) \end{bmatrix}$

Discretising 4.18 by applying forward Euler's approximation which is commonly used

in predictive control methods, results in 4.19.

$$\begin{aligned} V_{iA}(k+1) &= T_s \left(V_{iA}(k) \left(\frac{1}{T_s} - \frac{1}{R_d C_f} \right) + \frac{1}{R_d C_f} V_{sA}(k) + \frac{1}{C_f} (I_{lA}(k) - I_{iA}(k)) \right) \\ I_{lA}(k+1) &= I_{lA}(k) + \frac{T_s}{C_f} (V_{iA}(k) + I_{iA}(k)) \end{aligned} \quad (4.19)$$

From 4.19 , $I_s(k+1)$ can be calculated as shown below.

$$I_{sA}(k+1) = I_{lA}(k+1) + \frac{V_{sA}(k+1) - V_{iA}(k+1)}{R_d} \quad (4.20)$$

The voltage after the input filter, V_{iA} is obtained from the voltage transducers. It can be noted that input current control requires at least two additional measurement sensors.

4.5.2 Load model

4.5.2.1 RL load

To determine the load current in the next sampling interval, a mathematical model of the load is required. The continuous time model of a resistive-inductive load is given by

$$L \frac{di_o(t)}{dt} = v_o(t) - R i_o(t) \quad (4.21)$$

where R and L are the load resistance and inductance respectively, $i_o(t) = [i_a(t) \ i_b(t) \ i_c(t)]^T$ the load currents and $v_o(t) = [v_a(t) \ v_b(t) \ v_c(t)]^T$ the matrix converter output voltage.

Discretising 4.21, using forward Euler approximation (see 4.2) ,the discrete time model of the load can be obtained as shown in 4.22.

$$I_o(k+1) = \left(1 - \frac{RT_s}{L}\right) I_o(k) + \frac{T_s}{L} V_o(k) \quad (4.22)$$

4.5.2.2 Induction Machine

For an induction motor(IM), the stator current control is based on model of the machine [106] in a field oriented reference frame. The stator equations of an IM in this reference frame are given in 4.23.

$$\begin{aligned} V_{sd} &= I_{sd}R_s + \frac{d\Psi_{sd}}{dt} - \omega_e \Psi_{sq} \\ V_{sq} &= I_{sq}R_s + \frac{d\Psi_{sq}}{dt} + \omega_e \Psi_{sd} \end{aligned} \quad (4.23)$$

where V_{sd} and V_{sq} are the stator voltages, Ψ_{sd} and Ψ_{sq} the stator fluxes, I_{sd} and I_{sq} the stator currents in the dq co-ordinate frame and ω_e is the speed of the rotor flux. Discretising the stator equations in 4.23 by applying forward Euler approximation, the stator fluxes can be estimated. The stator currents can then be predicted in order to form the cost function. The current prediction is also done in the dq reference frame since a field oriented model is used.

$$\begin{aligned} I_{sd}(k+1) &= I_{sd}(k) + \epsilon(V_{sd}(k) - R_s I_{sd} + \sigma L_s \omega_e I_{sq}(k)) \\ I_{sq}(k+1) &= I_{sq}(k) + \epsilon(V_{sq}(k) - R_s I_{sq} - \sigma L_s \omega_e I_{sd}(k) - \omega_e \frac{L_m^2}{L_r} I_{sd}) \end{aligned} \quad (4.24)$$

where $\epsilon = \frac{T_s}{\sigma L_s}$ and $\sigma = 1 - \frac{L_m^2}{L_s L_r}$

The predicted currents in 4.24 are then used to formulate the cost function for model predictive control.

4.5.3 Cost function

Once the predictions of control variables are obtained, a cost function which translates the desired behaviour of the system is designed. An output current reference is used in the case of RL load to control the current. Thus the resulting cost function will

be as in 4.25.

$$G_a = | I_{ref}(k+1) - I_o(k+1) | \quad (4.25)$$

To demonstrate the multi-variable control capability of MPC, instantaneous input reactive power is also taken into account. The instantaneous input reactive power of the system at instant $k+1$ is given by 4.26.

$$Q(k+1) = V_{s\beta}(k+1)I_{s\alpha}(k+1) - I_{s\beta}(k+1)V_{s\alpha}(k+1) \quad (4.26)$$

The cost function corresponding to reactive power control can then be written as

$$G_b = | Q_{ref}(k+1) - Q(k+1) | \quad (4.27)$$

where $Q_{ref} = 0$ for unity power factor.

The final cost function to control both load current and instantaneous input reactive power is given by

$$G = G_a + \mathbf{W}G_b \quad (4.28)$$

where $\mathbf{W}$ is the weighting factor which dictates the prominence of one control variable over the other in case of multi-variable control. There is no specific procedure or calculation for the selection of $\mathbf{W}$ and this has been a topic of debate in the area of model predictive control over the last few years. One way to find the weighting factor is by normalization when there are more than one control variables. In this case, the normalization procedure will result in a specific weighting factor for each of the control variables and will be fixed. Another approach is to fix the weight of one variable and changing the weight of the other variable in the cost function until desired behaviour is attained.

As seen from 4.28 the cost function is essentially the magnitude of the error between the reference and the measured value of the controlled variables. The objective of the control is to minimize this error and find the switching combinations that will produce the minimum error.

4.5.4 Implementation issues- Computation delay

As explained earlier, MPC involves measurement of control variables, prediction of the variable at the next sampling instant and cost function optimization to select the optimum switching states. Ideally the time taken to perform these computations should be negligible. In that case the measurement of currents and application of the optimum switching state will take place in the same sampling instant. Thus the switching state which will produce minimum error at $(k+1)$ is applied at instant k .

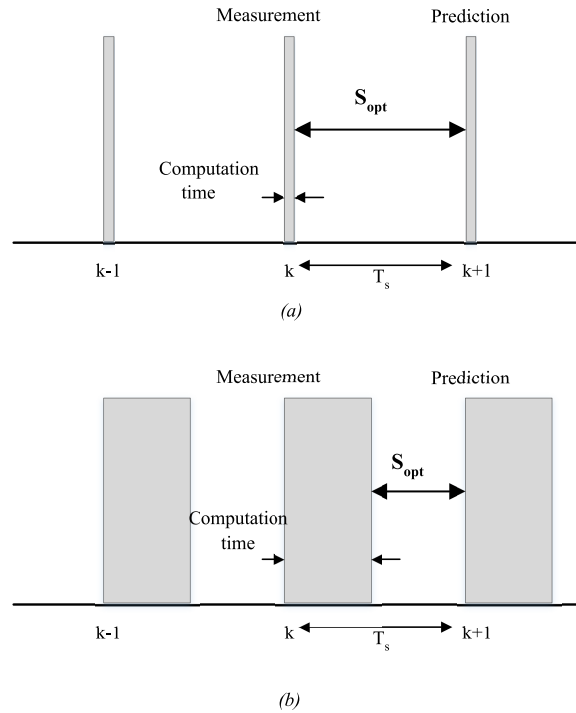


Figure 4.15: (a) Ideal and (b) actual implementation of FCS-MPC

For a matrix converter these computations have to be repeated 27 times as there are 27 switching states and practical implementation using microprocessors is not straightforward. In reality, the time taken to compute the predictions and cost functions for all switching states is significant enough to cause a delay. So the switching states are not applied exactly at instant k but with a delay as shown in Figure 4.15. This delay has to be compensated as it affects the performance of the control. This is done by predicting two steps ahead and optimizing the cost function at $(k+2)$ instead of $(k+1)$. The currents at $(k+1)$ can be estimated from the switching states applied

at k with the help of the system model as shown in 4.29.

$$I_o(k+2) = \left(1 - \frac{RT_s}{L}\right)I_o(k+1) + \frac{T_s}{L}V_o(k+1) \quad (4.29)$$

A flowchart of the MPC control strategy for matrix converter with delay compensation is shown in Figure 4.16. Even though the flowchart shows indicated two different loops for the cost function optimization and duty cycle calculation, in practice it is implemented in a single loop.

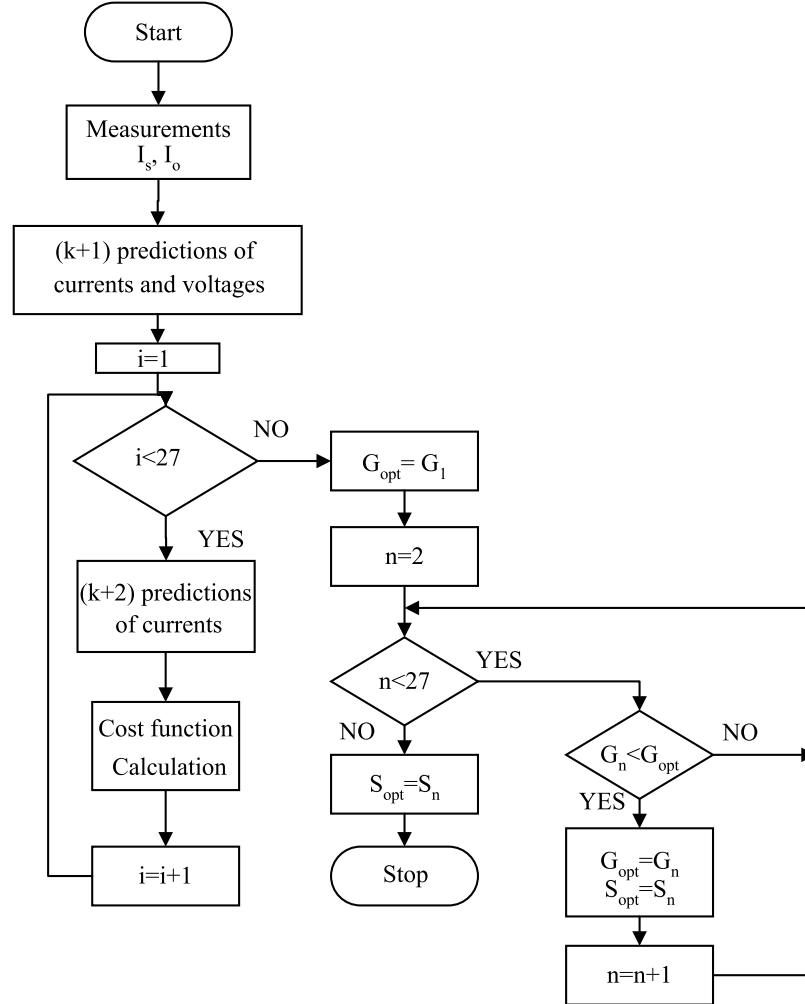


Figure 4.16: Flowchart of MPC with delay compensation

4.6 Simulation Results

Simulation of a direct matrix converter controlled by MPC is done for both RL load and an induction motor in Matlab Simulink (control algorithm) environment together with Ples blockset (converter model). It should be noted that the simulation tests are performed for load current control only and the supply current remains uncontrolled. The simulation results for these scenarios employing $(k+2)$ predictions are discussed in the following sections. The simulation also takes into account the four-step commutation in matrix converter and the measurement delays in the FPGA. A control block diagram of MPC for matrix converter is shown in Figure 4.17.

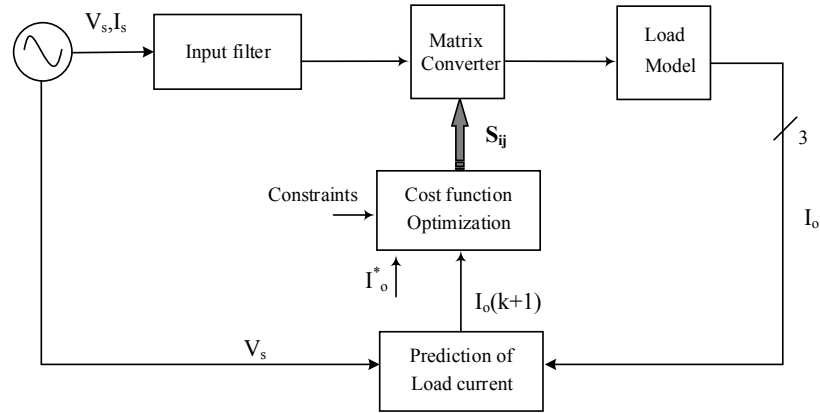


Figure 4.17: Block diagram of MPC for a matrix converter

4.6.1 Matrix Converter Feeding RL Load

The control of load current of matrix converter feeding RL load using MPC is discussed in this section. An open loop schematic of the matrix converter system is shown in Figure 4.12. In the case of RL load, control of the currents is achieved in alpha-beta reference frame to simplify and reduce the computation complexity. The parameters of the system are given in Table 4.3.

Table 4.3: Simulation parameters

Parameter	Value	Unit
<i>Load resistance, R</i>	10	Ω
<i>Load inductance, L</i>	3.75	mH
<i>Input filter capacitance, $C_f(\text{delta})$</i>	8.3	μF
<i>Input filter inductance, L_f</i>	0.7	mH
<i>Damping resistor, R_d</i>	10	Ω
<i>Supply frequency</i>	50	Hz
<i>Output frequency</i>	30	Hz
<i>Sampling time, T_s</i>	80	μs
<i>Supply Voltage, V_s</i>	80	V

MPC is employed to control the load current when a reference of 5A at 30Hz is demanded from the system. The predicted load current is obtained from the load model using 4.22 and is used to calculate the cost function. The switching state resulting in the minimum value for cost function is selected by the predictive controller. The three phase controlled load currents and output line voltage of matrix converter is shown in Figure 4.18. Figure 4.18 also shows the Fast Fourier Transform (FFT) of Phase A load current. It is evident from the figure that the harmonics are concentrated at the lower frequency range as expected from a predictive controller. This is because the controller chooses a switching state that produces the minimum current error and the switching state selected in the next instant does not depend on the number of commutations required but solely on the cost function value. THD of the controlled load current is approximately 8.09% which is not desirable and significantly high compared to methods involving modulation.

To demonstrate the dynamics of the controller, a step change in current reference magnitude from 2A to 4A is introduced at $t=0.2s$. The performance of the controller in this case is depicted by Figure 4.20. It is evident from Figure 4.20 that the controller responds to the change in magnitude swiftly.

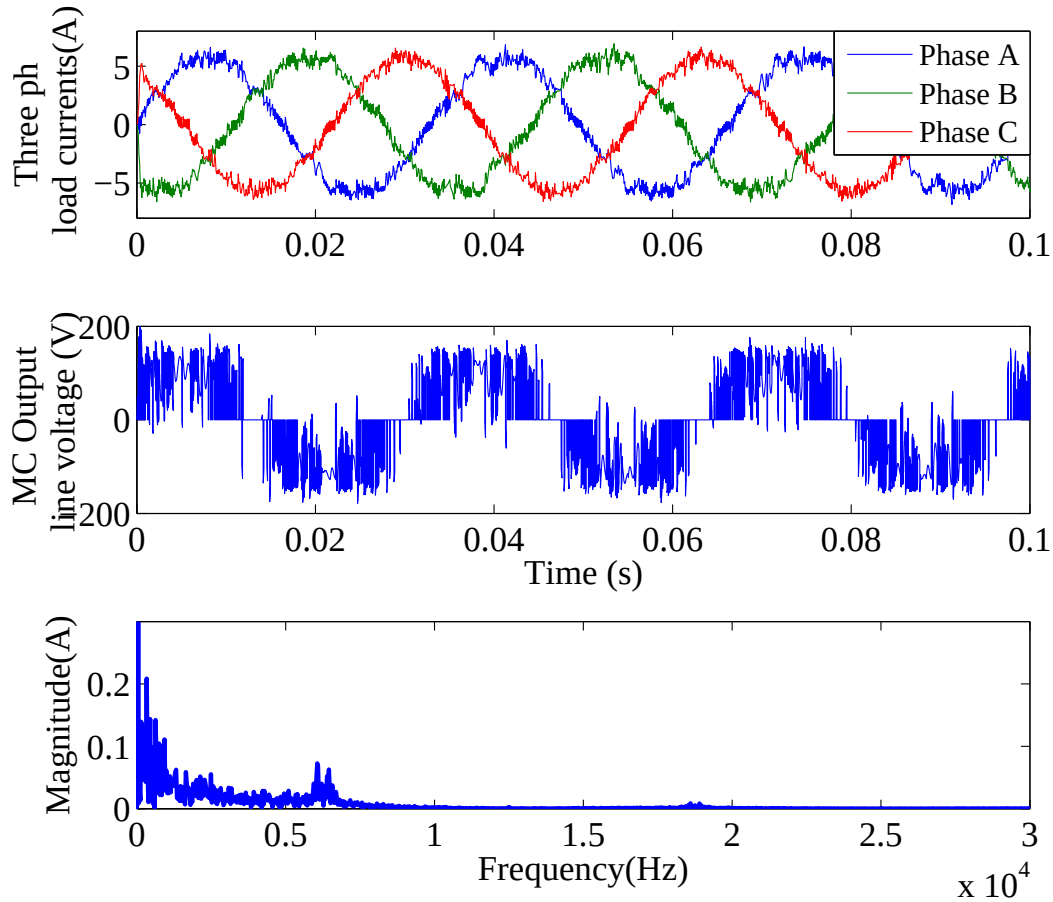


Figure 4.18: Three phase load currents (top), MC output line voltage (middle) and FFT of Phase A load current (bottom) at steady state

Since the source currents are not explicitly controlled, the only criterion the controller considers is the load current error. This results in an input current will be uncontrolled and not sinusoidal as shown in Figure 4.19.

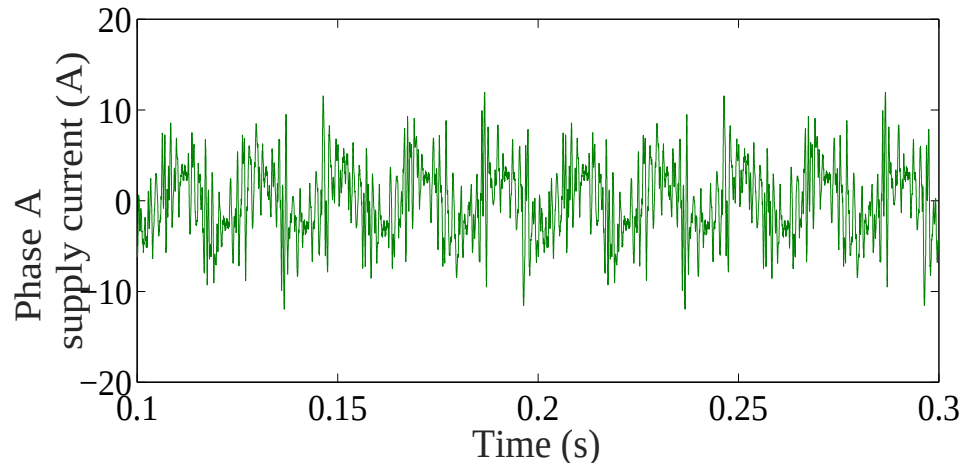


Figure 4.19: Phase A supply current of MC

Figure 4.21 shows the performance of the controller when the frequency of the output current reference is changed from 20Hz to 40Hz at $t=0.2\text{s}$. The load currents track the changes in the references swiftly, exhibiting a fast transient response.

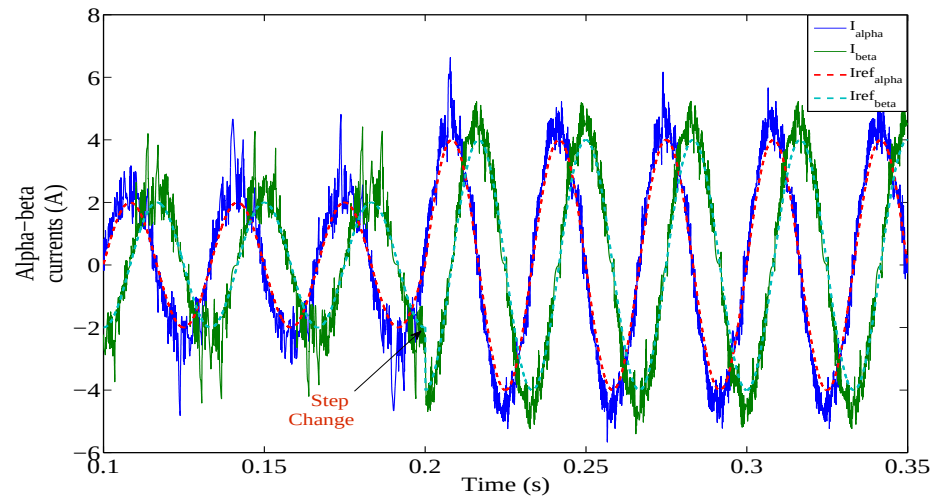


Figure 4.20: Step change in magnitude of the load current demand

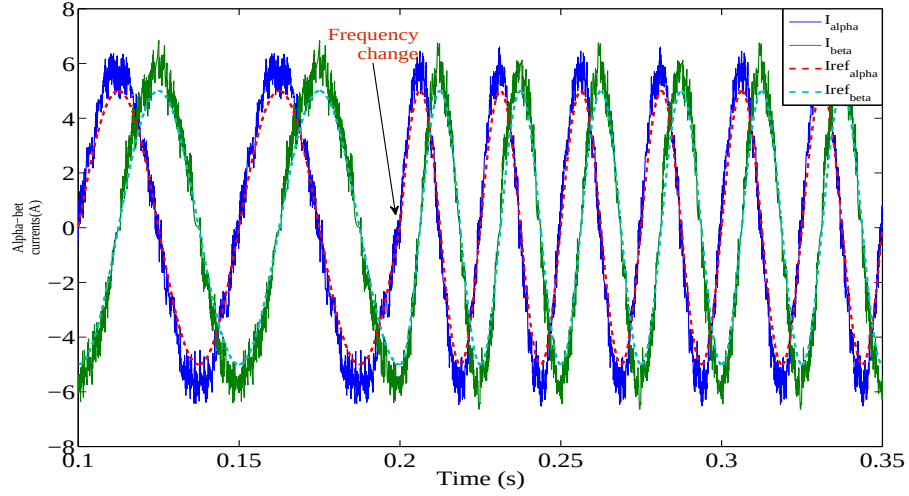


Figure 4.21: Change in frequency of the load current demand

4.6.2 Matrix Converter Feeding an Induction Machine

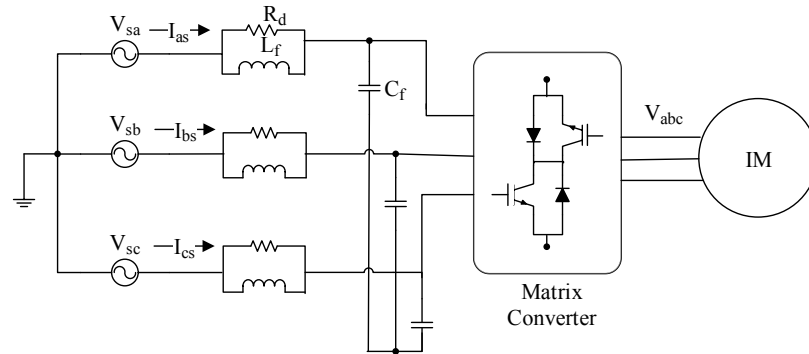


Figure 4.22: Open loop schematic of matrix converter drive

A 3kW, star connected induction motor is considered to implement MPC for the stator current control. The speed of the machine is controlled by a conventional PI controller. The q-axis current reference is obtained from the speed controller. An open loop schematic of the system is shown in Figure 4.22. The parameters of the machine are as shown in Table 3.2. The control block diagram for MPC of a matrix converter drive is shown in Figure 4.23.

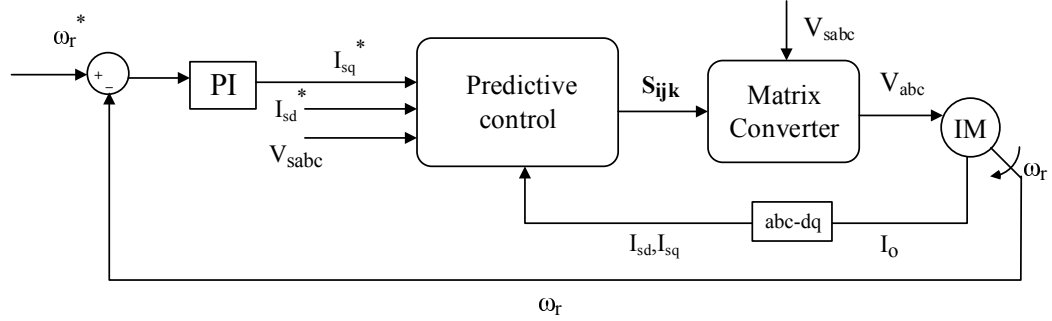


Figure 4.23: Control block diagram of MPC for matrix converter drive

A speed demand of 500rpm is applied to the speed controller at no load condition and the resulting waveforms are shown in Figure 4.24 and Figure 4.25. The reference tracking of d and q-axis stator currents are shown in Figure 4.24 and it is evident from the graph that the dq stator currents follow the references with negligible steady state error. It can be noted that the starting torque of the machine in Figure.4.25 reaches the nominal value slowly. This is because the speed demand was applied before the machine is magnetized. The THD of the controlled stator currents is approximately 5.1%.

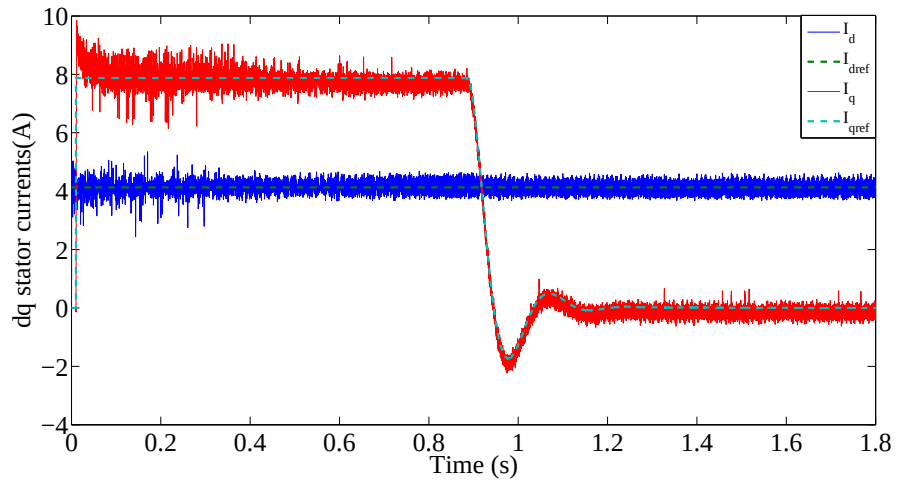


Figure 4.24: Reference tracking of dq stator currents at no load operation

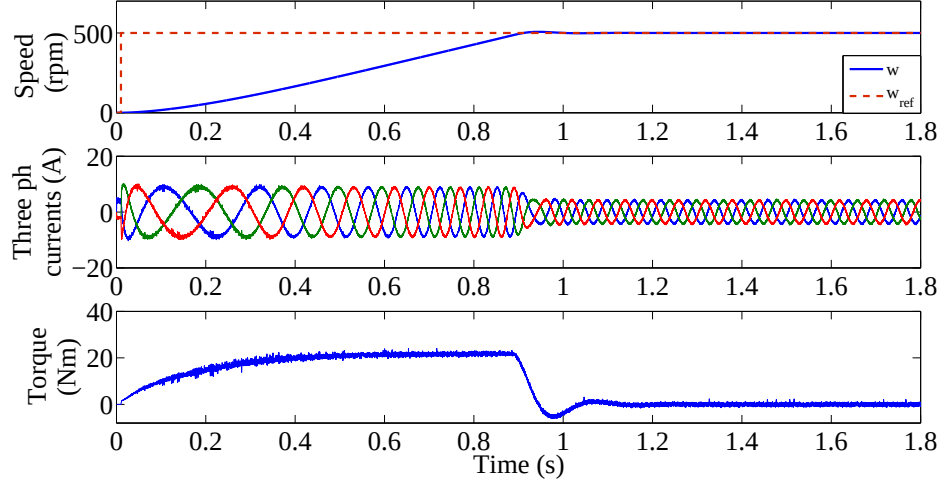


Figure 4.25: Speed, Three phase stator currents and torque of the machine during transient and steady state operation at no load

Figure 4.25 shows the speed of the machine together with the speed reference, three phase stator currents and electromagnetic torque during the transient and steady state operation. To study the dynamics of the controller with step changes in the reference quantities, the speed reference is changed from 200rpm to 500rpm at $t=1.2$ s. The resulting waveforms of the speed, three phase currents and torque are shown in Figure 4.26. In Figure 4.24 and Figure 4.25, it can be noted that the torque response is slow at the beginning of the test. This is because, in the simulation the speed demand was applied before the machine is magnetised. From these waveforms it is clear that MPC has a fast transient response and can yield good results with an accurate system model.

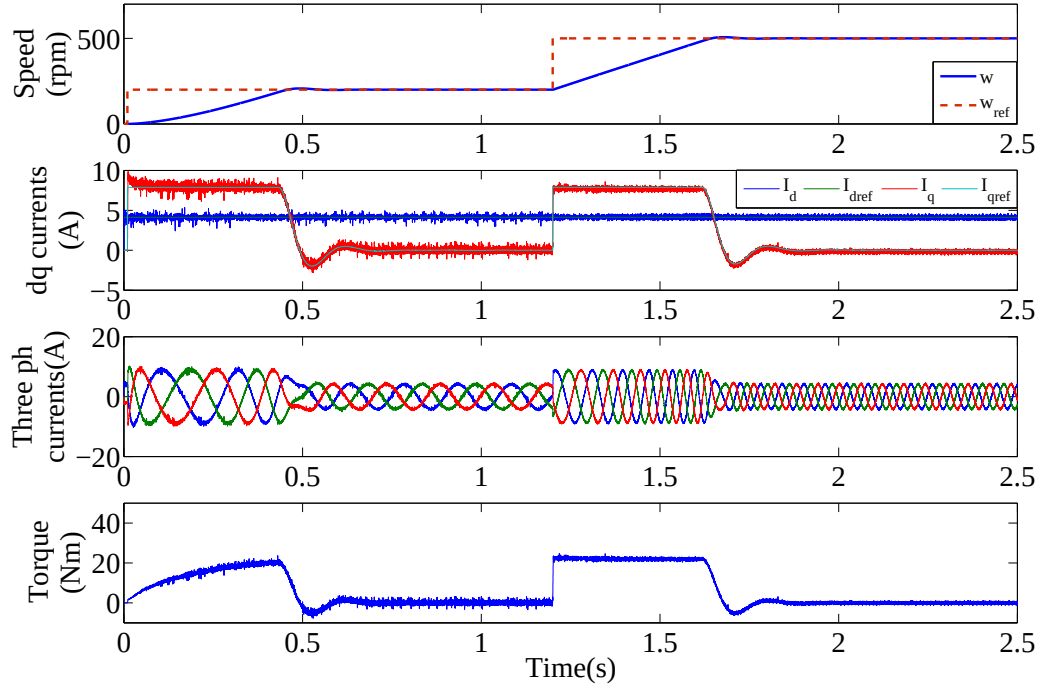


Figure 4.26: Speed, Three phase stator currents and torque of the machine during transient and steady state operation at no load

4.7 Summary

The basic principle of predictive control and various categories of the same were discussed in this chapter. Finite-set Model Predictive Control has been given special emphasis, being a relatively modern control strategy that has gained popularity in recent years for the control of power converters. Multi-variable control capability without the need for nested loops, inclusion of non linearities and control constraints in the cost function, fast transient response are some of the promising features of MPC. The shortcoming of this method is the variable switching frequency operation.

MPC applied to a 3-phase two level inverter is used to elaborate on the control strategy. Simulation results for control of matrix converter feeding an RL load and IM shows the ability of the controller to control more than one variable without cascaded loops. The advantages of this method shows its relevance in applications

that require fast transient response.

Chapter 5

Fixed Switching Frequency Predictive Control for a Direct Matrix Converter

5.1 Introduction

A fixed switching frequency predictive control attempts to overcome one of the major shortcomings of MPC which is variable switching frequency. This chapter studies two different strategies that can achieve a fixed switching frequency operation for a direct matrix converter. This is achieved by having an intrinsic modulation scheme in the control algorithm. The two methods discussed in this chapter are Modulated Model Predictive Control (M2PC) and Direct Predictive Current-error Vector Control (DPCVC). The basic principle of M2PC and DPCVC and their implementation procedure when applied to a direct matrix converter feeding a resistive inductive load and an induction machine are discussed in this chapter. Simulation results of load current control of a direct matrix converter using M2PC and DPCVC are also included to validate the control theory and analyse the performance.

5.2 M2PC control strategy

The main objective of M2PC is to combine the ideas of SVM and MPC to create a control method with the positive features of both [66]. The result is a multi-objective model predictive control strategy with fixed switching frequency. This is achieved by embedding a modulation scheme in the predictive control structure.

Table 5.1: Comparison of features of SVM with PI controller, MPC, M2PC and DPCVC

Features	PI with SVM	MPC	M2PC	DPCVC
Fixed switching frequency	✓	✗	✓	✓
Modulation scheme	✓	✗	✓	✓
Multi variable control without cascaded loops	✗	✓	✓	✓
Addition of non-linearities and Constraints	✗	✓	✓	✓

A comparison of the features of M2PC, MPC, DPCVC and PI controller with SVM is shown in Table 5.1. The key components of the M2PC method include predictive control using cost function optimization and a modulation scheme to generate the voltage vectors and their application times within a sampling interval as dictated by the predictive controller. The control variables are predicted for future sampling intervals to form the cost function which describes the desired behaviour of the system. Unlike MPC where the switching state that results in the minimum cost function is applied for the whole sampling duration, the M2PC applies more than one vector within one sampling interval. The number of vectors applied depends on the converter. The application times for the vectors applied are calculated using the cost functions for different switching states. The application of vectors for specified times within the sampling interval to generate the desired voltage vector is the main characteristic of M2PC strategy.

The working principle of the M2PC is based on the assumption that the dynam-

ics of the system/load within one sampling interval is linear. This means that any non-linearities in the load within the sampling interval will not be considered. This approximation is reasonable for a RL load and induction machine.

The M2PC strategy has been successfully applied to a three phase active front rectifier [63], three phase two level inverter [68], cascaded multilevel converters [66, 65, 64, 67] and indirect matrix converter [69, 70] with excellent results. This is being extended to a direct matrix converter for load current control in this thesis.

5.3 M2PC for 2-level inverter

The fundamental principle of M2PC can be explained using the example of a three phase 2-level inverter [68]. A three phase 2-level inverter in Figure 5.1 has eight switching states out of which six produce active or non-zero vectors and two produce zero vectors [68].

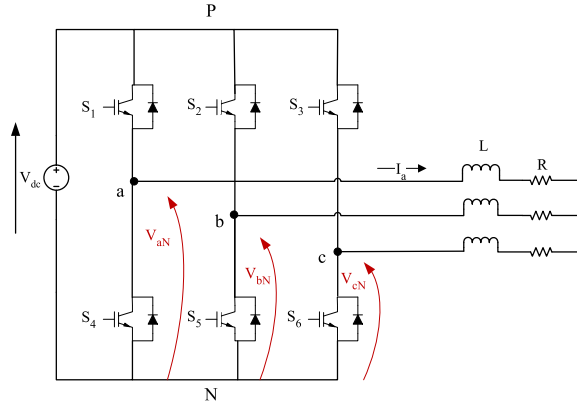


Figure 5.1: Schematic of a three phase 2-level inverter

The desired voltage in the inverter is produced by the combination of two consecutive active vectors and two zero vectors for their respective duty cycles. In standard SVM, the duty cycles are calculated based on the voltage vector magnitudes whereas in M2PC this is based on the value of cost functions / error for two adjacent switching states. The M2PC algorithm minimizes the cost functions for each switching state

and applies the vectors with minimum error for their respective duty cycles within a sampling interval.

The switching states and available voltage vectors for a three phase 2-level inverter are shown in Table 4.2. To elaborate on the M2PC strategy, load current control in the case of a RL load is considered. The predicted load currents in alpha-beta co-ordinate frame for all the 8 switching states can be derived from 5.1.

$$\begin{aligned} I_\alpha(k+1) &= (1 - \frac{RT_s}{L})I_\alpha(k) + \frac{T_s}{L}V_\alpha(k) \\ I_\beta(k+1) &= (1 - \frac{RT_s}{L})I_\beta(k) + \frac{T_s}{L}V_\beta(k) \end{aligned} \quad (5.1)$$

The cost functions for each switching state is calculated using 5.2 and minimized as in the predictive control strategy. This results in three cost functions G_1, G_2 and G_0 for each voltage vector needed for modulation.

$$G_j = |I_{\alpha ref}(k+1) - I_\alpha^j(k+1)| + |I_{\beta ref}(k+1) - I_\beta^j(k+1)| \quad (5.2)$$

where $j=0,1,2$. The voltage vectors produced by each switching state for a VSI is shown in Figure 5.2. The duty cycle for each vector is then calculated by minimizing the cost function for two adjacent states. This is done for all possible switching combinations such as (1,2),(2,3),(3,4),(4,5),(5,6),(6,1).

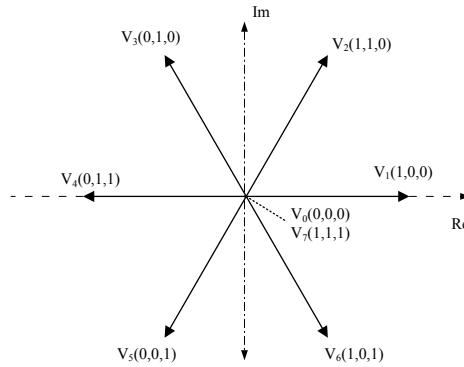


Figure 5.2: Voltage vectors generated by a 3-ph inverter

The duty cycles are derived by solving the set of linear equations in 5.3 based on the behaviour of predictive control that the application time of a vector will be inversely

proportional to its cost function.

$$\begin{aligned}
 t_1 &= \frac{k_1}{G_1} \\
 t_2 &= \frac{k_1}{G_2} \\
 t_0 &= \frac{k_1}{G_0} \\
 t_1 + t_2 + t_0 &= T_s
 \end{aligned} \tag{5.3}$$

where k_1 is the normalizing constant and t_1, t_2 and t_0 are the application times for the active and zero vectors. Once the value of k_1 is calculated from 5.3, the duty cycles can be determined using 5.4.

$$d = \frac{t}{T_s} \tag{5.4}$$

The resulting expression for the duty cycles of the active and zero vectors in terms of their cost functions is given by 5.5.

$$\begin{aligned}
 d_1 &= \frac{G_0 G_2}{G_1 G_0 + G_2 G_0 + G_1 G_2} \\
 d_2 &= \frac{G_0 G_1}{G_1 G_0 + G_2 G_0 + G_1 G_2} \\
 d_0 &= \frac{G_1 G_2}{G_1 G_0 + G_2 G_0 + G_1 G_2}
 \end{aligned} \tag{5.5}$$

Once the duty cycles are calculated, the new cost functions for each adjacent pair of vectors is minimized to obtain the optimal set of vectors with their respective duty cycles. The new cost function which determines the switching states to be applied in the next sampling interval is given by 5.6.

$$G_{new} = d_1 G_1 + d_2 G_2 + d_0 * G_0 \tag{5.6}$$

The minimization procedure is then to select the set of vectors that can produce the least cumulative error in a similar manner to that of a predictive control algorithm. The two active vectors which minimizes the cumulative cost function dictated by 5.6 is chosen to be applied in the next sampling interval together with the zero vectors. The vectors are applied in a symmetrical switching pattern as shown in Figure 5.3 with the objective of applying two active and two zero vectors.

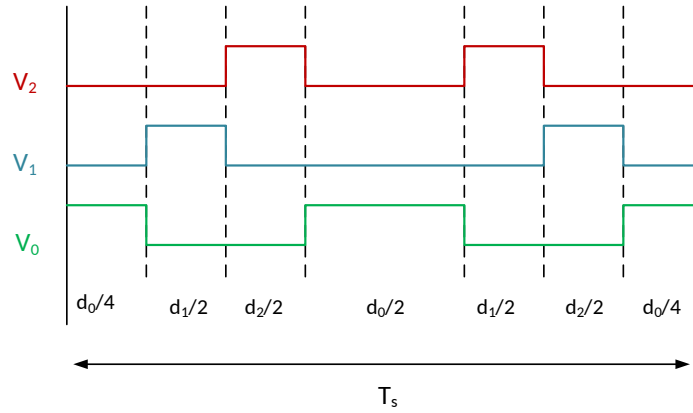


Figure 5.3: Switching pattern for 2-level inverter

5.4 M2PC for a Direct Matrix converter

Applying the M2PC to a direct matrix converter is a complex task due to the presence of 27 switching states [35] resulting in a higher number of iterations to select a set of voltage vectors. This increases the computational effort compared to that of a 2-level inverter, which only presents 8 switching states.

The procedure required for the implementation of M2PC is very similar to the conventional MPC method explained in chapter 4, with the addition of an algorithm to calculate the duty cycles for the vectors applied. This procedure is composed of four stages such as:

- Determination of system model.
- Prediction of the control variable.

- Cost function calculation.
- Duty cycle calculation.

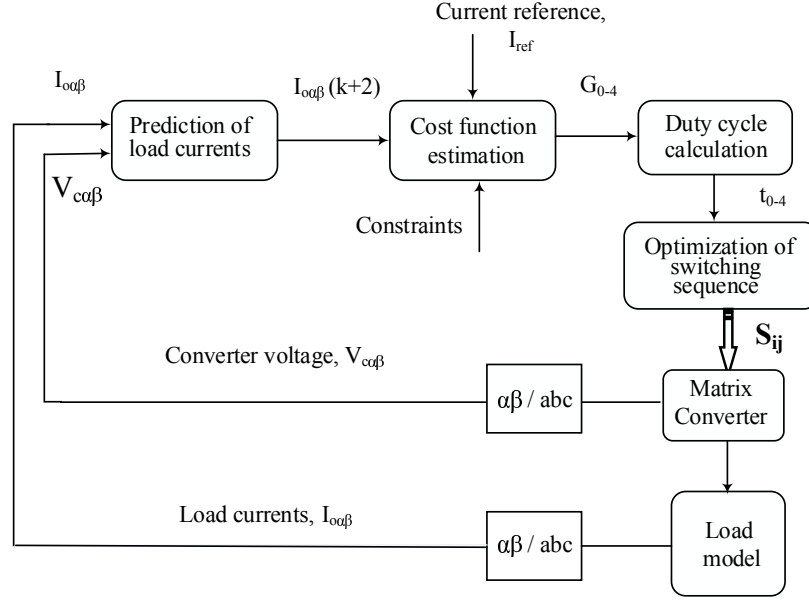


Figure 5.4: Control block diagram for M2PC

Figure 5.4 shows the control block diagram for the M2PC method applied to a direct matrix converter. A reference current is imposed upon the system and the controller is designed using the system model discussed in section 5.4.1, so that the load current tracks the reference. The measured currents are then used to predict the value of the current at the future sampling instant. The reference and predicted currents are then used to calculate the cost function which will be used to compute duty cycles for the active and zero vectors. Unlike 2-level inverters with two active and two zero vectors, matrix converter usually makes use of four active and three zero vectors to obtain sinusoidal currents and voltages.

5.4.1 System Model

The system model plays a vital role in model predictive control strategies. For load current control, it includes the converter model and the load model. A schematic of

the matrix converter circuit considered in the design of the control algorithm is shown in Figure 5.5. The circuit consists of direct matrix converter, an input filter and a RL load.

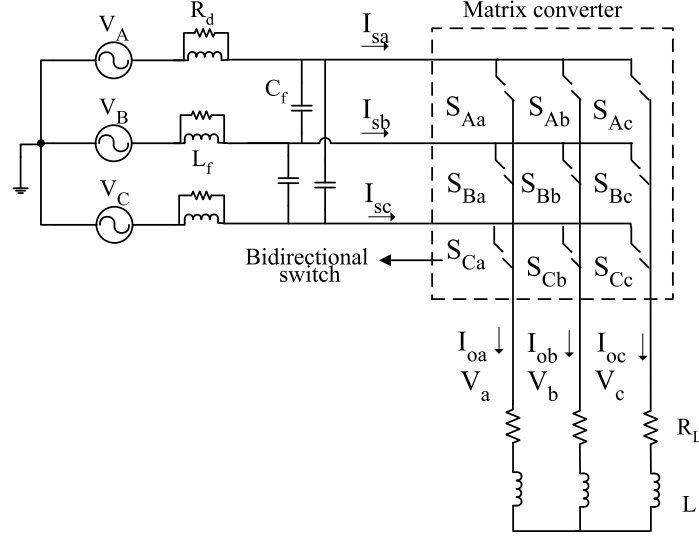


Figure 5.5: Schematic of matrix converter feeding RL load

5.4.1.1 Converter Model

A model of the converter is required to obtain the output voltages of the converter for each switching state. The converter output voltages and input currents can be obtained from Figure 5.5 as follows:

$$\begin{bmatrix} v_a(t) \\ v_b(t) \\ v_c(t) \end{bmatrix} = \underbrace{\begin{bmatrix} m_{Aa} & m_{Ba} & m_{Ca} \\ m_{Ab} & m_{Bb} & m_{Cb} \\ m_{Ac} & m_{Bc} & m_{Cc} \end{bmatrix}}_M \cdot \begin{bmatrix} v_A(t) \\ v_B(t) \\ v_C(t) \end{bmatrix} \quad (5.7)$$

$$\begin{bmatrix} i_A^-(t) \\ i_B^-(t) \\ i_C^-(t) \end{bmatrix} = \underbrace{\begin{bmatrix} m_{Aa} & m_{Ab} & m_{Ac} \\ m_{Ba} & m_{Bb} & m_{Bc} \\ m_{Ca} & m_{Cb} & m_{Cc} \end{bmatrix}}_{M^T} \cdot \begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} \quad (5.8)$$

where $\bar{\mathbf{v}}$ and $\bar{\mathbf{i}}$ are the output voltage and input current vectors respectively. $\mathbf{M}$ is the transfer matrix consisting of the modulation indices for all the nine bidirectional switches.

5.4.1.2 Load Model

The two types of load considered in this study are resistive-inductive (RL) load and an induction machine. The discrete time load equations for an RL load can be derived from the continuous time model of the circuit shown in Figure 5.5 as follows:

$$V(k) = (R - \frac{L}{T_s})I_o(k) + \frac{L}{T_s}I_o(k+1) \quad (5.9)$$

where $V(k)$ - Converter voltage at instant k

$I_o(k)$ - Load current at instant k

$I_o(k+1)$ - Load current at instant $k+1$

T_s - Sampling time

The discrete time equations for the stator voltages of an induction machine can be obtained from the continuous time equations in the field oriented reference frame [106]. The resulting model for the IM is shown in 5.10.

$$\begin{aligned} V_{sd} &= I_{sd}R_s + \frac{d\Psi_{sd}}{dt} - \omega_e \Psi_{sq} \\ V_{sq} &= I_{sq}R_s + \frac{d\Psi_{sq}}{dt} + \omega_e \Psi_{sd} \end{aligned} \quad (5.10)$$

where Ψ_{sd} and where Ψ_{sq} are the d-axis and q-axis stator flux respectively.

5.4.2 Prediction of Control Variables

For an RL load, prediction of load current makes use of the current measurement at the present instant or instant 'k' and the load parameters as in 5.9. In MPC, the prediction of load current is carried out once using the optimal switching state from the previous instant. Since the M2PC incorporates a modulation scheme, more than one active vector will be applied during one sampling interval. For a matrix converter, out of the 27 switching states, the six rotating vectors are not considered for modulation resulting in 21 switching states to choose from. As in a standard SVM strategy [26] it is assumed that a minimum of four active vectors is required to get sinusoidal waveforms for a matrix converter. Hence the M2PC strategy will also consider four active vectors and three zero vectors to apply during one sampling interval. The four active vectors are obtained by selecting the sequence adopted by a standard SVM scheme [26]. The sequence in SVM is formulated by considering the input current sector and output voltage sector which will amount to a total of 36 sequences. Each sequence consists of four active and three zero vectors. Since the load current is controlled in this case, the output voltage sector is unknown. Therefore the selection of the sequences is made based on the input current sector that can be obtained from the supply voltages. Thus for every input sector, there will be 18 sequences of vectors that can produce the desired load current. The M2PC controller selects the sequence which produces the minimum cumulative error. The predictions for four active vectors and zero vectors are done as if they were applied for the whole sampling interval. For example, for one sequence containing four active and three zero vectors, the predictions of load current are calculated as shown in 5.11 and 5.12 respectively. The predicted value of load current for all the three zero vectors will be the same and hence it is only required to calculate it once.

For active vectors :

$$\begin{aligned}
 I_o^1(k+1) &= (1 - \frac{RT_s}{L})I_o^1(k) + \frac{T_s}{L}V_o^1(k) \\
 I_o^2(k+1) &= (1 - \frac{RT_s}{L})I_o^2(k) + \frac{T_s}{L}V_o^2(k) \\
 I_o^3(k+1) &= (1 - \frac{RT_s}{L})I_o^3(k) + \frac{T_s}{L}V_o^3(k) \\
 I_o^4(k+1) &= (1 - \frac{RT_s}{L})I_o^4(k) + \frac{T_s}{L}V_o^4(k)
 \end{aligned} \tag{5.11}$$

For zero vectors:

$$I_o(k+1) = (1 - \frac{RT_s}{L})I_o(k) \tag{5.12}$$

For an IM, the procedure remains the same. The equations for stator current predictions in dq co-ordinate frame are obtained as shown in 5.13.

$$\begin{aligned}
 I_{sd}(k+1) &= I_{sd}(k) + \epsilon(V_{sd}(k) - R_s I_{sd}(k) \\
 &\quad + \sigma L_s \omega_e I_{sq}(k)) \\
 I_{sq}(k+1) &= I_{sq}(k) + \epsilon(V_{sq}(k) - R_s I_{sq}(k) \\
 &\quad - \sigma L_s \omega_e I_{sd}(k) - \omega_e \frac{L_m^2}{L_r} I_{sd}(k))
 \end{aligned} \tag{5.13}$$

where $\epsilon = \frac{T_s}{\sigma L_s}$ and $\sigma = 1 - \frac{L_m^2}{L_s L_r}$.

As in the case of RL load, predictions have to be made for each of the active and zero vectors using 5.13. The resulting equations are then used to formulate the cost function.

5.4.3 Cost Function Calculation

Just as in MPC, the switching states are selected based on a cost function minimization. The cost function can include different performance factors according to the control variables and constraints required. More than one variable can be controlled with the same control loop, providing a multi-objective control approach and avoiding nested loops.

The cost function for load current control is essentially the error between the current demand and predicted current at (k+1) instant. This results in five cost functions G_0, G_1, G_2, G_3 and G_4 for each active and zero vectors. The set of quadratic cost functions can be calculated using the pre-calculated predicted currents in 5.11 as shown in 5.14.

$$\begin{aligned}
 G_0 &= (I_{oref}(k+1) - I_o^0(k+1))^2 \\
 G_1 &= (I_{oref}(k+1) - I_o^1(k+1))^2 \\
 G_2 &= (I_{oref}(k+1) - I_o^2(k+1))^2 \\
 G_3 &= (I_{oref}(k+1) - I_o^3(k+1))^2 \\
 G_4 &= (I_{oref}(k+1) - I_o^4(k+1))^2
 \end{aligned} \tag{5.14}$$

where $I_{oref}(k+1)$ is the current demand at (k+1) instant.

For an induction machine load, the cost function for stator current control can be represented as follows:

$$G_j = (I_{sdref}(k+1) - I_{sd}^j(k+1))^2 + (I_{sqref}(k+1) - I_{sq}^j(k+1))^2 \tag{5.15}$$

where j=0,1,2,3,4 for each of the vectors.

Once the cost functions for each of the active and zero vectors are calculated, the predictive control will choose the best sequence of vectors which produces the least

error. These vectors are then applied within the sampling interval with respective duty cycles. The duty cycles are calculated based on the cost function values of the active and zero vectors.

5.4.4 Duty Cycle Calculation

The application times for each of the vectors are calculated from the cost functions computed for each of the switching state. This method is based on the assumption that, per sampling interval, the system dynamic is linear in nature. From this it is possible to calculate approximate application times for each vector as a percentage of the total sampling time. The relation between the application times of the vectors and their corresponding cost functions can be written as follows.

$$t = \frac{k_1}{G} \quad (5.16)$$

where k_1 is the proportional constant. Using the above equation for four active vectors and zero vector for a MC will result in a set of five equations each dictating the application times for a particular vector in terms of the proportional constant. The condition that holds true for all the vectors applied is that the sum of the application times for all the vectors should be equal to the sampling time as shown in 5.17.

$$t_1 + t_2 + t_3 + t_4 + t_0 = T_s \quad (5.17)$$

where t_1, t_2, t_3, t_4 are the times for the four active vectors and t_0 is the total time for three zero vectors. By substituting 5.16 in 5.17, the proportional constant can be obtained as shown in 5.18.

$$k_1 = \frac{G_0 G_1 G_2 G_3 G_4 T_s}{(G_1 G_2 G_3 G_4 + G_0 G_2 G_3 G_4 + G_0 G_1 G_3 G_4 + G_0 G_1 G_2 G_4 + G_0 G_1 G_2 G_3)} \quad (5.18)$$

The application times for the vectors can then be derived by substituting the value of the proportional constant. The resulting set of duty cycles is shown in 5.19.

$$\begin{aligned}
 t_1 &= \frac{(G_0 G_2 G_3 G_4 T_s)}{A} \\
 t_2 &= \frac{(G_0 G_1 G_3 G_4 T_s)}{A} \\
 t_3 &= \frac{(G_0 G_1 G_2 G_4 T_s)}{A} \\
 t_4 &= \frac{(G_0 G_1 G_2 G_3 T_s)}{A}
 \end{aligned} \tag{5.19}$$

where $A = (G_1 G_2 G_3 G_4 + G_0 G_2 G_3 G_4 + G_0 G_1 G_3 G_4 + G_0 G_1 G_2 G_4 + G_0 G_1 G_2 G_3)$.

Once the application times for active vectors are obtained, the time for zero vectors can be calculated from 5.17.

$$t_0 = T_s - (t_1 + t_2 + t_3 + t_4) \tag{5.20}$$

The duty cycles for each of the vectors will be related to the error associated with that voltage vector. The active and zero vectors are then arranged in a symmetrical manner as shown in Figure 5.6; to achieve minimum switching losses and harmonic content. The switching pattern is the same as that used in a SVM scheme in [26]. In Figure 5.6, t_{01}, t_{02} and t_{03} are the application times for the three zero vectors and t_1, t_2, t_3 and t_4 for the four active vectors.

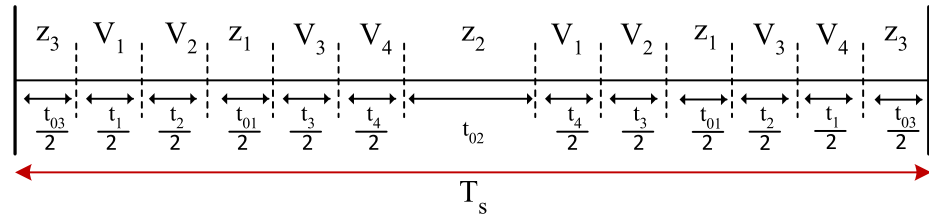


Figure 5.6: Double sided symmetrical arrangement of vectors for a MC

To account for the computation delay in practical implementation of this control method, all the results shown in this thesis compensates for this delay by predicting the control variable for two sampling intervals. Figure ?? shows the experimental

results of matrix converter output current control with a RL load, when the cost function is calculated using $(k+1)$ and $(k+2)$ predictions of control variables.

Figure 5.7 shows the flowchart of M2PC algorithm incorporating delay compensation. As opposed to the two-loop method shown in the flowchart, in practice the prediction, cost function optimization and duty cycle calculation algorithm is carried out in a single loop to reduce computational time.

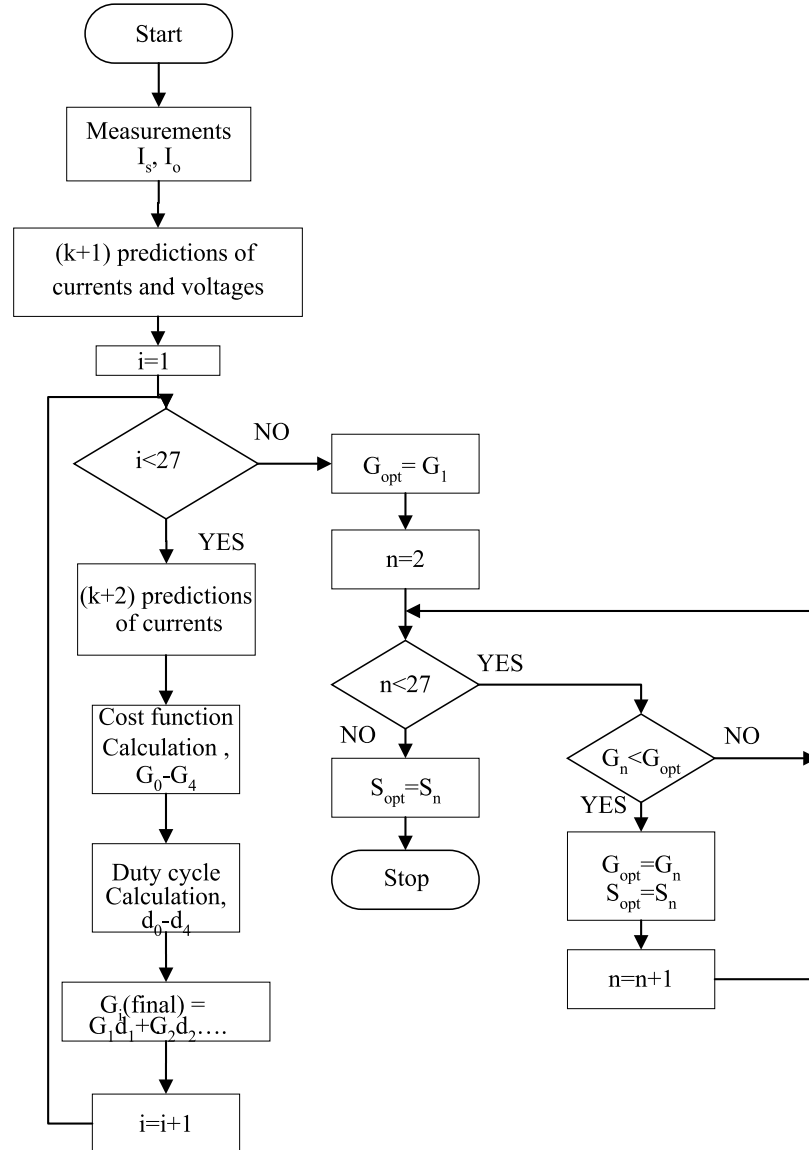


Figure 5.7: Flowchart of M2PC algorithm

5.5 Simulation Results

The M2PC is applied to a direct matrix converter feeding an RL load and an induction motor and simulations are done in Matlab Simulink environment to study the performance. The control algorithm is implemented in Matlab Simulink and the switching model of the converter is modelled using Plecs. A sampling time of $80\mu s$ is considered for both cases. The simulation also takes into account the four-step commutation [102] in direct matrix converter and the measurement delays in the control platform. The simulation tests consider the control of load currents while the supply currents remain uncontrolled.

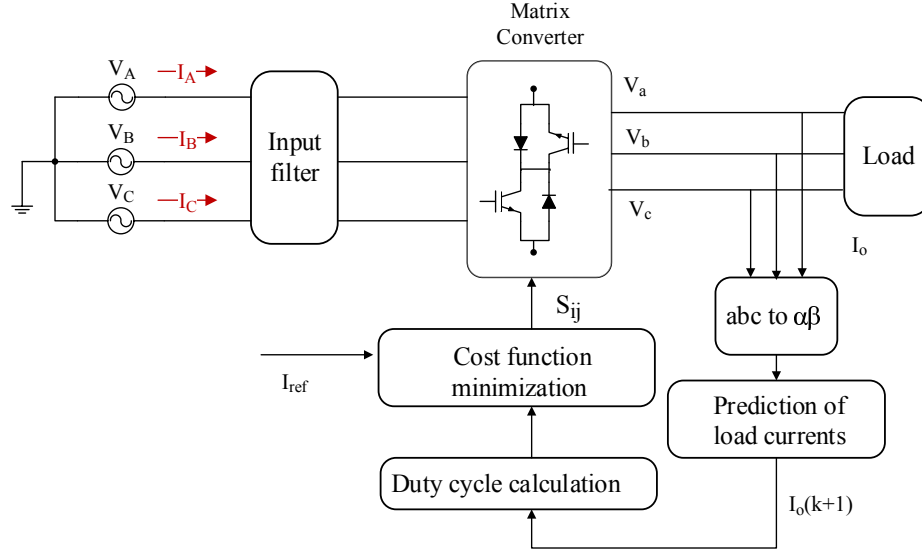


Figure 5.8: Control block diagram of M2PC for a matrix converter

5.5.1 Matrix converter feeding RL load

The load current of a matrix converter feeding an RL load is controlled using the M2PC and the resulting waveforms are presented in this section. A load current reference of 5A at 30Hz is demanded from the system. Figure 5.9 shows the three phase load currents and the corresponding MC line voltage at steady state. A FFT of Phase A load current at steady state is also shown in Figure 5.9. It is worth noting the

presence of harmonics in the range of switching frequency, 12.5kHz and its multiples. This is a result of fixed switching frequency operation. THD of the controlled load current is approximately 6.3 % which is lower than that of the current controlled by MPC. The three phase currents on the supply side of the converter is shown in Figure 5.10. Since the supply side currents are not controlled, it can be noted that they are not sinusoidal.

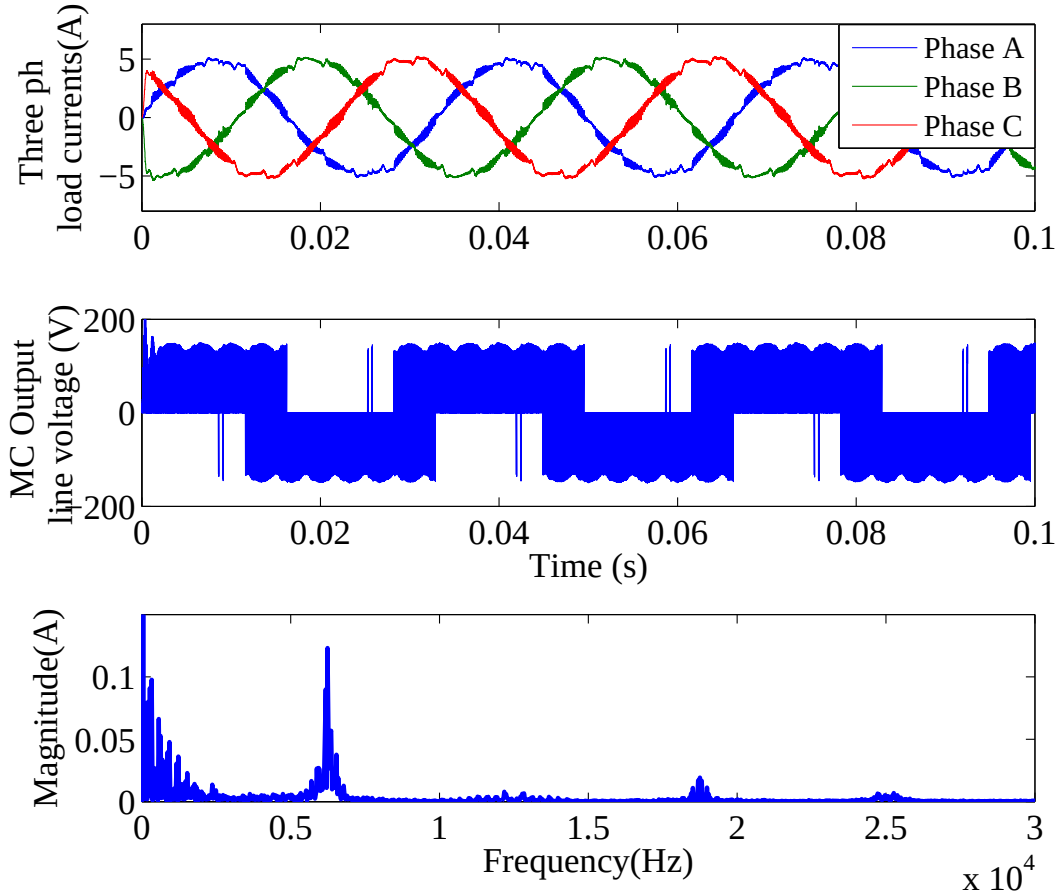


Figure 5.9: Three phase load currents of MC controlled by M2PC

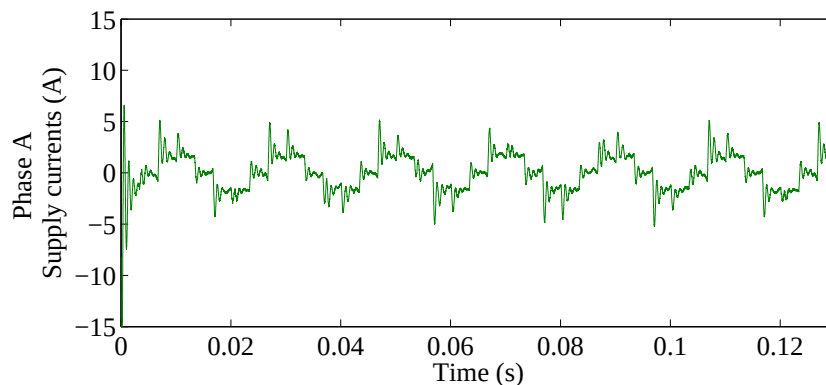


Figure 5.10: Three phase supply currents of MC controlled by M2PC

To demonstrate the fast dynamic response of M2PC, a step change from 2A to 4A is introduced at $t=0.2$ s. The resulting waveform is shown in Figure 5.11. It is evident from the figure that the load current follows the step demand in reference current reaches steady state within a few samples. This proves that M2PC can provide fast transient response which is a characteristic it inherited from MPC.

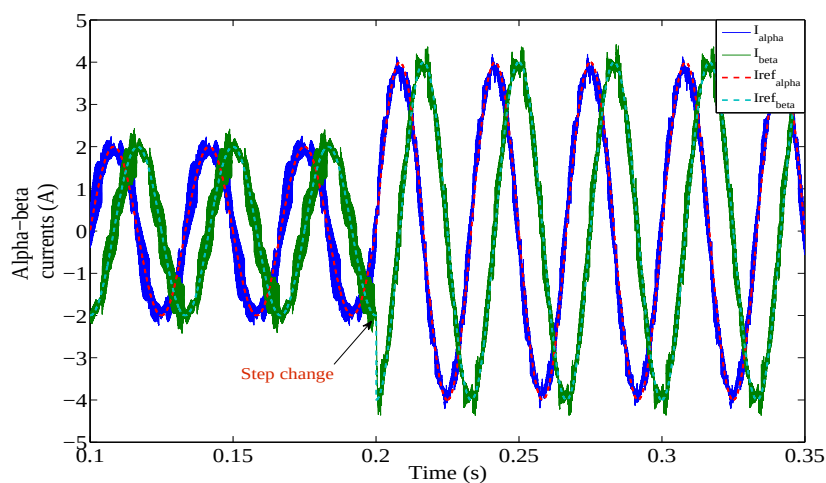


Figure 5.11: Alpha-beta load currents during step change in magnitude of current reference

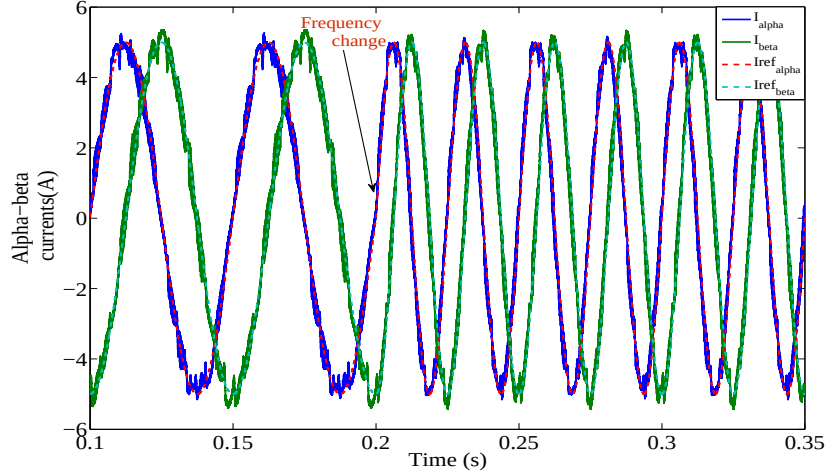


Figure 5.12: Alpha-beta currents when the frequency of reference is changed

In addition to the step change in current reference magnitude, a step change in the frequency of the current reference from 20Hz to 40Hz is also introduced. The resulting waveform is shown in Figure 5.12. The tests show that the M2PC is able to handle any abrupt changes in the reference signal without any overshoots or instability.

5.5.2 Matrix converter feeding an induction machine

A 3kW induction machine is fed using a matrix converter and the stator currents at no load condition are controlled using the M2PC strategy. A schematic of the whole system is shown in Figure 4.22 in section 4.6.2 and a control block diagram for the same is shown in Figure 5.13.

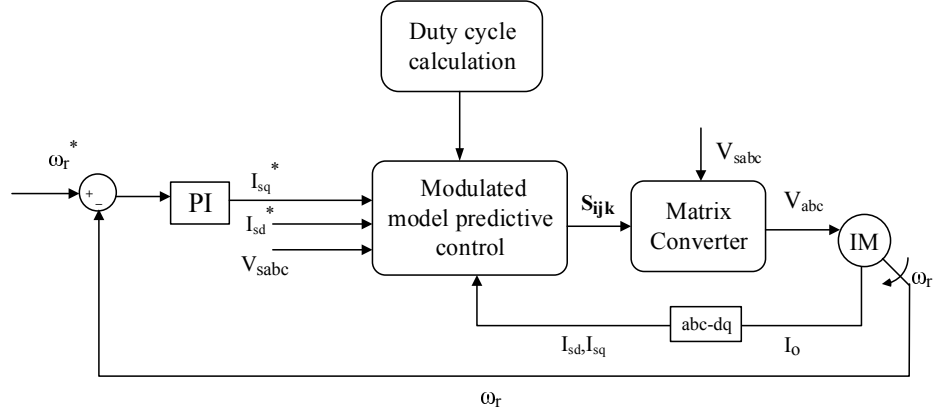


Figure 5.13: Control block diagram of M2PC for matrix converter drive

The parameters of the machine are given in Table 3.2. The speed of the machine is controlled by a PI controller which outputs the q-axis stator current reference. A speed demand of 500rpm is applied to the machine and Figure 5.14 shows the performance of the M2PC current controller compared to the reference currents. It is evident from Figure 5.14 that the stator currents are controlled without any steady state error and present a very fast dynamic response. The THD of the controlled stator currents is approximately 4.6%. However, it was observed that the quality of the controlled currents diminishes with decreasing voltage ratios.

The speed of the machine, three phase currents and the electromagnetic torque of the machine during the same speed transient of Figure 5.14 is shown in Figure 5.15. The stator currents reach the steady state value as expected without any overshoots or ripple.

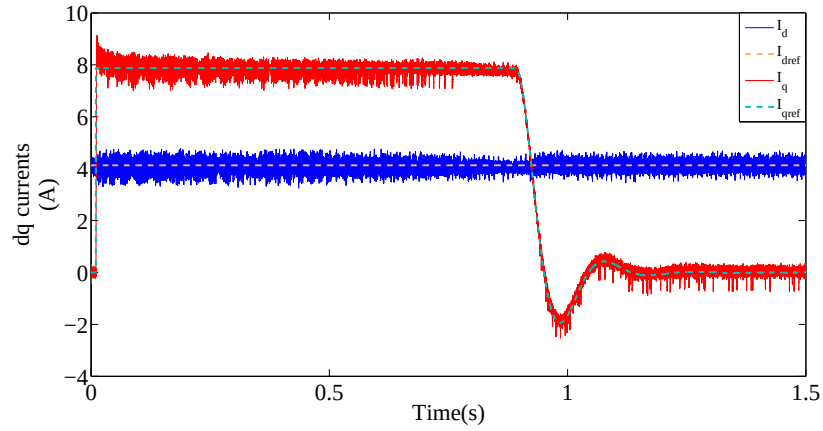


Figure 5.14: Reference tracking of dq stator currents

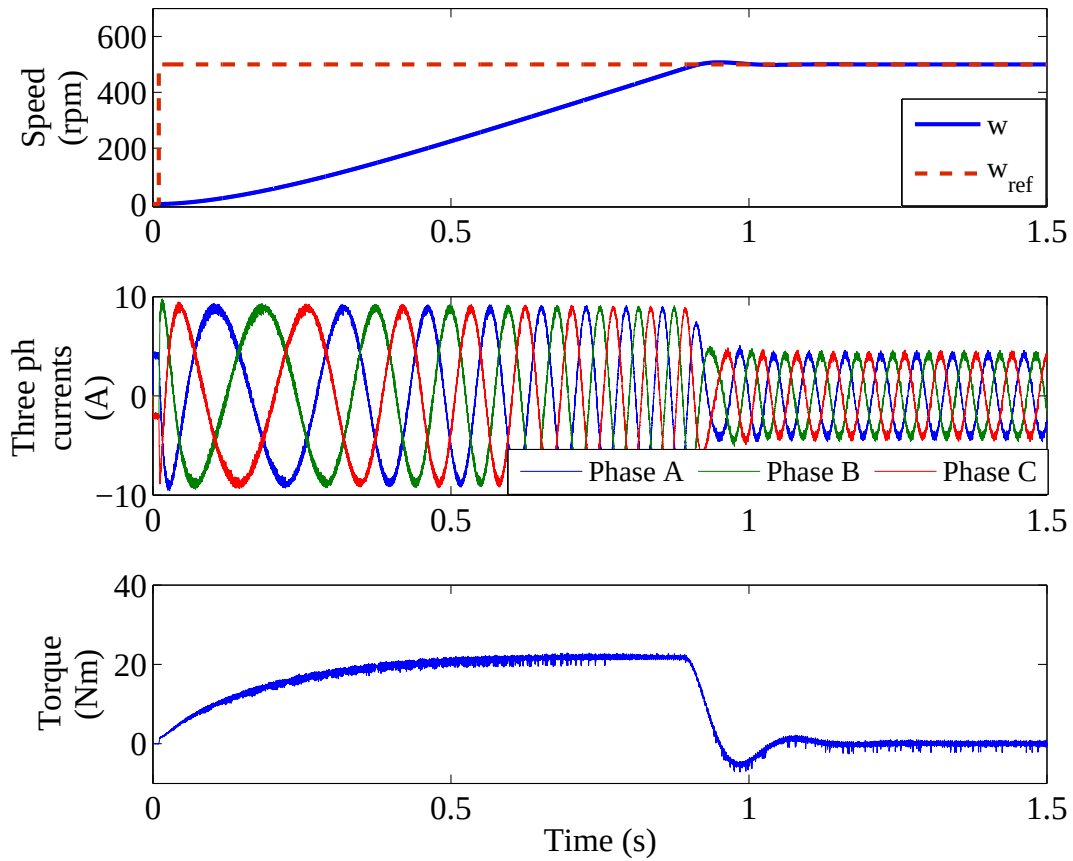


Figure 5.15: Transient response of the machine at no load : Speed, Three phase stator currents and torque during a speed demand of 500rpm

Figure 5.16 shows the performance of the drive when step change in speed from 200rpm to 500rpm is applied at $t=1.2\text{s}$. This shows the fast transient response of the controller when there is a sudden change in the reference signal which is a characteristic of MPC. The quality of the currents at higher modulation index is however better than that of a drive controlled using MPC due to the modulation scheme involved.

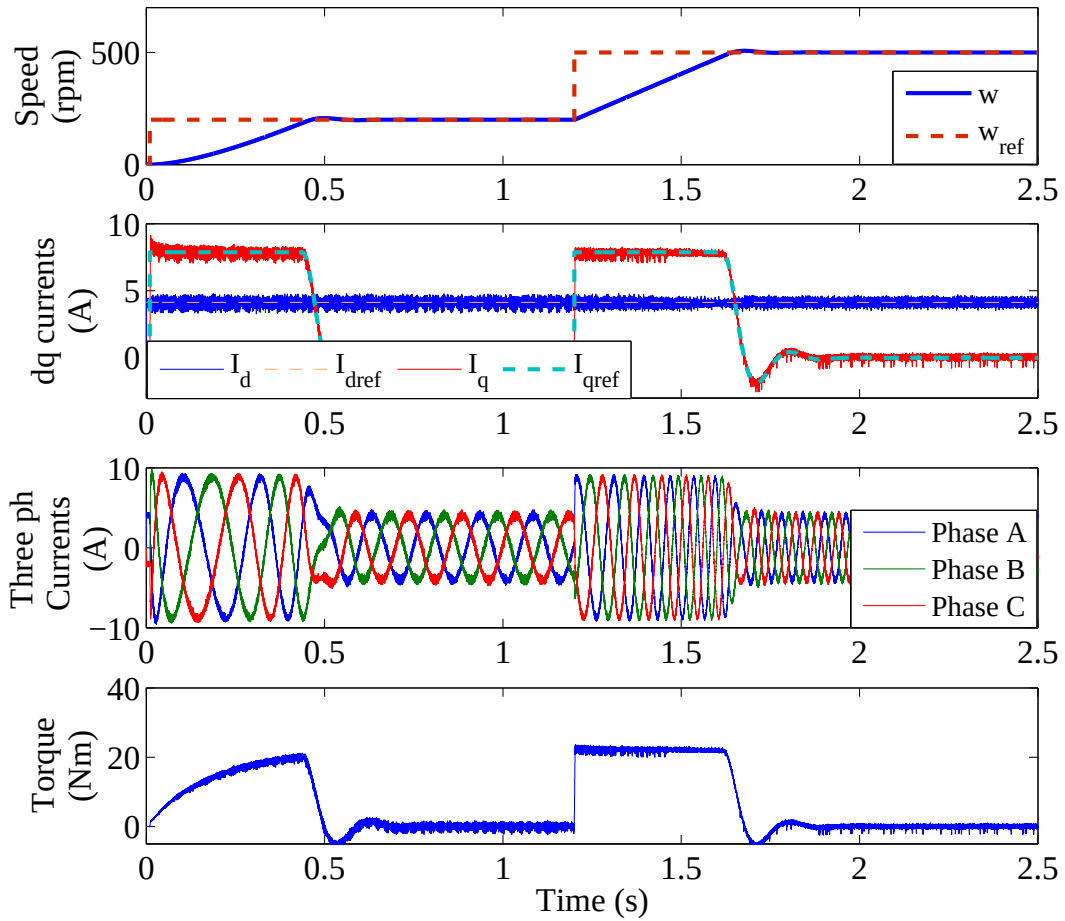


Figure 5.16: Speed, Three phase stator currents and Torque during step change in the speed of the machine at no load

5.5.2.1 Multi-objective M2PC

One of the positive features of the M2PC is the ability to control more than one control variable simultaneously without the use of cascaded loops. To analyze the

performance of the strategy for mutli-objective control, a matrix converter feeding an IM is considered where both speed and stator current control is achieved by the M2PC. A control block diagram that depicts the procedure is shown in Figure 5.17.

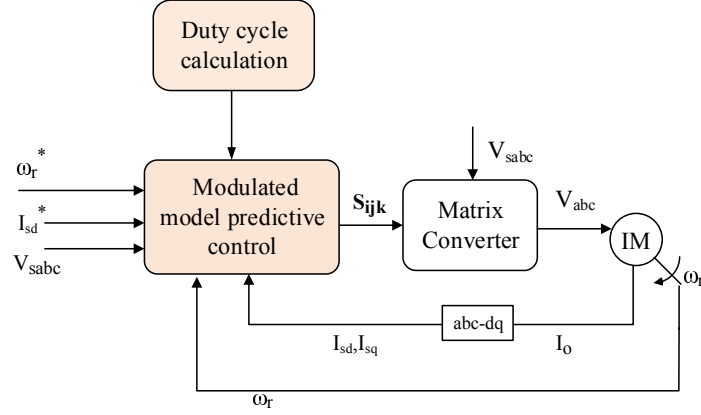


Figure 5.17: Control block diagram of M2PC for matrix converter drive

To demonstrate the multi-objective control capability of the M2PC, control of the speed of the induction machine is also included in the M2PC cost function, thereby eliminating the need for a PI speed controller. The speed of the machine at the future instant is predicted from the torque equation of the machine as shown in 5.21.

$$T_e = T_{load} + J \frac{dw_r}{dt} + B_v w_r \quad (5.21)$$

Rearranging 5.21 and applying forward Euler approximation, the predicted value of speed can be obtained as shown in 5.22.

$$w_r(k+1) = w_r(k) + T_s(T_e - T_{load} - B_v w_r) \quad (5.22)$$

The electromagnetic torque, T_e can be expressed in terms of the stator currents as shown in 5.23. It can be noted that this equation is accurate only during the steady state operation of the machine and does not account for the transients. This is due to the assumption that the flux in the machine remains constant which is applicable only in steady state.

$$T_e = \frac{3}{2} \frac{P}{2} \frac{L_m^2}{L_r} I_{sd} I_{sq} \quad (5.23)$$

The reference for the torque producing current can be obtained by substituting 5.23 in 5.22 as shown in 5.24.

$$I_{sqref} = \frac{JL_r(w_{ref} - w_r)/T_s + B_v w_r + T_{load}}{3L_m^2 I_{sd}} \quad (5.24)$$

The cost function for the speed controller is the quadratic error between the reference speed and the predicted speed and is expressed as shown in 5.25.

$$G_w = (w_{ref}(k+1) - w_r(k+1))^2 \quad (5.25)$$

By merely adding the cost function for speed control given by 5.25 to the current control cost function, G_{curr} together with appropriate weighting factor, the control of speed and stator current control can be achieved without cascaded loops. A weighting factor determines the emphasis on each of the control variables. A higher weighting factor for one variable indicates that priority is given for that variable compared to the another variable in the same cost function. The choice of weighting factor is often done by trial and error and is still a debatable topic in the area of predictive control as there is no concrete procedure to calculate it. The final cost function for the simultaneous control of both speed and stator currents of an induction machine is given by 5.26.

$$G_{final} = G_{curr} + A_w * G_w \quad (5.26)$$

where A_w is the weighting factor for speed control. In this case, the weighting factor is determined by fixing the co-efficient of G_{curr} to 1 and changing the A_w until the desired response is achieved. The weighting factor that has been used in this simulation is $A_w = 0.8$.

5.5.2.2 Simulation Results

Simulation of the M2PC strategy to control both speed and stator currents of an induction machine fed by a direct matrix converter is performed in Matlab Simulink

environment. The simulation results when a speed reference of 500rpm is demanded from the system is shown in Figure 5.18. As seen from Figure 5.18, the speed, dq-axis and three phase stator currents follows the reference with negligible steady state error. This indicates that the need for a separate PI controller to control the speed of the machine can be eliminated by including the speed error in the cost function. Thus it can be inferred that the M2PC method is capable of multi-objective control without using nested loops.

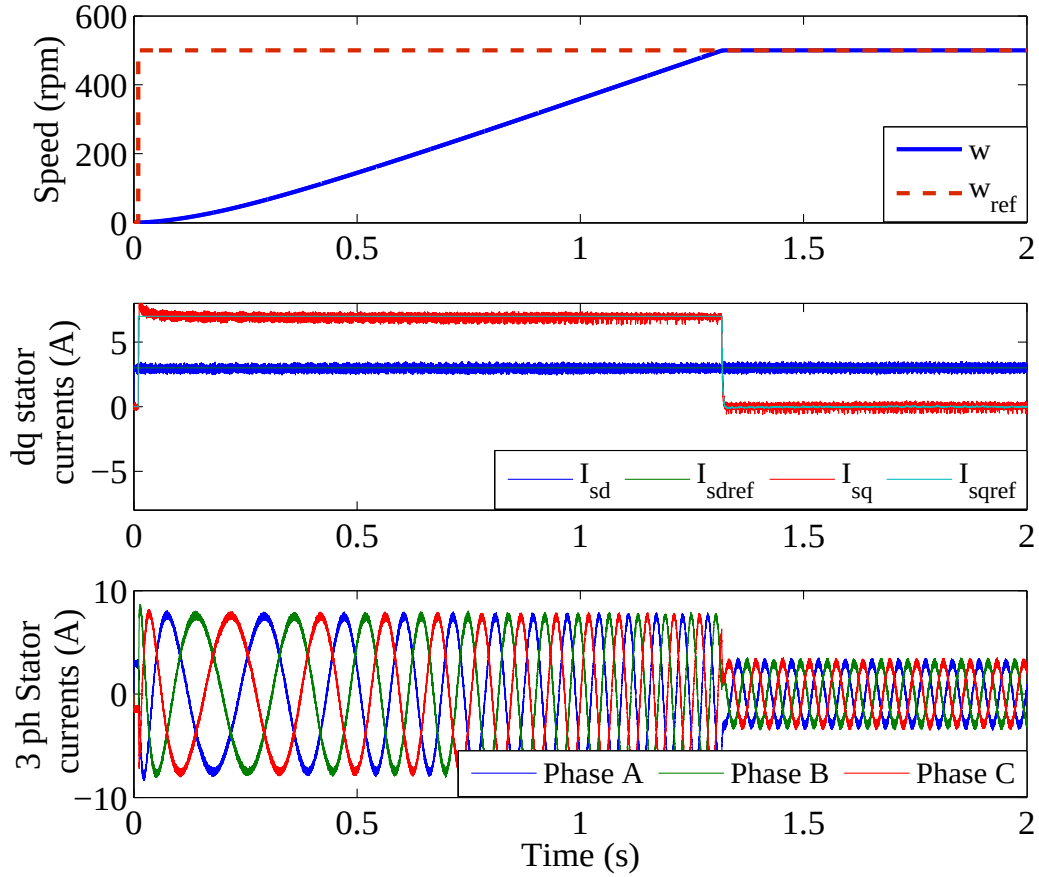


Figure 5.18: Speed, Three phase stator currents and torque during step change in the speed of the machine.

5.6 Limitations of M2PC

From the simulation results, it can be noted that the key advantages of a predictive controller like fast transient response holds true for the M2PC. However the steady state performance of this strategy is not comparable to the results from PI controller with SVM in section 3.3.6. There could be two reasons for this difference. Firstly, the M2PC algorithm discussed in this chapter only considers the control of load currents while the input current remains uncontrolled. Since a modulator like SVM ensures the control of both input and output currents, a fair comparison of these two methods cannot be made. On the other hand, the results from M2PC with an RL load are largely improved compared to that from MPC which was also the motivation behind the M2PC method.

Even though the simulation results shown in this chapter are promising, the procedure to derive the duty cycles based on the cost function magnitudes has been identified as an approximation. It is mentioned in [62] that at high modulation index the converter waveforms were distorted for a 7-level cascaded H-bridge converter. To address this issue, a new approach is considered that takes into account the magnitude and direction of the errors/cost functions. This would give an accurate value for the duty cycles. This new control strategy is elaborated in the section 5.7.

5.7 Direct Predictive Current-error Vector Control

A second control strategy named the Direct Predictive Current-error Vector Control (DPCVC) based on predictive control and including a modulation scheme is developed to overcome the drawbacks of M2PC while maintaining the good features like fast dynamic response. This method is based on the principle proposed in [86] for a three phase 2-level inverter. The fundamental idea is to consider the errors/cost function

vectors in $\alpha\beta$ plane to calculate the duty cycles or application times for the voltage vectors. In M2PC, only the magnitude of the cost function is considered, resulting in a approximate solution for duty cycles.

The basic principle of DPCVC is based on the error vectors of the control variable formed by each switching state of the converter. For example, for load current control, the error vectors when plotted in $\alpha\beta$ plane results in a skewed hexagon similar to that of the voltage vectors for each switching state. A matrix converter has 27 switching states out of which six produce rotating vectors that changes in magnitude and direction. Similar to the standard SVM technique, rotating vectors are not considered for this modulation strategy; resulting in 21 switching states. To simplify the problem and as a starting point, it is decided to modulate between two active and one zero vector for the matrix converter. This is sufficient to control the load current of MC. If input current control is required, two more active vectors need to be considered. The voltage and input current vectors for a matrix converter for each switching state is shown in Figure 5.19.

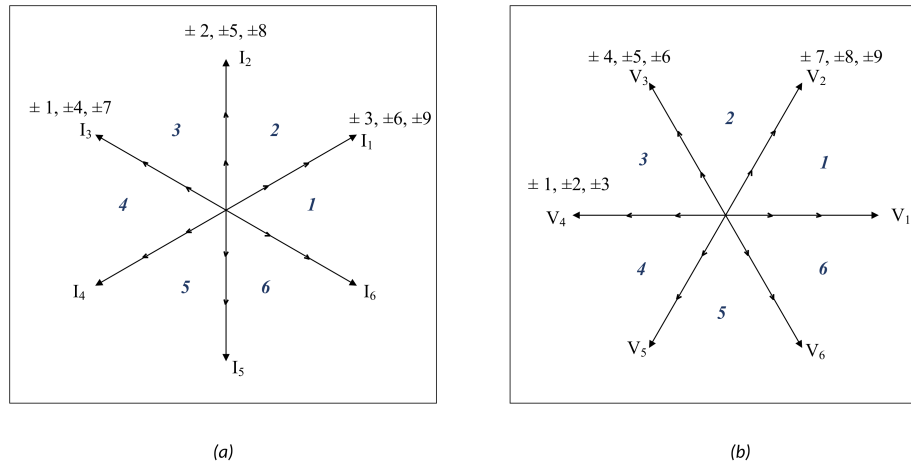


Figure 5.19: (a) Input current and (b) Output voltage vectors of a MC

Once the desired input current sector is determined from the supply voltages, the number of available voltage vectors/switching states for modulation can be narrowed down to 12. For example, in Figure 5.19, if the desired input current sector is 1, then the voltage vectors that can produce an input current in that sector are $\pm 3, \pm 6, \pm 9, \pm 1, \pm 4$ and ± 7 . These switching states produce vectors of varying am-

plitude and direction as shown in Figure 5.20.

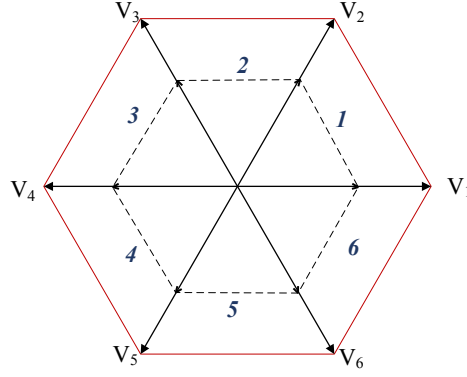


Figure 5.20: Output voltage vectors of a MC

Out of the 12 voltage vectors available, the voltage vectors with higher magnitude forming the outer hexagon (shown in red) is considered for modulation. In order to select the two active vectors to modulate, the output current error vectors for each switching state is plotted. The output current error vectors for the six voltage vectors in the outer hexagon and zero vectors is shown in Figure 5.21. It is evident from Figure 5.21 that the predicted current errors form a hexagon in the $\alpha\beta$ plane.

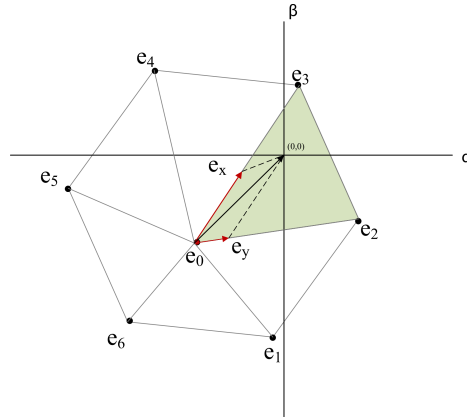


Figure 5.21: Output current error in $\alpha\beta$ plane

Based on predictive control algorithm, the objective is to minimize or obtain zero load current error. Therefore the target is the origin of the plane. The control problem is then to find the linear combination of three vectors which will result in zero load current error. In Figure 5.21 this is obtained by the linear combination of the error vectors e_2 , e_3 and e_0 in a certain proportion. A solution exists only if the

target (origin) lies within the triangle formed by the two consecutive vectors. If the target lies outside the triangle, it is considered as the over modulation condition and different measures need to be taken to address it. The relevant triangle or consecutive set of vectors can be obtained from the following conditions mentioned in [86] for each consecutive set of vectors. For every pair of consecutive vectors like (1,2), (2,3), (3,4), (4,5), (5,6), (6,1) a linear combination of e_x , e_y and e_0 exists if the following conditions are met.

$$\begin{aligned} ((e_x - e_0) \times (-e_0) \cdot (e_y - e_0) \times (-e_0)) &\leq 0 \\ \wedge (e_x - e_0) \cdot (-e_0) &> 0 \\ \wedge (e_y - e_0) \cdot (-e_0) &> 0 \end{aligned} \quad (5.27)$$

where $\langle \cdot \rangle$ is the dot operator and $\langle \times \rangle$ is the cross operator. The pair of vectors that satisfy the conditions in 5.27 is selected for modulation. The linear combination of vectors to obtain the target can then be obtained by solving the set of linear equations in 5.28.

$$\begin{aligned} (e_{x\alpha} - e_0)d_1 + (e_{y\alpha} - e_0)d_2 &= -e_0 \\ (e_{x\beta} - e_0)d_1 + (e_{y\beta} - e_0)d_2 &= -e_0 \\ d_1 + d_2 + d_0 &= 1 \end{aligned} \quad (5.28)$$

where e_x and e_y are the two current error predictions that satisfy 5.27 and d_1 , d_2 and d_0 are the duty cycles for each vector. If $d_1 + d_2 > 1$ that means the target point is outside the hexagon formed by the error vectors and the duty cycles are given by 5.29. In this case, the target point is achieved by modulating between the two active vectors only.

$$\begin{aligned} d_1 &= \frac{d_2}{d_1 + d_2} \\ d_2 &= \frac{d_1}{d_1 + d_2} \\ d_0 &= 0 \end{aligned} \quad (5.29)$$

The resulting voltage vector will be a combination of the two active vectors applied

for their respective duty cycles and can be expressed as in 5.30.

$$v_o = d_1 v_x + d_2 v_y \quad (5.30)$$

The reference voltage vectors is then synthesized by the standard SVM technique for MC using two active and three zero vectors. The double sided switching pattern for MC in this case is shown in Figure 5.22.

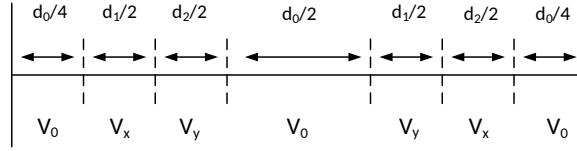


Figure 5.22: Double sided switching pattern for MC

This strategy generates accurate duty cycles depending on the instantaneous error predictions and results in fixed switching frequency operation. Due to the predictive control based algorithm, this modulation strategy will also have very fast transients. However it is worth noting that for simple systems that only requires a current control, DPCVC will be very similar to that of a dead-beat controller. Hence in such cases, a dead-beat controller is a better option compared to DPCVC as it requires less computational effort.

5.7.1 Simulation Results

To validate the theory explained in the previous section, the DPCVC is used to control the load current of a direct matrix converter feeding a RL load and an induction machine. The control block diagram showing the steps involved is shown in Figure 5.23.

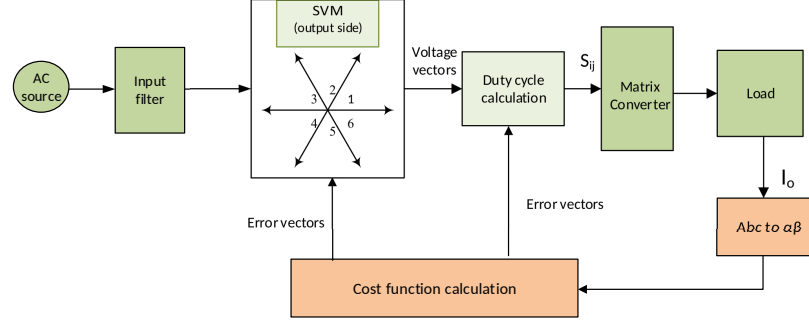


Figure 5.23: Control block diagram for Direct Predictive Current-error Vector Control

A reference current of 5A at 30Hz is demanded from the system and the resulting three phase load currents, corresponding MC line voltage and FFT of Phase A load current are shown in Figure 5.24.

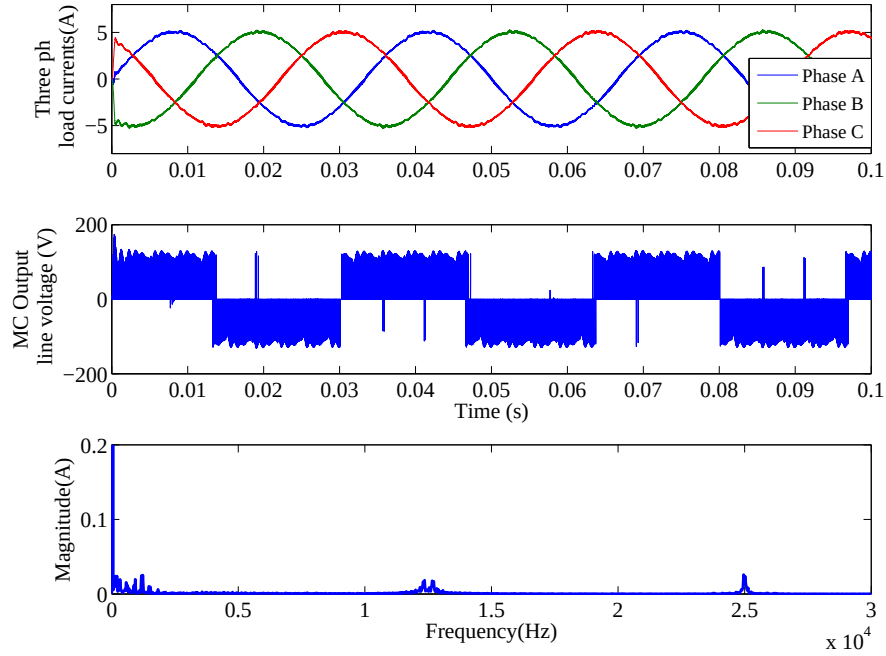


Figure 5.24: Controlled 3-phase load currents and MC output line voltage of a MC feeding a RL load

Since DPCVC is used to implement load current control, the input current of the converter is uncontrolled. This results in waveforms as shown in 5.25.

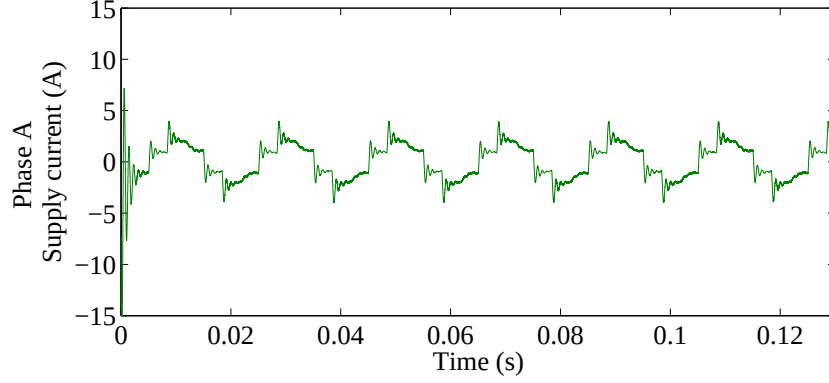


Figure 5.25: Phase A supply current of MC

From Figure 5.24, it is evident that the load currents attain the steady state value without any steady state error. The FFT spectrum shows the presence of harmonics in the range of 12.5kHz and its multiples which is the switching frequency. This proves the fixed switching frequency operation. It can be noted that the waveform quality of this method is improved compared to M2PC. The THD of the load current waveforms controlled using DPCVC is approximately 2.9 % which is a considerable improvement compared to that of MPC or M2PC. This proves that the calculation of duty cycles in terms of the current errors in alpha-beta plane is more accurate than merely considering the magnitude of the errors as in the case of the M2PC.

To demonstrate the fast dynamic response of the DPCVC method, a step in the current demand from 2A to 4A is applied to the converter and the response of the DPCVC is shown in Figure 5.26. It can be noted that the response of the controller to change in the magnitude of the reference is fast.

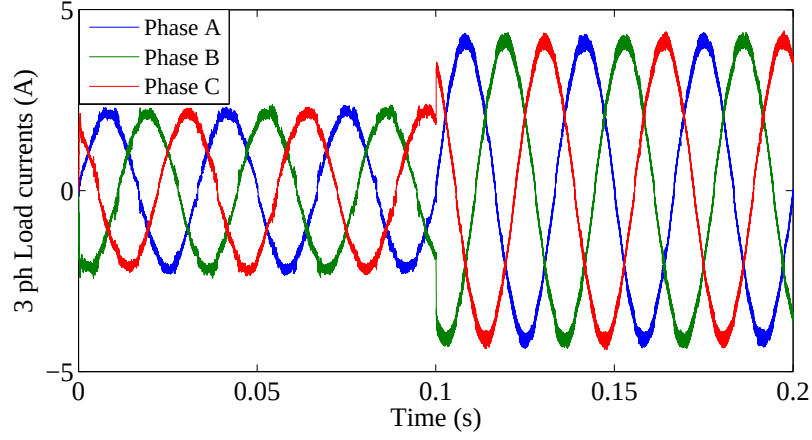


Figure 5.26: Dynamic response of the DPCVC strategy when a step demand in current is applied to the system

To study the performance of this method under different loading conditions, an induction machine is considered as the load. The speed of the machine at no load is controlled by a PI controller and the stator currents are controlled using DPCVC technique. A speed demand of 500rpm is applied to the system and the controlled speed and dq-axis stator currents are shown in Figure 5.27. From Figure 5.27, it is evident that control variables track the references without any steady state error. The controlled stator currents have a THD of around 2.4%. This indicates that DPCVC is capable of delivering good quality controlled currents very similar to the case when controlled by a PI controller with SVM.

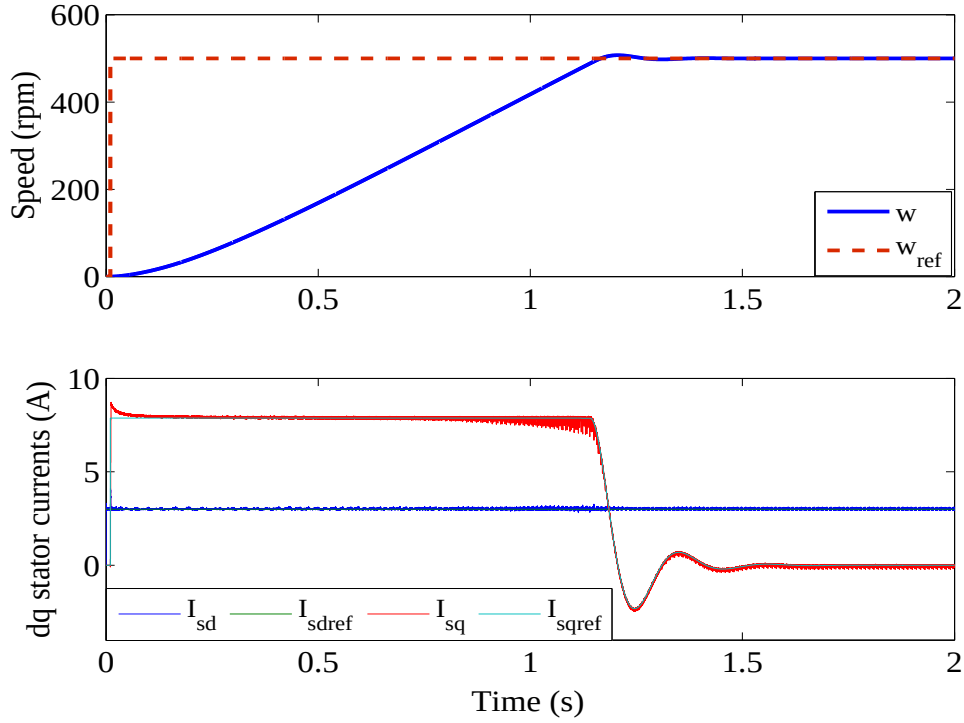


Figure 5.27: Speed and dq-axis stator current reference tracking during a speed transient of 0-500rpm at no load

5.8 Summary

Two control methods with the features of MPC and a modulation scheme similar to SVM are proposed in this chapter for a direct matrix converter. The ability to control different parameters simultaneously, high controller bandwidth and constant switching frequency are the main highlights of these methods. The procedure for implementing both these control methods are discussed in detail in this chapter. Implementation of M2PC for a direct matrix converter is done in Matlab Simulink environment and the resulting waveforms are also included in this chapter. From the results it is evident that the quality of the currents is improved when using M2PC compared to MPC without losing its advantages like fast transient response and multi-objective controllability.

Another method that provides fixed switching frequency operation is the DPCVC

strategy which is proposed for load current control of a direct matrix converter. The simulation results of this method for a RL load and induction machine are promising and shows improved waveform quality compared to the M2PC and MPC.

Chapter 6

Experimental Validation

6.1 Introduction

This chapter focuses on the implementation of a direct matrix converter laboratory setup to validate the control strategies presented in this thesis namely PI control, MPC, M2PC and DPCVC. The experimental setup consists of the converter hardware, control hardware and the control software which are discussed in section.6.2. Experimental tests are carried out for all the control strategies mentioned in chapters 3, 4 and 5 for a direct matrix converter feeding an RL load and an induction machine. The tests are done keeping the parameters of the load and input filter constant. The tests conducted experimentally are the same as that of simulation for the purpose of comparing the results. Outcomes of the experimental tests showing the controlled waveforms are discussed in section.6.3. An analysis of the experimental results and the conclusions inferred from them are also discussed in this chapter.

6.2 Description of the experimental setup

The experimental setup of a direct matrix converter system has been used to implement the control methods mentioned in chapters 3, 4 and 5. A block diagram showing the different components of the setup is shown in Figure 6.1. The setup includes an input filter, matrix converter power circuit, the load and a control platform. Each of these components will be explained in detail in this section.

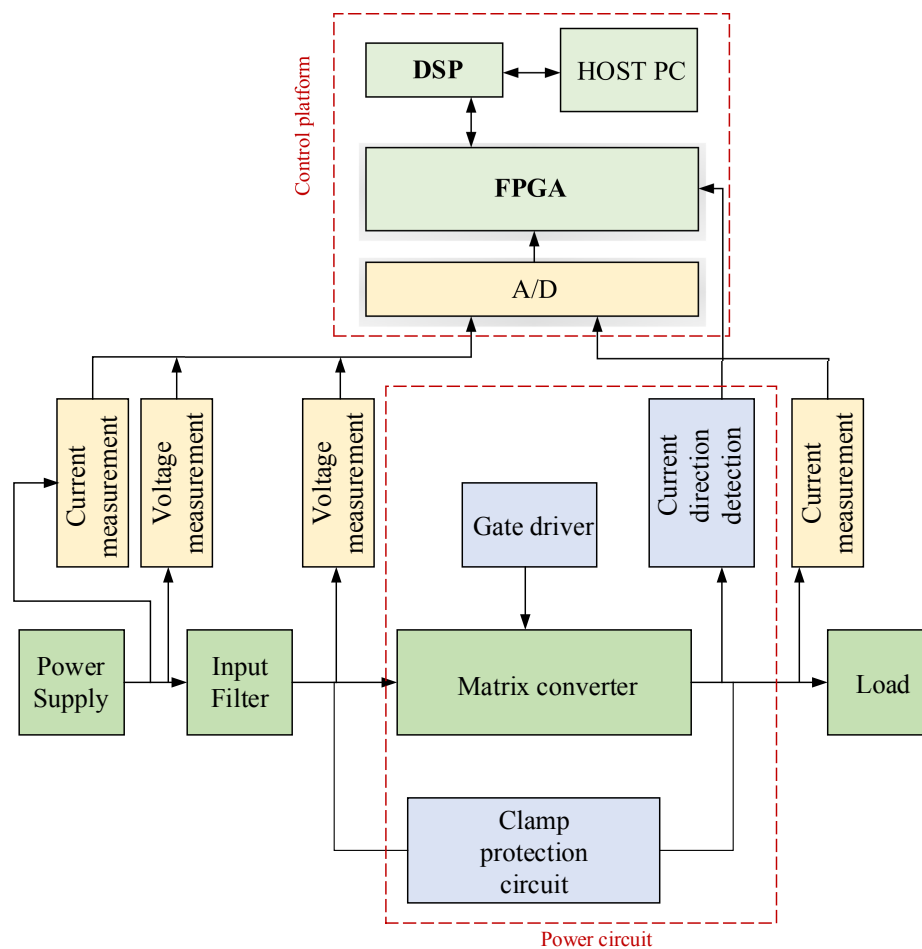


Figure 6.1: Block diagram of the experimental setup

6.2.1 Input Filter

As mentioned in chapter 2, an input filter is required to prevent harmonic injection into the grid. The harmonics are caused by the switching of the matrix converter and it is important to filter the distorted source currents. An LC filter with a damping resistor is used in this setup to satisfy the filtering requirements of the circuit.

The input filter values are selected based on the switching frequency and the supply frequency of the system. If a supply frequency of 50-60Hz and switching frequency of 12.5kHz is considered, a resonant frequency of 1kHz is assumed to avoid interference with the supply and switching frequencies and their harmonics.

The values of the input filter components calculated based on the above criteria are shown in Table 6.1

Table 6.1: Input filter parameters

Parameter	Value	Unit
<i>Filter inductance, L_f</i>	3.75	<i>mH</i>
<i>Filter capacitance, $C_f(\text{delta})$</i>	8.3	μF
<i>Damping resistor, R_d</i>	15	Ω

A photograph of the input filter used in the experimental set up is shown in Figure 6.2.



Figure 6.2: Photograph of input filter of the matrix converter

6.2.2 Power Circuit

The circuit schematic of the direct matrix converter system is shown in Figure 6.3. It shows the different stages of a typical MC system such as the supply, input filter, MC power circuit, Clamp protection circuit and the load.

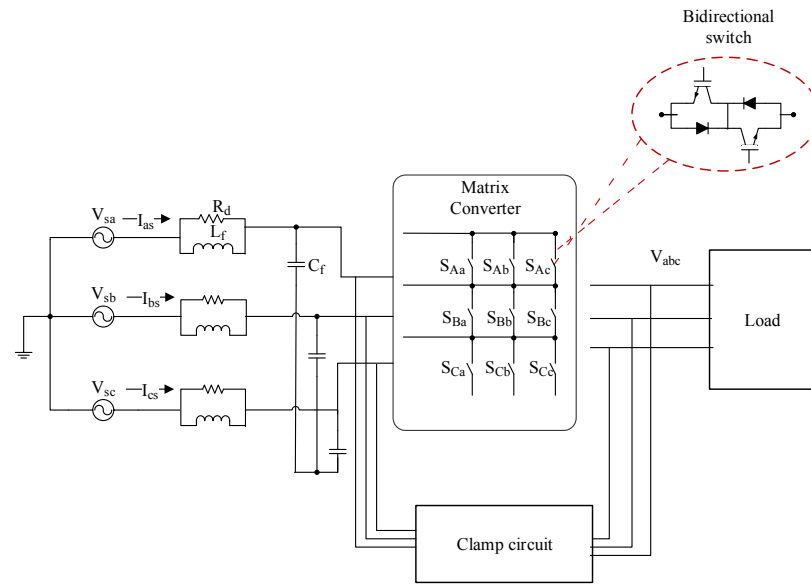


Figure 6.3: Schematic of MC power circuit with supply and load

A photograph of the power circuit is shown in Figure 6.4. The matrix converter power circuit consists of the Power modules, the gate drivers, clamp protection circuit and the current direction detection circuit. Each of these parts will be discussed in detail in the following subsections.

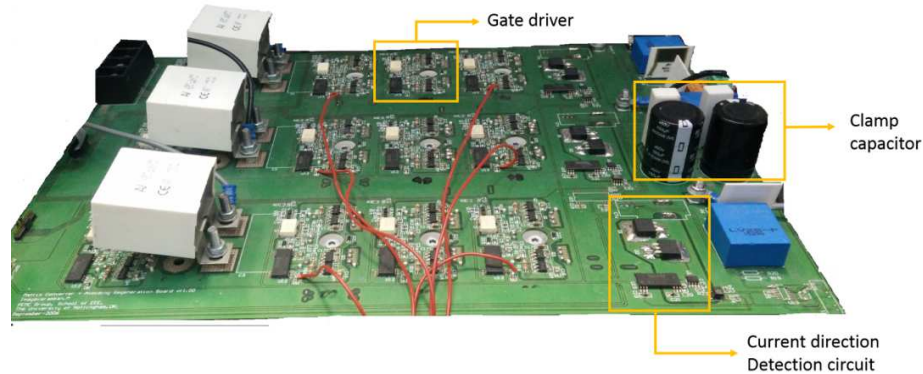


Figure 6.4: Photograph of the power circuit

6.2.2.1 Power modules

The matrix converter is formed of nine bidirectional switches implemented with IGBT modules. The IGBT modules used in the setup are the Semikron SK60GM123 [115] (see Figure 6.5) and they are arranged in the form of a matrix to form a three-phase to three-phase matrix converter. The modules are rated at 1200V and 60Amps. Each IGBT module consists of two anti-parallel IGBTs connected in common emitter configuration together with two diodes.



Figure 6.5: Semikron power module

The switching signals generated from the control platform cannot directly be used to turn on and off the power devices in matrix converter. These signals need to be amplified with the help of gate drivers. The IGBT is a voltage-controlled device and to change the voltage at the gate, a significantly large current is required which is

provided by the push-pull amplifier stage after the gate driver chip. The gate driver is responsible for providing +15V or -15V between the gate and emitter of the IGBT when the switching command state of '1' or '0' is sent from the Field Programmable Gate Array (FPGA), respectively. The gate driver circuit also provides good isolation between the low power control platform and the power devices.

6.2.2.2 Clamp protection circuit

A clamp circuit is essential in a matrix converter for protection against over voltages due to voltage disturbances or open circuit faults. A schematic of a typical clamp circuit is shown in Figure 6.6.

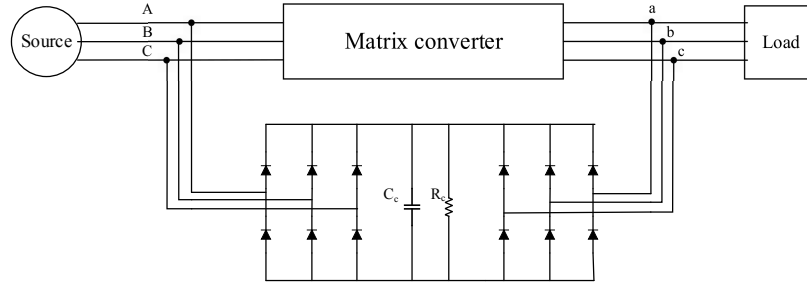


Figure 6.6: Schematic of the clamp protection circuit

The circuit consists of 12 fast recovery diodes (8EWF12S), two clamp capacitors and two discharge shunt resistors. These diodes are rated at 1200V and 8A. Two $150\mu F$, 450V electrolytic capacitors are connected in parallel with $47k\Omega$ balancing resistors to distribute the voltage evenly across these capacitors. In the event of a fault, the diode bridge will provide a path for the current and to transfer the inductive load energy to the clamp capacitors, which in turn is dissipated in the shunt resistors. The selection of the clamp capacitor is vital as the inductive load energy has to be stored in the capacitor without creating dangerous over-voltage.

The clamp capacitor value can be calculated from the maximum energy stored in the inductive load and the maximum peak to peak source voltage. The energy stored in

a three phase inductive load carrying balanced currents is given by 6.1.

$$\begin{aligned} Q &= \frac{1}{2}L(i_a^2 + i_b^2 + i_c^2) \\ &= \frac{3}{4}LI^2 \end{aligned} \tag{6.1}$$

where Q is the total energy stored in the load inductor, L is the load inductance and I is the sum of three phase load currents. The value of the clamp capacitor can then be calculated from 6.2.

$$\begin{aligned} C &= \frac{2Q}{V_i^2 - V_{cmax}^2} \\ &= \frac{3/2LI^2}{V_{cmax}^2 - V_i^2} \end{aligned} \tag{6.2}$$

where V_i is the initial clamp voltage and V_{cmax} is the maximum clamp capacitor voltage.

If a line to neutral voltage of 200V is considered, the clamp capacitor voltage initially will be 490V. The maximum voltage is fixed at 700V and load inductance is 3.75mH. Thus using (6.2), the value of clamp capacitance is calculated and it resulted in 71.95 μF . Since this value is not available commercially, 75 μF is chosen in this implementation. The energy stored in the clamp capacitors is then discharged by the resistors connected in parallel.

6.2.2.3 Data acquisition

The measurement of currents and voltages required for the implementation of the control strategy are obtained from the respective transducers. The different transducers used in this setup are four voltage transducers (LEM LV 25) [116] to measure the supply line voltages and the line voltage after the input filter and five current transducers (LEM LA 55)[117] to measure the supply and load currents. Both the current and voltage transducers output a current proportional to the primary signal which can be voltage or current. This signal is then converted to a voltage at the

FPGA interface with the help of burden resistors.

The voltage transducers (LEM LV 25) are rated at 500V and the connection diagram of the same is shown in Figure 6.7. As seen from Figure 6.7, the transducer needs $\pm 15V$ supply and a resistor on the primary and secondary side. The resistor on the primary side, R_b is used to limit the current through the device to 10mA and the secondary side resistor (burden resistor), R_m converts the current at the output of the transducer back to voltage on the FPGA card.

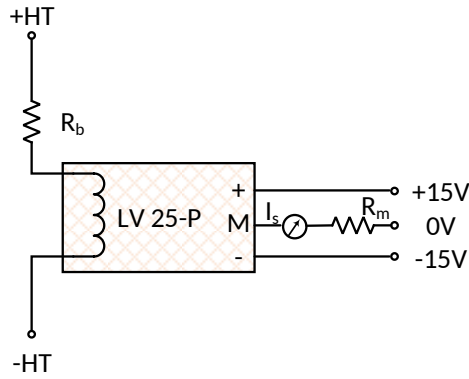


Figure 6.7: Connection diagram for voltage transducer

The current transducer (LEM LA 55) has a rated current of 50A. A connection diagram for the current transducer is shown in Figure 6.8. These transducers requires $\pm 15V$ supply and a burden resistor (R_m) to convert the current measured to voltage at the FPGA board.

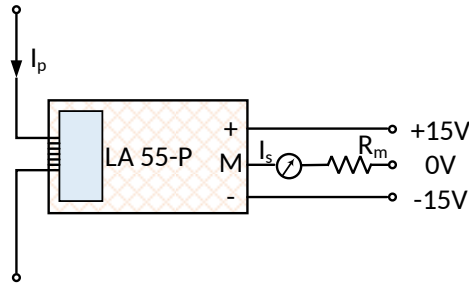


Figure 6.8: Connection diagram for current transducer

A photograph of the sensor board is shown in Figure 6.9.

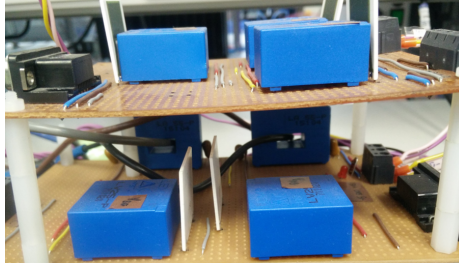


Figure 6.9: Photograph of voltage and current transducers

In addition to the current and voltage transducers, an encoder is used to get the instantaneous rotor position of the induction machine. Speed of the machine is essential as a feedback signal for vector control, MPC, M2PC and DPCVC of induction machines. Figure 6.10 shows the photograph of the encoder mounted on to the shaft of the motor.



Figure 6.10: Photograph of encoder mounted on the motor shaft

The encoder used in this experimental setup is a Model 755RG High Precision servo mount encoder[118] manufactured by British Encoders. It operates with a 5V supply and a frequency of upto 1MHz. It is a differential incremental encoder with three output channels A, B and Z. Channel A determines the rotor speed while channel B provides information about the direction of rotation. Channels A and B are square waves positioned 90° apart. For example if A leads B, then the rotor is rotating in clockwise direction and if B leads A, it is rotating in anticlockwise direction. Thus it is possible to track both the position and direction of rotation by observing both the number of pulses and the relative phase between A and B. The Z channel also called zero channel gives the precise position of the encoder shaft. Figure 6.11 shows the graphical representation of the encoder signals.

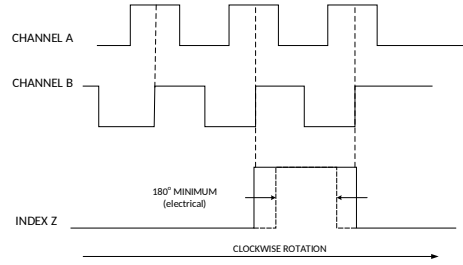


Figure 6.11: Encoder output signals

6.2.2.4 Current direction detection circuit

As described in chapter 2, the four-step commutation requires the current direction of each output phase. Also to ensure a safe commutation between the switches, it is vital that the FPGA gets this information. The current direction detection circuit consists of two anti-parallel Schottky diodes (30CTQ045S) [119], a voltage comparator (LM311D)[120], and an opto-coupler (HCPL-061N)[121]. A schematic of the current direction detection circuit is shown in Figure 6.12.

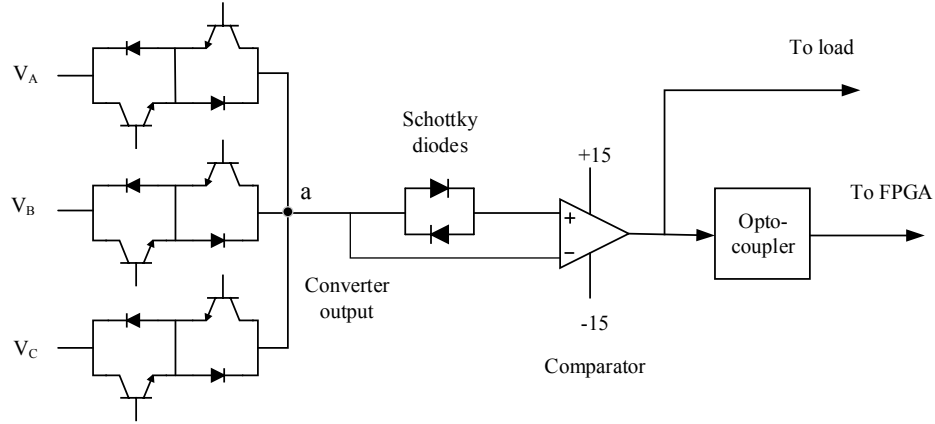


Figure 6.12: Schematic of the current direction detection circuit for output Phase a

As shown in Figure 6.12, there are two anti-parallel Schottky diodes with a current rating of 30A connected in series with the output of each phase of the matrix converter. These diodes determine the current direction at each output phase. As they have very low forward voltage drop, these diodes are perfect for detecting the output current direction quickly and accurately in the most efficient manner. A comparator with the voltage drop across the diodes as inputs provides the gate signal to the FPGA.

To allow voltage isolation and noise immunity, an opto-coupler is introduced between the output of the comparator and the FPGA. The circuit is supplied by an isolated dc/dc power converter (DCP020515). Figure 6.13 shows the photograph of the current direction detection circuit on the power board.

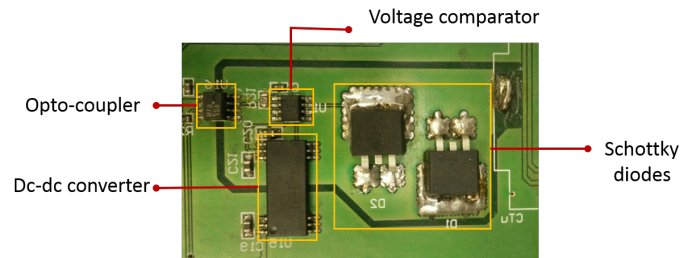


Figure 6.13: Photograph of current direction detection circuit

6.2.3 Control Platform

As shown in Figure 6.1 the control platform consists of the control circuit (DSP and FPGA), HPI interface circuit and encoder interface board. The DSP board used is by Texas Instruments and the FPGA board is designed by Dr. Lee Empringham at Power Electronics Machines and Control (PEMC) Research Group, The University of Nottingham. A host port interface (HPI) card is used to access the DSP memory and allow bidirectional data transfer between the host PC and the DSP.

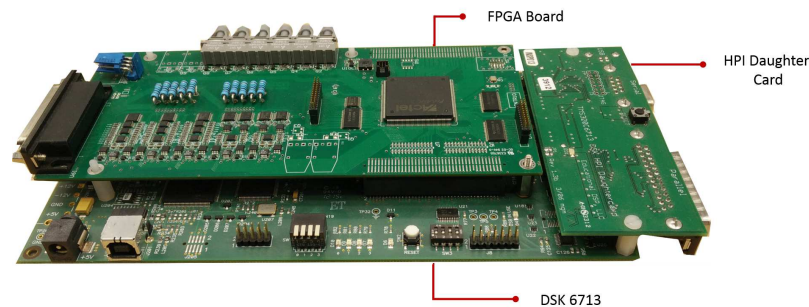


Figure 6.14: Control platform with DSP, FPGA and HPI daughter board

6.2.3.1 DSP board

The control algorithm is implemented in a Texas Instruments 32 bit, Digital Signal Processor (TMS320C6713) [122] which operates at 225MHz. A parallel port interface of C6713DSK connects the DSP to the HPI interface. The internal memory of the DSP can be accessed without interrupting the CPU by means of the HPI port. This enables bidirectional data transfer between the host PC and the DSP. This is essential to control the matrix converter and download sampled data for analysis.

The software implementation of control algorithm in DSP is programmed in C language and compiled using Code Composer Studio from Texas Instruments. Figure 6.17 shows a flowchart that summarizes the software implementation of the control algorithm. The whole algorithm can be divided into three routines: initialization routine, the main loop and the interrupt service routine (ISR). Initialization routine initializes the timer, FPGA, the host parameters and the controller parameters. Until the interrupt signal is detected the program will run the main loop. Once the interrupt signal is detected, ISR starts the A/D conversion of the measurement signals.

The software trips for over current and over voltage protection circuit are also included in the ISR. The measured values of current and voltages are compared with the pre-set trip values. The converter will be disconnected if the measured values exceed the pre-set values for trip. Else the program continues to implement the control strategy and outputs the switching signals to the converter.

6.2.3.2 FPGA card

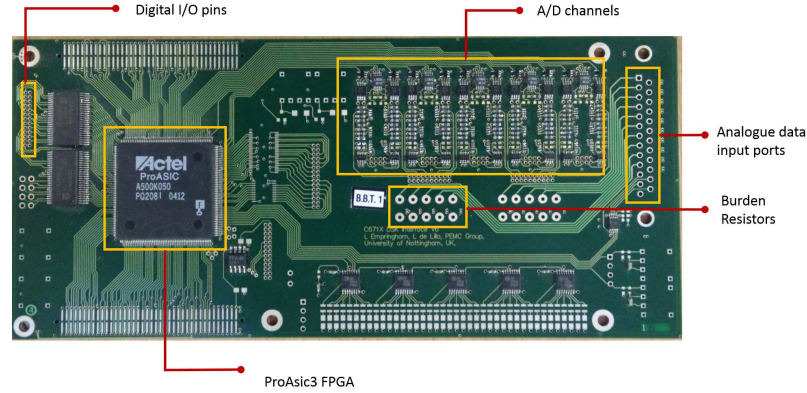


Figure 6.15: FPGA board

The FPGA card (see Figure 6.15) consists of an ACTEL ProASIC3 FPGA module [123], digital input-output ports, analog to digital (A/D) converters and analogue input and analogue output. It has ten 14-bit analog to digital (A/D) converters with signal conditioning units for system data acquisition. The control code for the converter is triggered by an interrupt signal in the DSP. The interrupt signal is given by the FPGA and the interrupt routine carries out the control and modulation and sends the switching signals back to the FPGA.

The implementation of control strategies like MPC and M2PC requires information about the output phase current and voltage, input phase voltage and the phase voltage at the input filter capacitor. The current and voltage transducers mentioned in section 6.2.3 feeds the measurement signals to the FPGA through the A/D channels. The reference signals on the other hand are demanded by the user by means of the HPI daughter card which is the interface between the user and the control platform.

The FPGA generates the interrupt signals for the DSP which initiates the interrupt routine. The FPGA also includes a watchdog timer to ensure continuous communication between the DSP and FPGA. Every time the timer receives a signal from the DSP, it will reset. This signal is sent at the beginning of every interrupt routine. If the timer fails, it means that there is a miscommunication between the DSP and FPGA and the converter will trip. This ensures that the converter remains protected

if there is a communication failure between the DSP and the FPGA card.

6.2.3.3 Encoder Interface Board

Measurement of speed of the rotor is necessary for closed loop control of the motor. The differential signals from the encoder are fed into the interface board, which then directs three channel signals A, B and Z to the FPGA. The interface board consists of a quadruple differential line receiver (SN75173), a digital isolator(opto-coupler)(HCPL090) and an isolated dc/dc converter(NML0505SC). A photograph of the encoder interface board showing different components is shown in Figure 6.16.

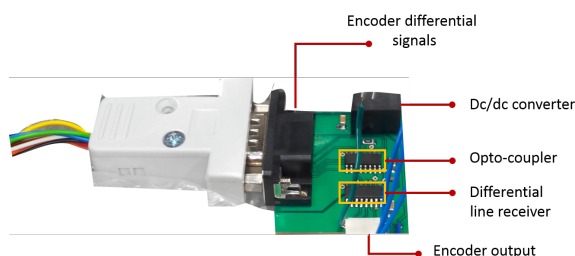


Figure 6.16: Encoder interface board

The dc/dc converter which generates a single output 5V from the 5V isolated power supply powers the opto-coupler. The differential line receiver obtains the differential signals from the encoder and generates the three channel encoder output signals, which is then fed to the FPGA through the digital isolator.

A flowchart explaining the algorithm of the control software is shown in Fig.6.17.

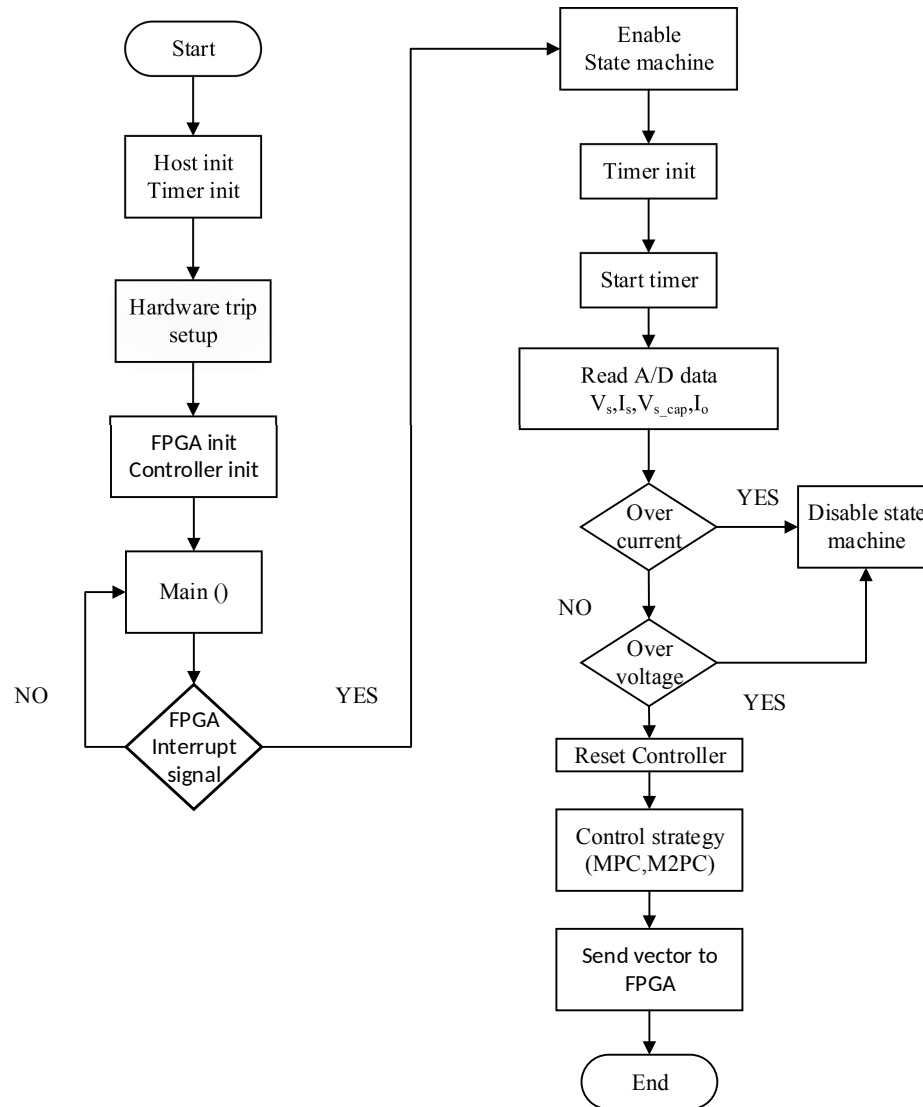


Figure 6.17: Flowchart illustrating the sequence of execution of the control software

Photographs of the complete experimental setup for a direct matrix converter feeding a RL load and an induction machine are shown in Figure 6.18 and Figure 6.19 respectively. The system parameters for all the tests conducted for RL load and the induction machine load are given in Table 6.2 and Table 6.3 respectively.

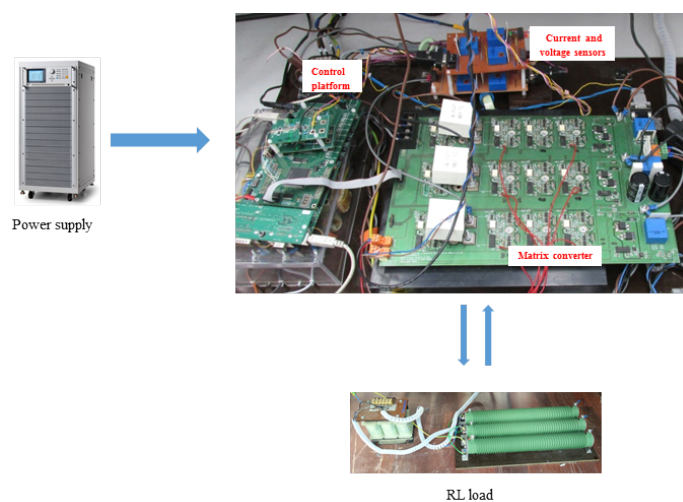


Figure 6.18: Experimental setup of MC with RL load

Table 6.2: Parameters of the experimental setup with RL load

Parameter	Value	Unit
<i>Load resistance, R</i>	10	Ω
<i>Load inductance, L</i>	3.75	mH
<i>Input filter capacitance, $C_f(\text{delta})$</i>	8.3	μF
<i>Input filter inductance, L_f</i>	0.7	mH
<i>Damping resistor, R_d</i>	10	Ω
<i>Supply frequency</i>	50	Hz
<i>Output frequency</i>	30	Hz
<i>Sampling time, T_s</i>	80	μs
<i>Supply Voltage, V_s</i>	70	V

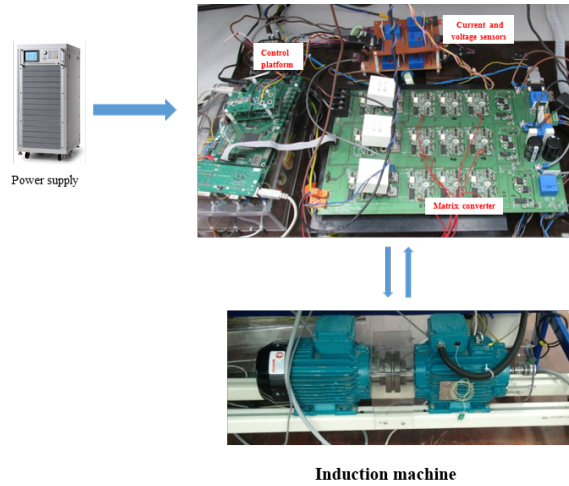


Figure 6.19: Experimental setup of MC with an induction machine

Table 6.3: Parameters of the test with induction machine

Parameter	Value	Unit
<i>Rated Power</i>	3	<i>kW</i>
<i>Rated RMS voltage(star)</i>	380	<i>V</i>
<i>Rated current</i>	10.3	<i>A</i>
<i>Rated speed</i>	1445	<i>rpm</i>
<i>Stator leakage inductance, Ll_s</i>	5.1*	<i>mH</i>
<i>Rotor leakage inductance, Ll_r</i>	7.3916	<i>mH</i>
<i>Mutual inductance, L_0</i>	0.2388	<i>H</i>
<i>Stator resistance, R_s</i>	1.9	Ω
<i>Rotor resistance, R_r</i>	1.33	Ω
<i>Supply Voltage, V_s</i>	120	<i>V</i>

* The leakage inductances are calculated by considering the ratio of the rotor and stator resistances, $R_r/R_s = L_s/L_r$.

6.3 Experimental Results

Control of a direct matrix converter feeding both RL load and an induction motor are implemented in the lab using the experimental setup explained in the previous section. Four different control strategies for load current control of a direct matrix converter system; namely PI controller, Model Predictive Control, Modulated Model Predictive Control and Direct Predictive Current-error Vector Control are considered to investigate the performance of the controllers. The converter and input filter parameters remain the same for all the control strategies and both types of load.

6.3.1 Control of MC with PI controllers

The load current of the MC with both RL load and an induction machine is controlled using PI controllers. The standard SVM modulation scheme is used to generate the voltage vectors.

6.3.1.1 Matrix converter feeding a RL load

An RL load is used to validate the simulation results for load current control using PI controllers in dq reference frame synchronous with the supply frequency. The dq current controllers are designed from the plant model as explained in section 3.2.1. A block diagram of the control scheme is shown in Figure 6.20.

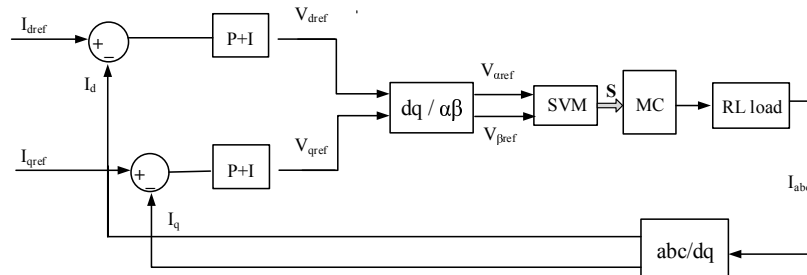


Figure 6.20: Control block diagram of matrix converter with RL load

A load current of 5A at 30 Hz is demanded from the system. The PI controller gains used in this test are $k_p = 15$ and $k_i = 1000$. The controlled load currents, MC output line voltage and FFT of Phase A load current are shown in Figure 6.21. From Figure 6.21, it is evident that the three phase load currents are sinusoidal without any steady state error. The THD is of the controlled load current is approximately 3.85%. This proves that the PI controller can provide a good steady state response.

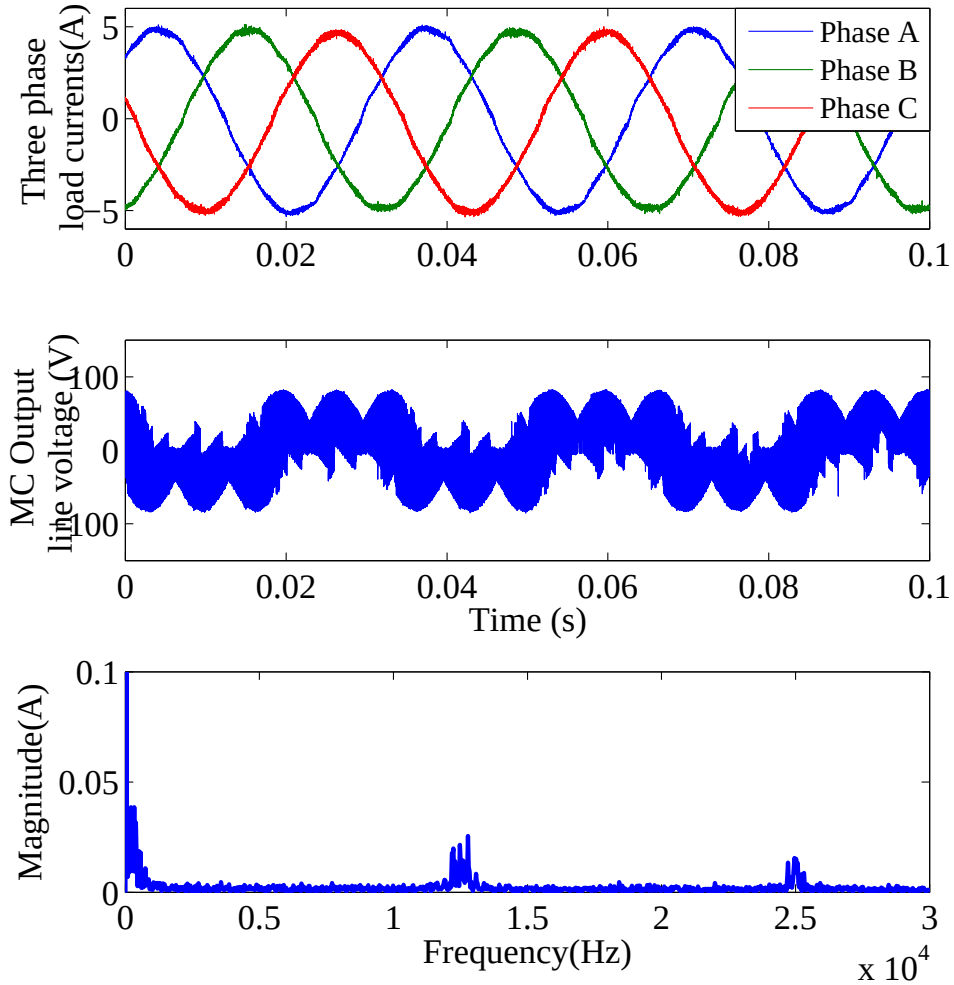


Figure 6.21: Experimental result for load current control using PI controllers

To demonstrate the dynamic response of the PI control approach, a step change of 2A to 4A is introduced to the magnitude of reference current. A step change of 20Hz

to 40Hz in frequency of the reference current is also introduced. Figure 6.22 shows the resulting waveforms.

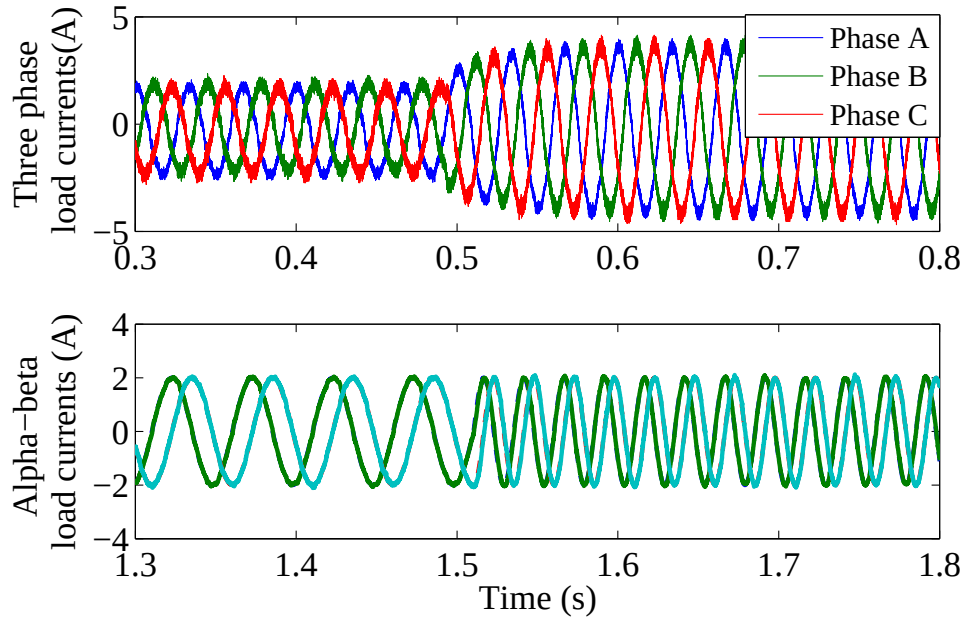


Figure 6.22: Dynamic performance of PI control during step change in amplitude(top) and frequency(bottom) of the reference current waveform

From Figure 6.22 it can be noted that the controller takes a couple of cycles to reach the desired amplitude of reference waveform after the step demand is applied. The transient response of the controller can be tuned to be faster but that may affect the steady state response adversely.

6.3.1.2 Matrix converter feeding an induction machine

A 3kW induction machine is used as a load to the direct MC. The stator currents and speed of the machine are controlled by PI controllers using a cascaded structure shown in Figure 3.12. The design of PI controller gains for both current and speed controllers for the control of IM is elaborated in Appendix C. The machine is magnetized prior to applying a speed demand. A reference speed of 500rpm is demanded at $t=0$ s at no load condition and the resulting waveforms for speed and stator currents are shown

in Figure 6.23.

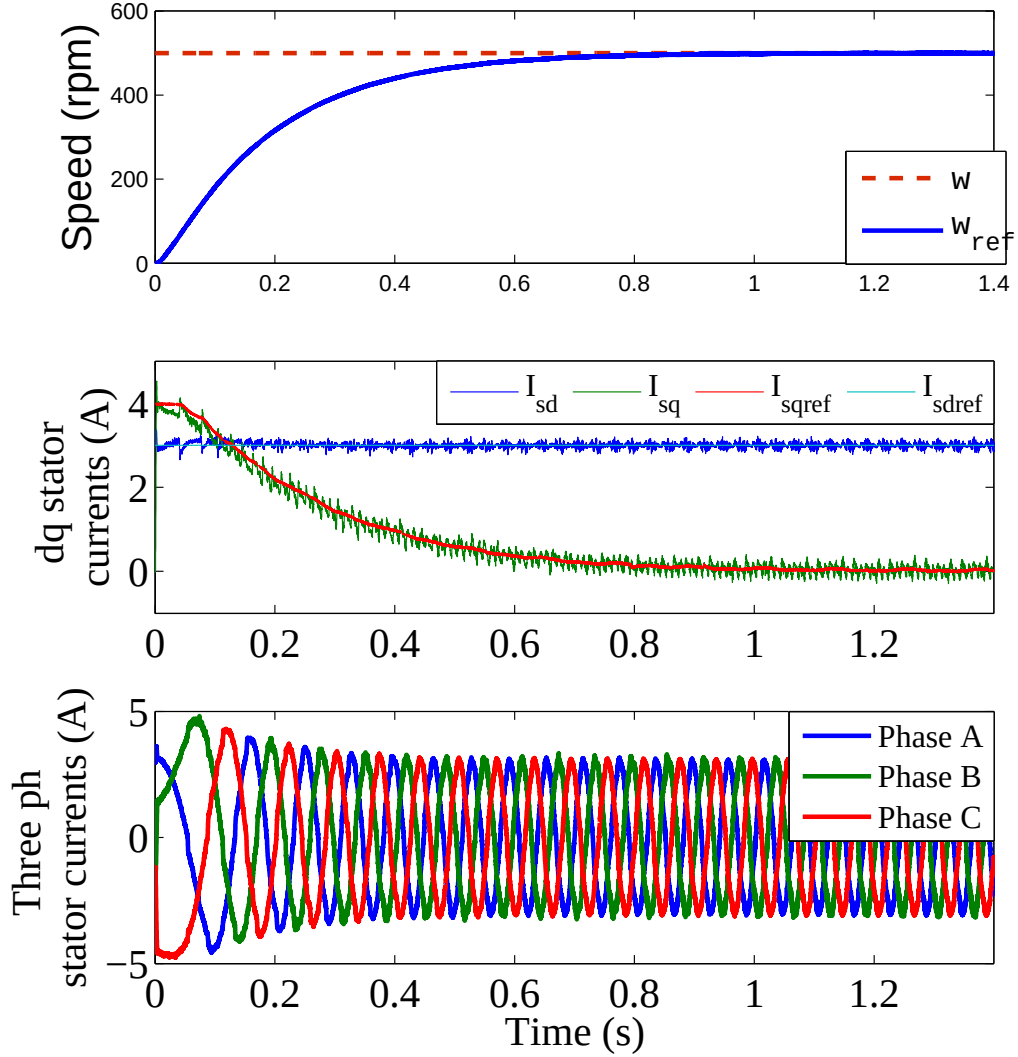


Figure 6.23: Speed, dq stator currents and 3phase stator currents during transient response when controlled by PI controllers

The steady state stator currents at no load and FFT of Phase A stator current are shown in Figure 6.24. The FFT reveals harmonics around switching frequency (12.5kHz) and its multiples as expected and the THD of the current waveform is about 3.3%. This verifies the good steady state performance seen in simulation results in section 3.3.6.

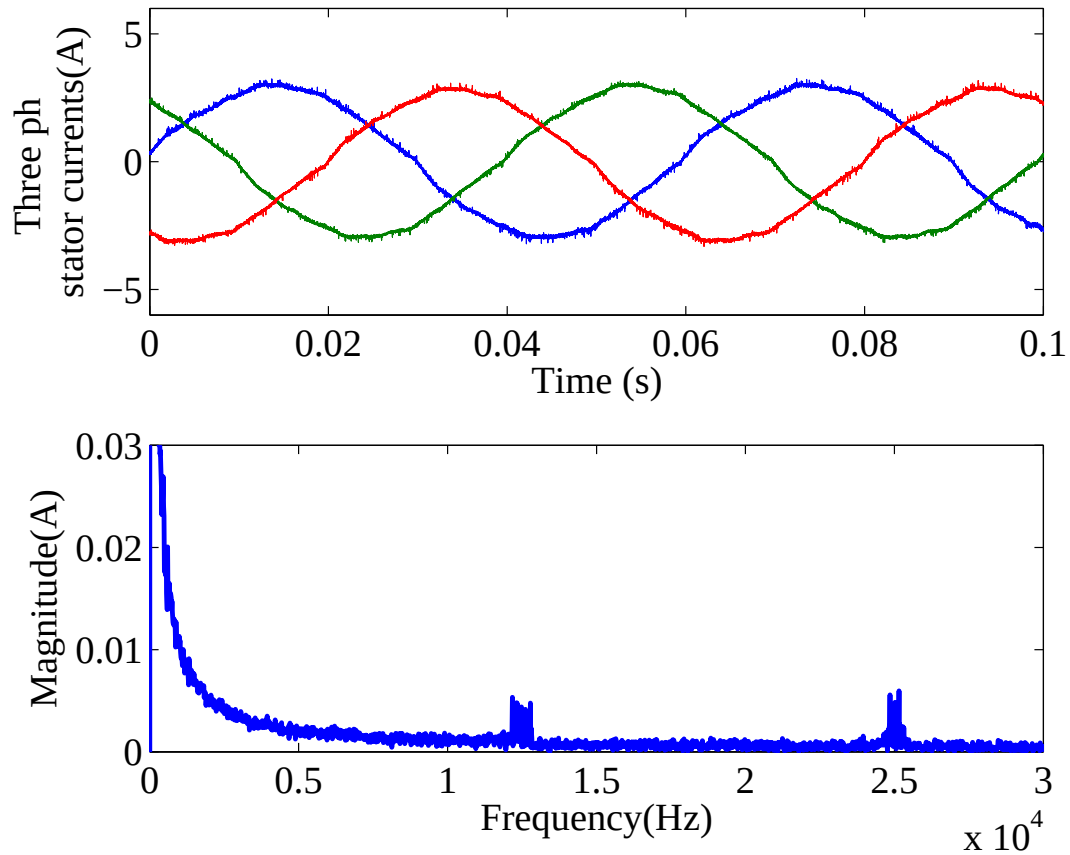


Figure 6.24: Three phase stator currents and FFT of Phase A stator current at steady state

To demonstrate the transient response of the controller a step in the speed demand from 200rpm to 500rpm is applied to the matrix converter. The resulting waveforms are shown in Figure 6.25. Figure 6.25 shows that the currents track the references without any steady state error and has a good dynamic response. The speed controller has slower dynamics than the current controllers as seen from Figure 6.23 and Figure 6.25.

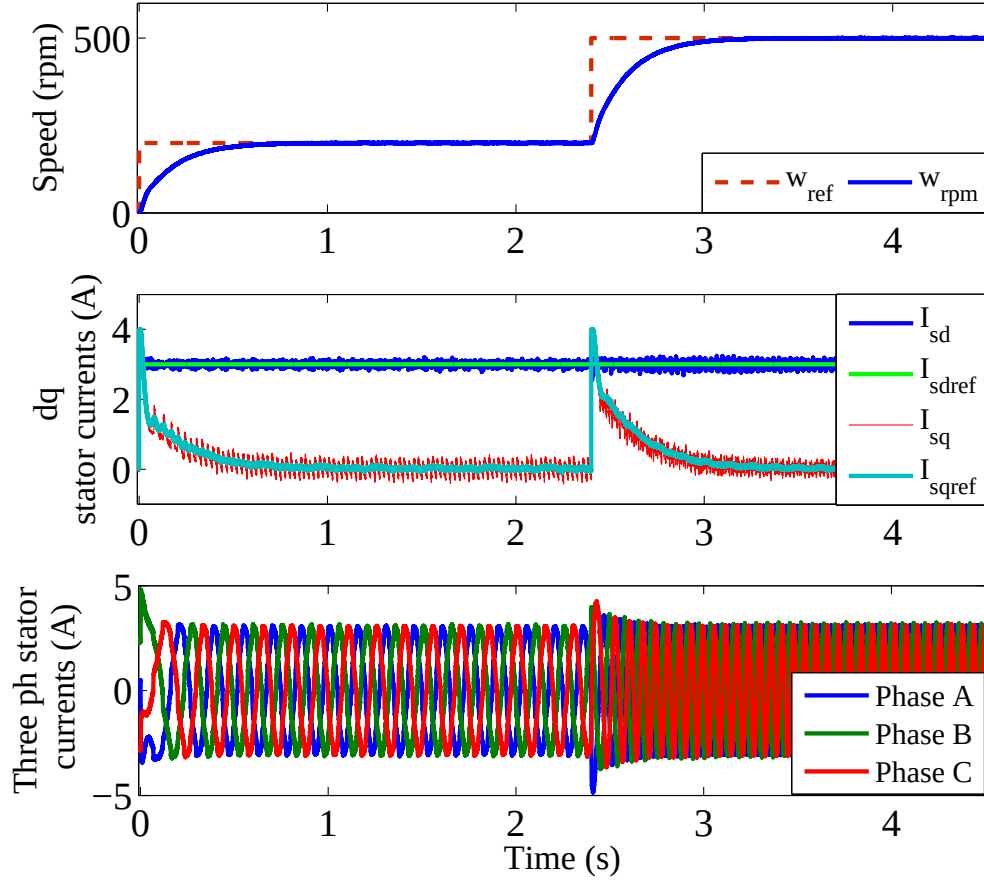


Figure 6.25: Transient response of PI controllers with SVM during a step change in speed demand of an IM

6.3.2 Control of MC with MPC

MPC is applied to the MC experimental setup to control the load currents with RL load and induction machine. The resulting waveforms for each case are shown below.

6.3.2.1 Matrix converter feeding a RL load

Experimental validation of load current control of a direct MC feeding a RL load using MPC will be discussed in this section. A reference current of 5A at 30 Hz

is demanded from the system and the resulting waveforms at steady state and its harmonic spectrum are shown in Figure 6.26. As seen from Figure 6.26, the harmonics spectrum contains frequencies in the lower range and are wide spread as expected from a predictive controller. The THD of Phase A load current is approximately 14.09 %. This shows that the steady state performance of MPC in terms of waveform quality is not desirable.

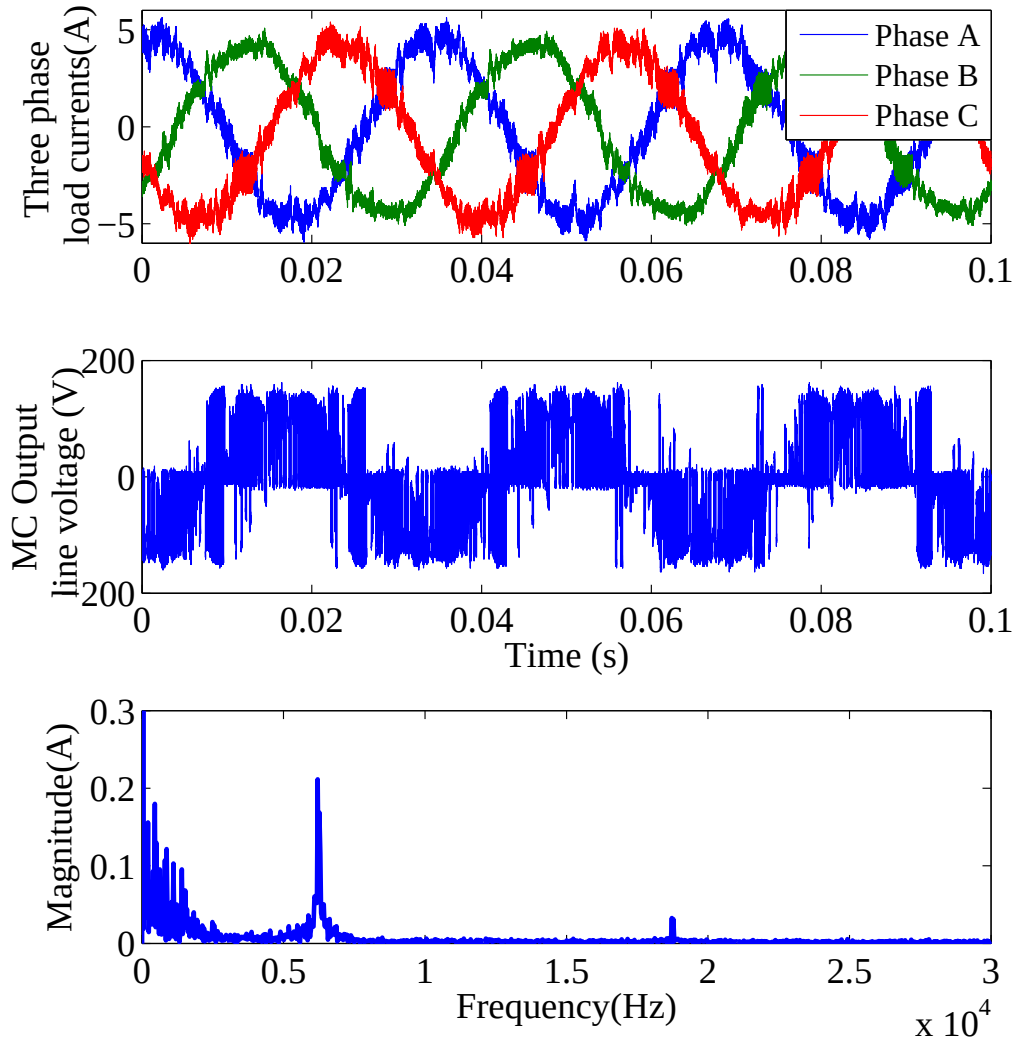


Figure 6.26: Load current of matrix converter controlled using MPC

To demonstrate the fast dynamic performance of MPC, a step change in the magni-

tude and frequency of the reference waveform is introduced. The resulting waveforms are shown in Figure 6.27. From Figure 6.27, it is evident that the load currents reacts to the step change in amplitude and frequency instantaneously. The test results conforms with the simulation results presented in section 4.6.1.

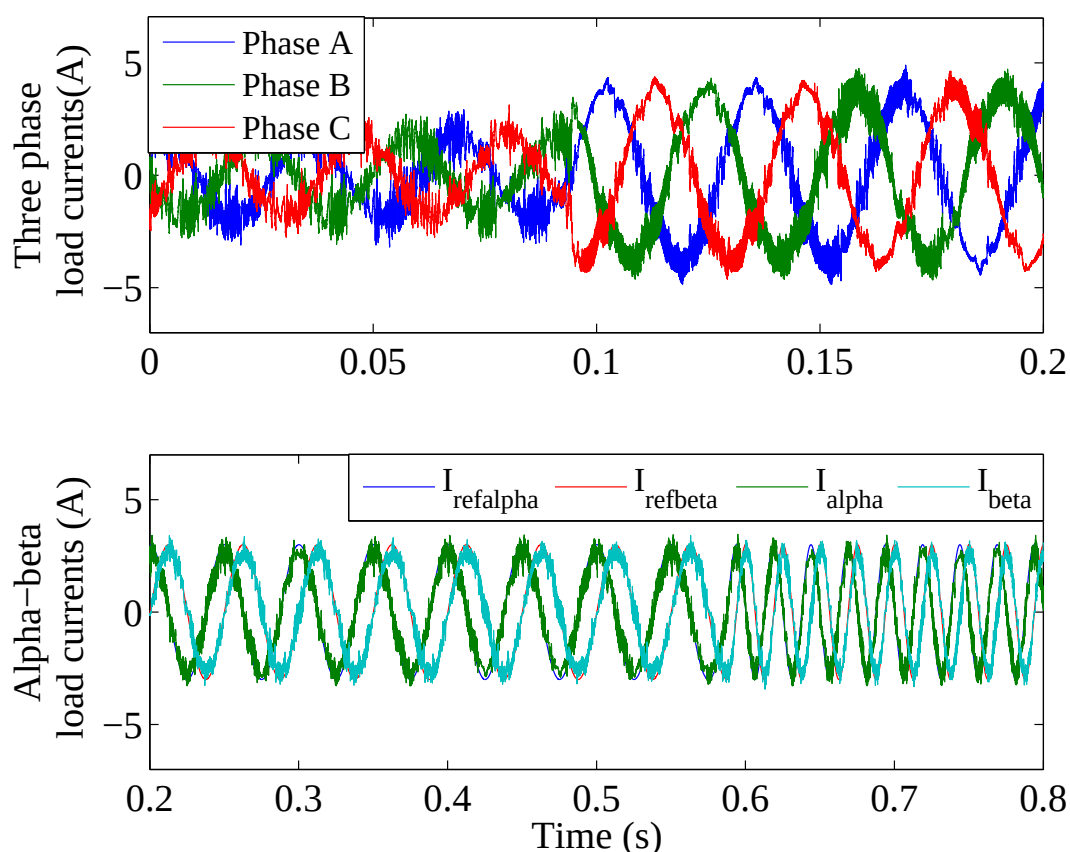


Figure 6.27: Dynamic performance of MPC during step change in the magnitude of the reference current

6.3.2.2 Matrix converter feeding an induction machine

A 3kW induction machine is used as a load for a direct MC to implement stator current control using MPC. The speed of the machine is controlled using a PI controller and the torque producing current reference is fed into the predictive current controller. The machine is magnetized initially and the current control is tested at no

load condition. A speed reference of 500rpm is demanded from the system and the waveforms of speed and stator currents showing the machine start up from 0-500rpm and following the references is shown in Figure 6.28.

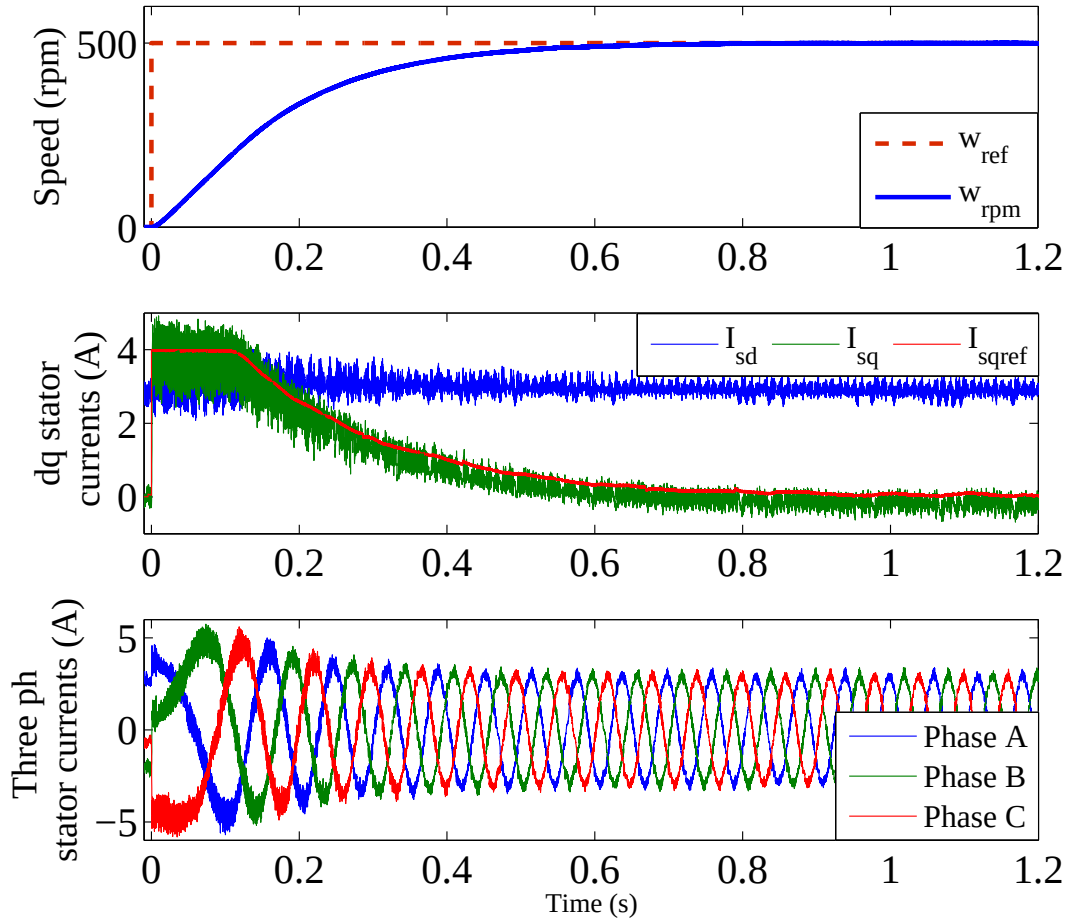


Figure 6.28: Reference tracking of speed and dq-stator currents and 3-ph stator currents during machine start-up when controlled using MPC at no load

The three phase stator currents at steady state and the harmonic spectrum of Phase A stator current is shown in Figure 6.29. It is evident from Figure 6.29 that the harmonics present are mostly in the lower harmonic range and there are no frequencies in the range of switching frequency (12.5kHz). This indicates a variable switching frequency operation. The THD of the controlled stator current waveform is approximately 5.8%.

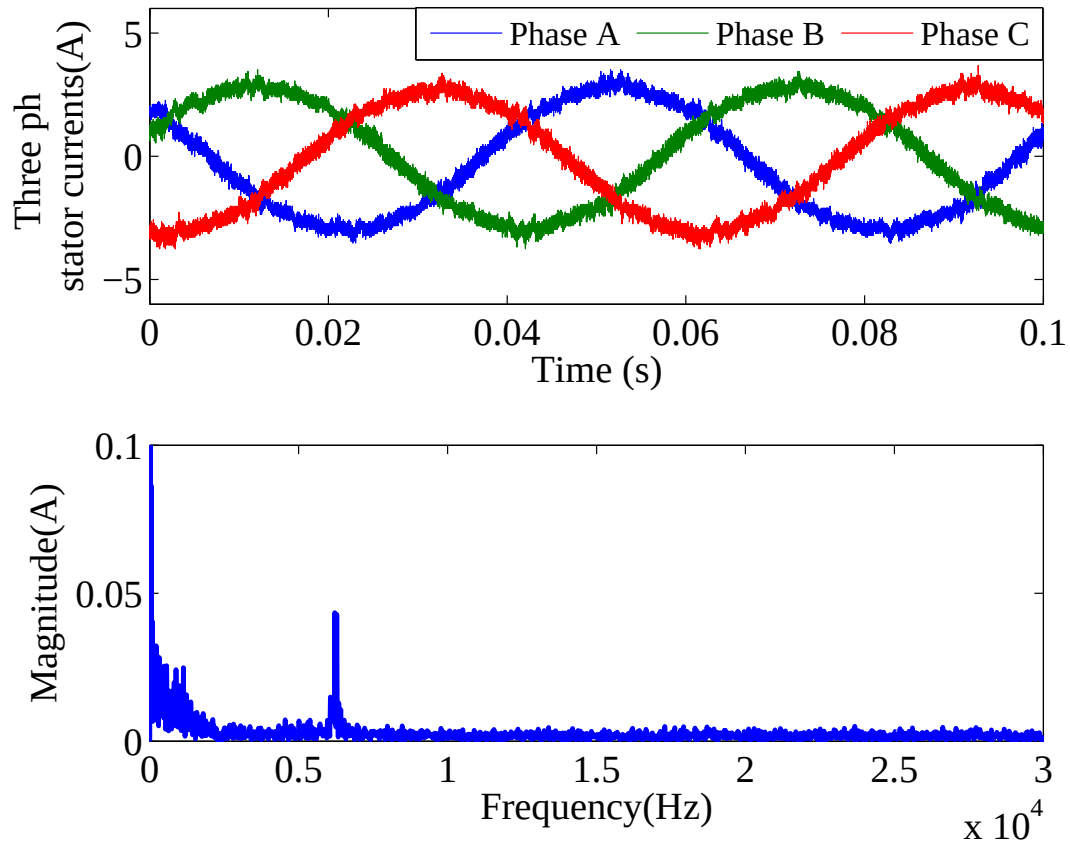


Figure 6.29: Three phase steady state stator currents(top) and harmonic spectrum for Phase A stator current(bottom) controlled by MPC at no load

To demonstrate the dynamic performance of the controller, the speed demand to the machine is changed from 200rpm to 500rpm at $t=2.4$ s and the reference tracking of speed and dq-stator currents during the same are shown in Figure 6.30.

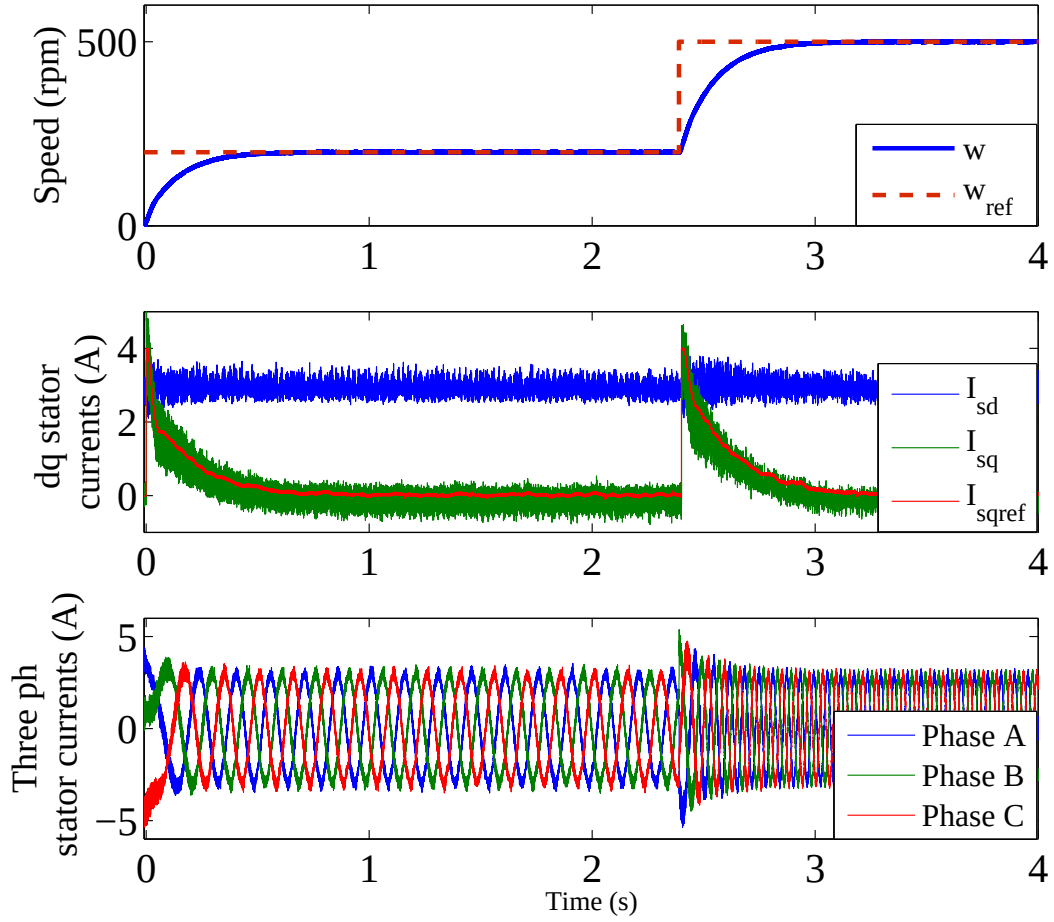


Figure 6.30: Dynamic response of MPC during a step change in speed from 200rpm-500rpm at no load

6.3.3 Control of MC with M2PC

Experimental validation of M2PC strategy for matrix converter current control is obtained by testing the control algorithm for two different types of load- RL load and an induction machine. The experimental results for these tests are shown in the next section.

6.3.3.1 Matrix converter feeding a RL load

The load currents of the direct matrix converter are controlled using M2PC strategy. A reference current of 5A at 30Hz is demanded from the system and the experimental results of controlled load current, MC output line voltage and the harmonic spectrum of the controlled load current are shown in Figure 6.31.

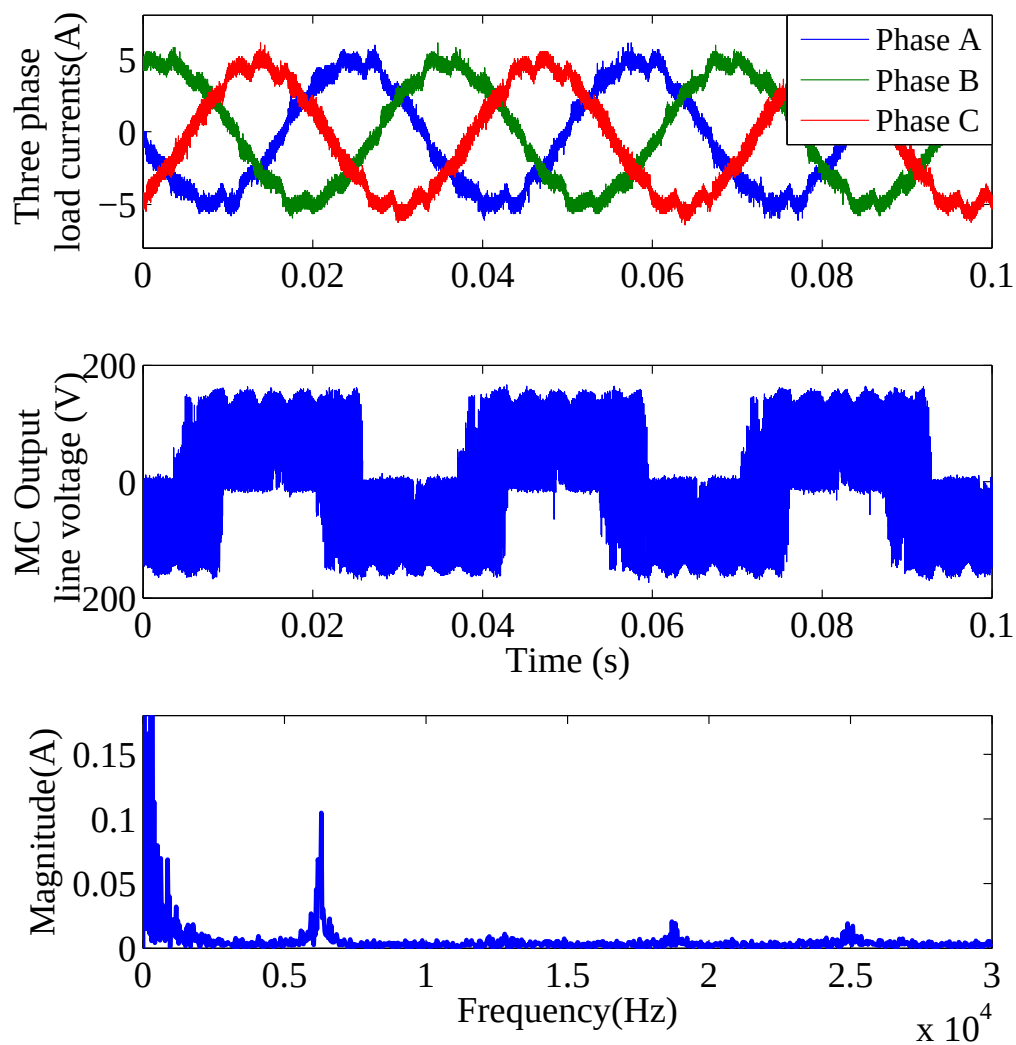


Figure 6.31: Phase A load current, MC output line voltage and harmonic spectrum of load current at steady state

Figure 6.31 indicates that the load currents are controlled without any steady state error. The FFT of Phase A load current shows the presence of harmonics around 12.5 kHz and its multiples which validates the fixed switching frequency operation. The THD of the waveform is about 10.8%. It can be noted that the current quality is improved compared to the results from MPC in section 6.3.2.1.

To analyse the transient behaviour of the control strategy, a step demand in the amplitude and frequency of the reference current waveform is applied. A step in the amplitude of reference current waveform from 2A to 4A is applied and the frequency of the reference load current is changed from 20Hz to 40Hz and the resulting waveforms are shown in Figure 6.32. It is evident from these results that M2PC is capable of fast transient response.

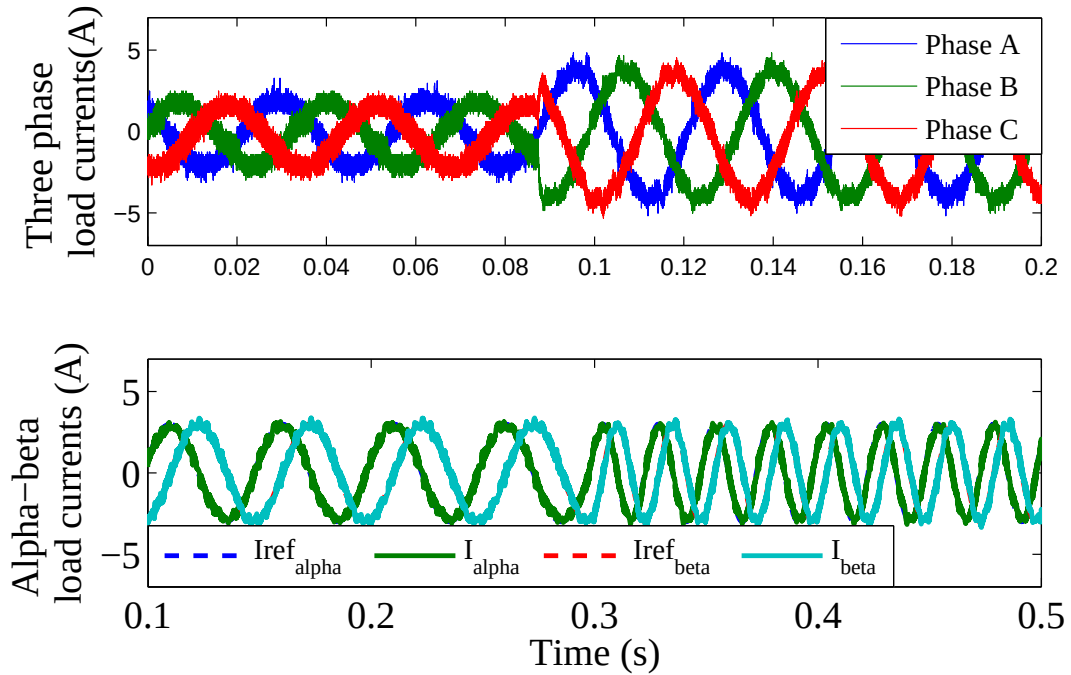


Figure 6.32: Dynamic performance of M2PC during step change in the magnitude(top) and frequency(bottom) of the reference load current

6.3.3.2 Matrix converter feeding an induction machine

The stator currents of a 3kW induction machine are controlled using the M2PC strategy to validate the simulation results shown in chapter 5. The speed of the machine is controlled by a PI controller and the dq-axis stator currents are controlled using the M2PC strategy. The machine is magnetized before demanding a speed reference of 500rpm. The experimental results showing reference tracking of speed and dq-axis stator currents and three phase stator currents during the machine start up from 0-500rpm are shown in Figure 6.33.

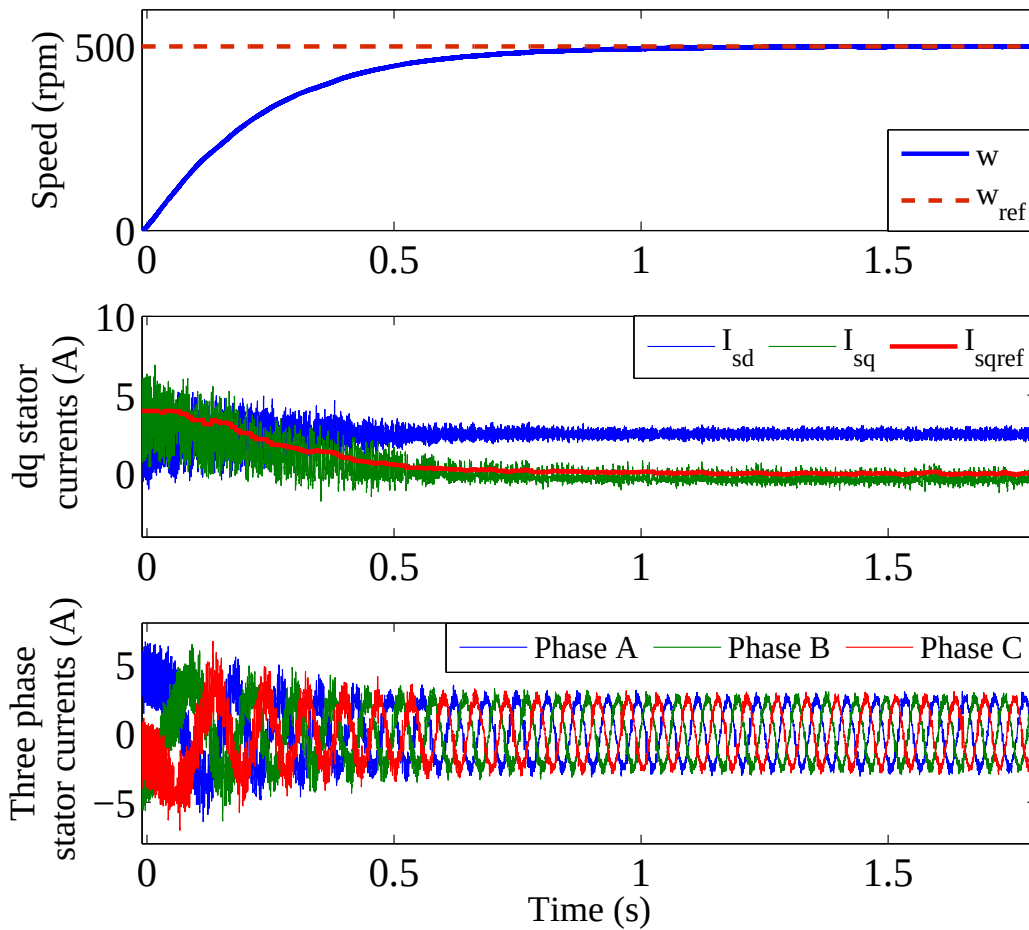


Figure 6.33: Transient response of speed, dq-axis and three phase stator currents for matrix converter feeding an IM controlled by M2PC at no load

The controlled three phase stator currents at no load condition and the harmonic spectrum of Phase A stator current are shown in Figure 6.34. From Figure 6.34, it can be seen that the steady state stator currents are not improved compared to that of MPC and the THD of the currents is approximately 12.1%. The results are also not similar to the simulation results for the same system parameters. The exact reason for this anomaly has not been identified; however one of the reasons could be the approximate duty cycle calculation approach employed in this strategy. Another reason could be parameter variation at higher speeds. The frequency spectrum of the stator current however shows harmonics in the range of switching frequency and its multiples confirming fixed switching frequency operation.

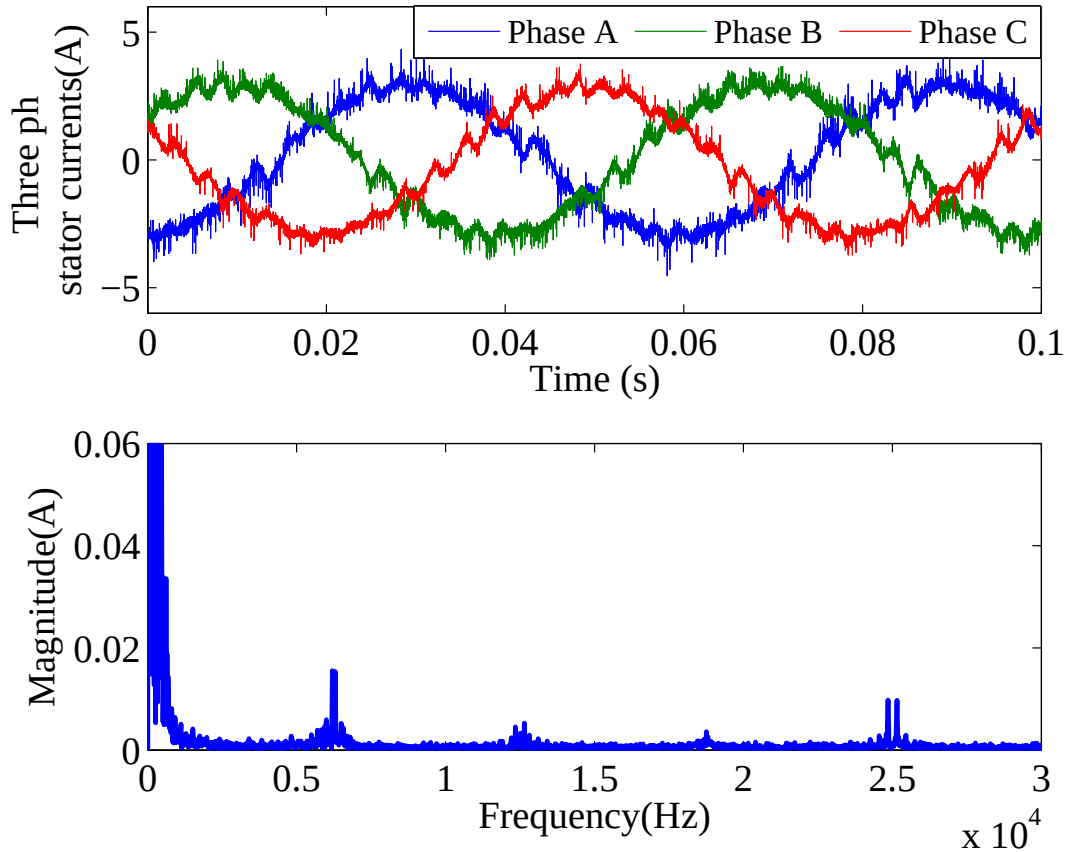


Figure 6.34: Three ph stator currents and FFT of Phase A stator current at steady state under no load condition

The dynamic response of the controller during a step demand in speed from 200rpm

to 500rpm is shown in Figure 6.35. From Figure 6.35 it is evident that the currents and speed follow the reference without any steady state error.

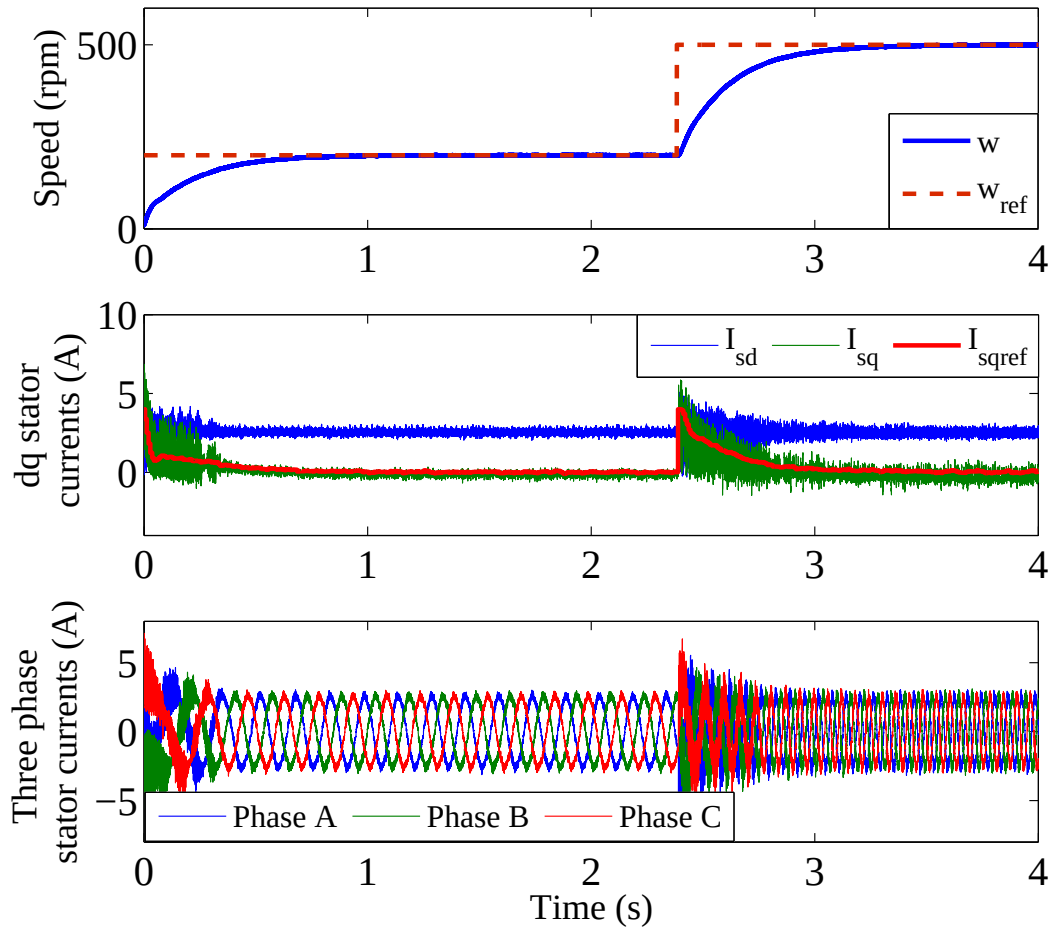


Figure 6.35: Reference tracking of speed and dq-stator currents and three phase stator currents during a step change from 200rpm to 500rpm when controlled by M2PC

6.3.3.3 M2PC speed and current controller

In order to study the multi-objective control capability of M2PC strategy, both speed and stator currents of the machine are controlled by M2PC. This eliminates any cascaded control compared to the previous section. The speed and current controllers can be tuned by changing the weighting factors in the cost function. A speed reference

of 500rpm is demanded after magnetizing the machine. The resulting waveforms showing the speed and dq-axis stator currents tracking the reference signals are shown in Figure 6.36. The speed and current controller is being controlled by minimizing the same cost function. The weighting factor for speed control is chosen to be 0.6 based on trial and error.

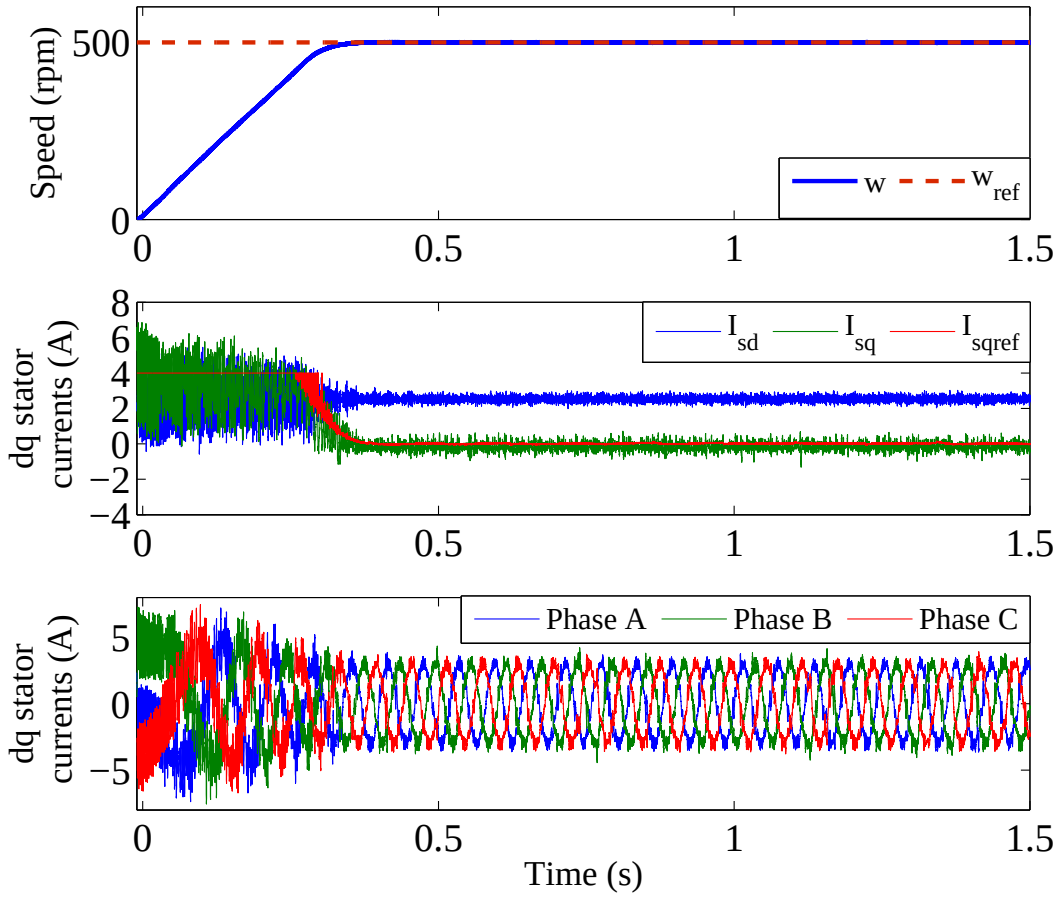


Figure 6.36: Reference tracking of speed and dq-axis stator current during the machine start up from 0-500rpm at no load

The three phase stator currents at steady state and the FFT of the controlled stator current is shown in Figure 6.37. The controlled Phase A stator current has a THD of 12.1%. The machine reaches the demanded speed in approximately 0.3s which is much lower than that achieved by PI controller (more than 0.5s) in the previous cases. This is due to the predictive based control of speed. If faster response is

required the weighting factor of the corresponding variable can be increased.

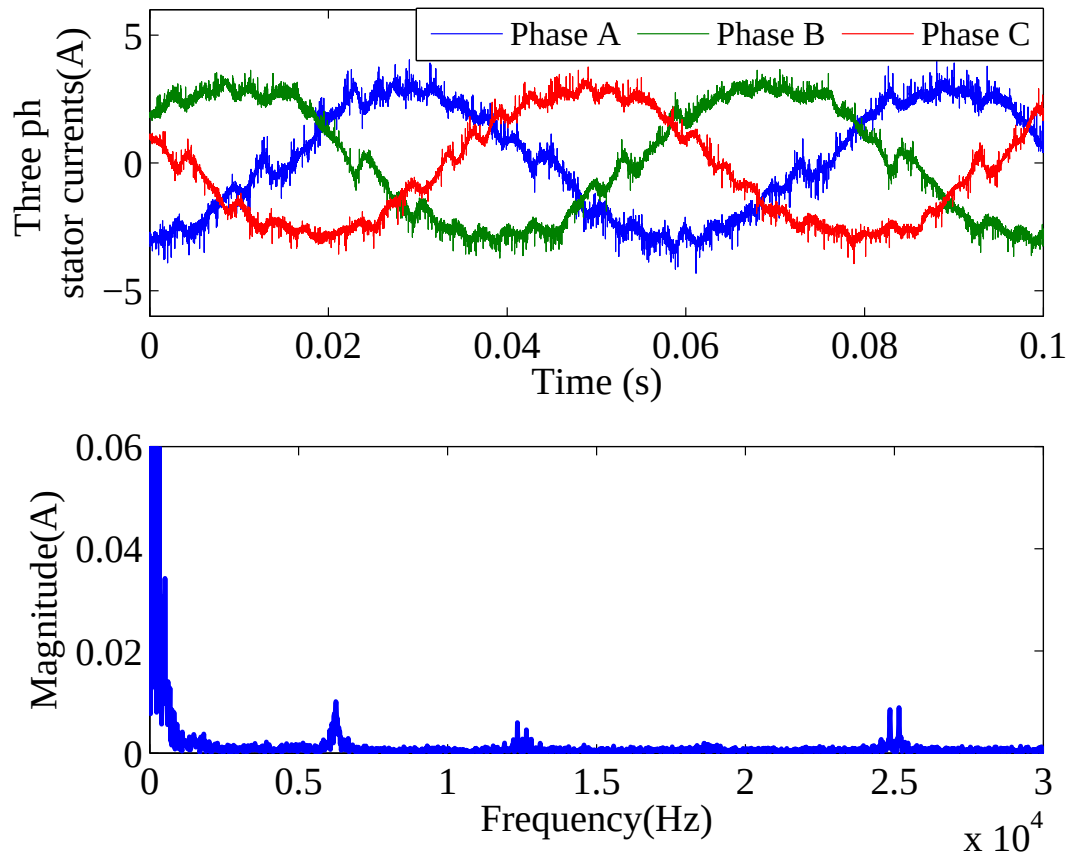


Figure 6.37: Three phase stator currents at steady state and the FFT of Phase A stator current for multi-objective M2PC with speed and current control at no load

Even though the current quality is not similar to the simulation test results in section 5.5.2.2, the multi-objective control capability of M2PC has been proved. This can also be extended to control the input reactive power together with load current and speed control.

6.3.4 Control of MC with DPCVC

Another strategy that results in a fixed switching frequency operation called DPCVC is proposed for a direct matrix converter in section 5.7. Simulation tests were carried out to study the performance of this method on a direct matrix converter feeding an RL load and an induction machine at no load condition. To validate the simulation results, this method is applied to the matrix converter feeding a RL load and an induction machine and the performance of the control method for different loading conditions are investigated.

6.3.4.1 Matrix converter feeding a RL load

DPCVC is tested for load current control of a direct matrix converter feeding an RL load. A load current of 5A at 30Hz was demanded from the system and the resulting waveforms of three phase load currents, MC output line voltage and the harmonic spectrum of Phase A load current are shown in Figure 6.38. The THD of the controlled waveform is about 3.97%. The current quality in this case is very good compared to MPC and M2PC for the same loading conditions.

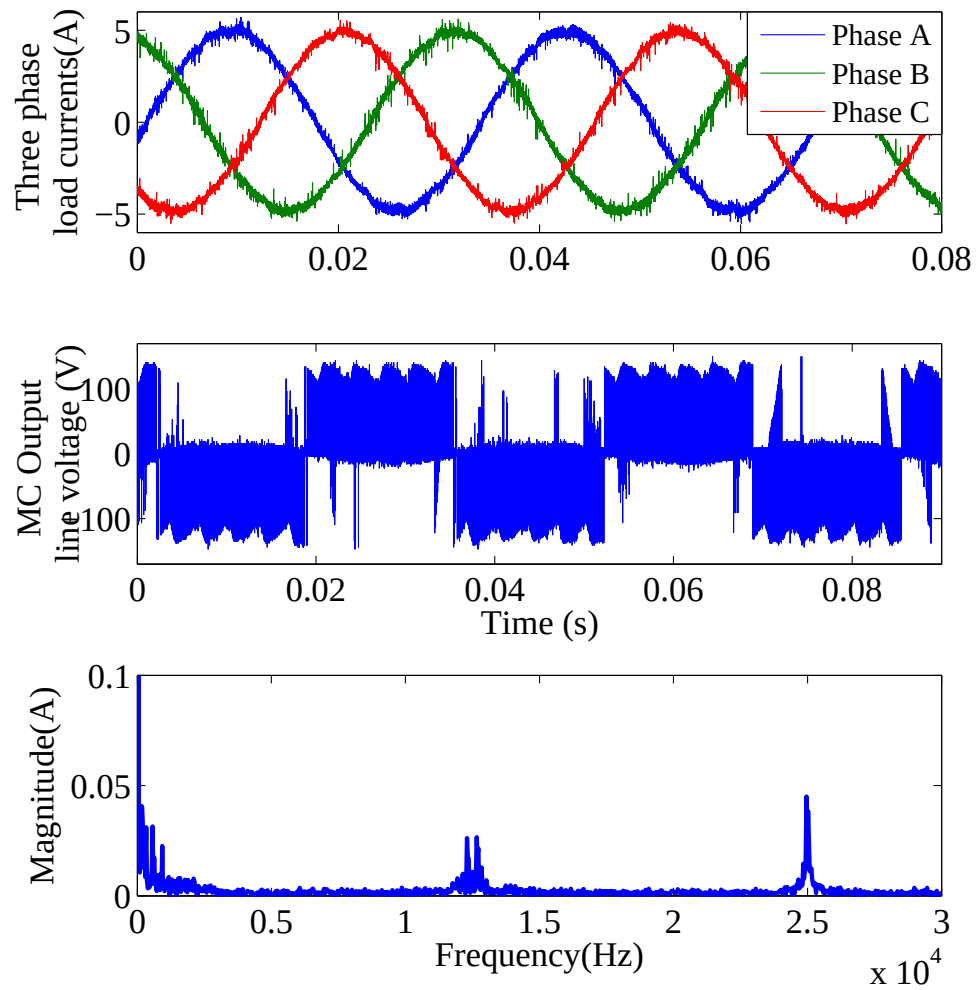


Figure 6.38: Three phase load currents (top), MC output line voltage (V)(middle) and FFT of Phase A load current (bottom) at steady state

From Figure 6.38, it is evident that the steady state performance of this method is superior to that of M2PC and MPC. The harmonic spectrum indicated harmonics at switching frequency (12.5kHz) and its multiples. In order to test the transient behaviour of this strategy, a step demand in the magnitude from 2-5A and frequency of the reference current from 20-40Hz are demanded separately. The resulting load current waveforms are shown in Figure 6.39.

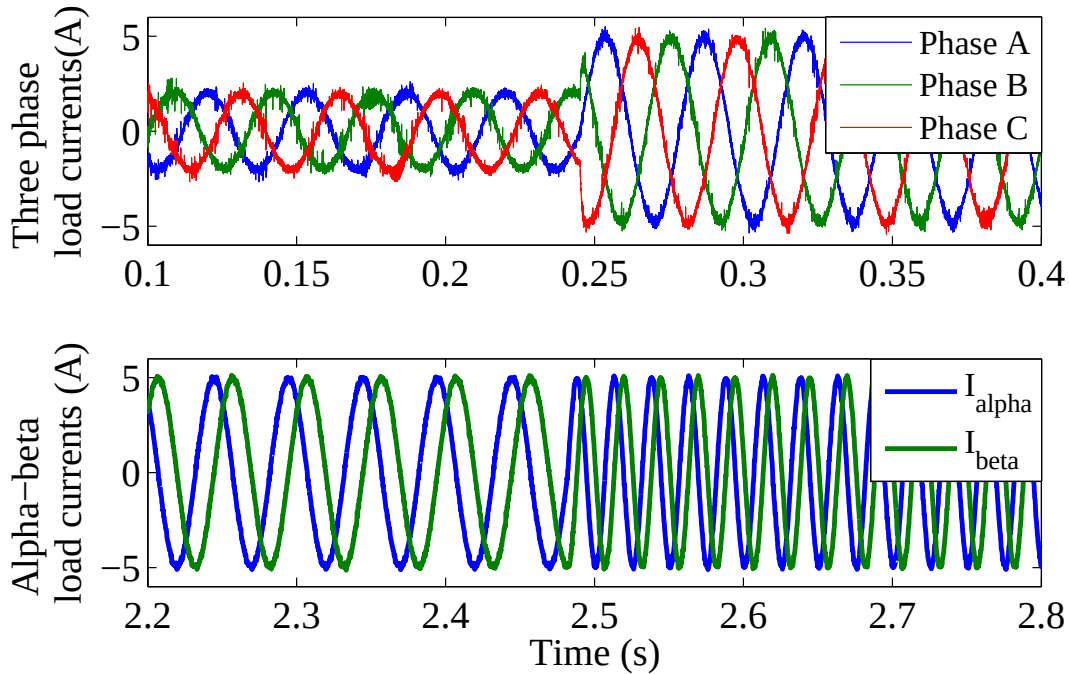


Figure 6.39: Transient response of predictive control based modulation strategy for step changes in magnitude (top) and frequency (bottom) of the reference current

Figure 6.39 indicates the fast transient response achieved by this method. The results are similar to the simulation results which validates the control theory. Thus with an RL load, it is proved that this method has improved current quality compared to the M2PC and MPC while maintaining a fast dynamic response.

6.3.4.2 Matrix converter feeding an induction machine

The motivation behind the control strategy mentioned in section 5.7 is to develop a control method with fast transient response and good waveform quality that comes with the presence of a modulation technique. DPCVC strategy is applied to a matrix converter feeding an induction motor under no load condition. The speed of the machine is controlled by a PI controller and the stator current control is achieved by the DPCVC. The machine is initially magnetized and a speed demand of 500 rpm is applied to the system. The resulting waveforms of the speed and d and q-axis stator current tracking their respective references is shown in Figure 6.40. It is evident from Figure 6.40 that the steady state response of the controller is improved compared to the M2PC and MPC as the stator currents have minimal ripple content. This validates the simulation results presented in section 5.7.1.

In order to analyse the transient behaviour of the controller, a step in the speed demand from 200rpm to 500rpm is applied and the resulting waveforms for speed of the machine and stator currents are shown in Figure 6.41. The figure indicates fast transient response comparable to the MPC strategy.

The three phase stator currents and MC output line voltage at a constant speed of 500rpm is presented in Figure 6.42. As seen from the figure, the quality of the stator current are improved compared to that produced by the M2PC. The harmonic spectrum of the stator current in Figure 6.42 shows frequencies in the range of switching frequency (12.5kHz) and its multiples which indicates a fixed switching frequency operation. The THD of the stator current waveforms is about 3.7%.

To summarise, the experimental results for the DPCVC strategy indicates that it delivers improved current quality waveforms with very fast transient behavior.

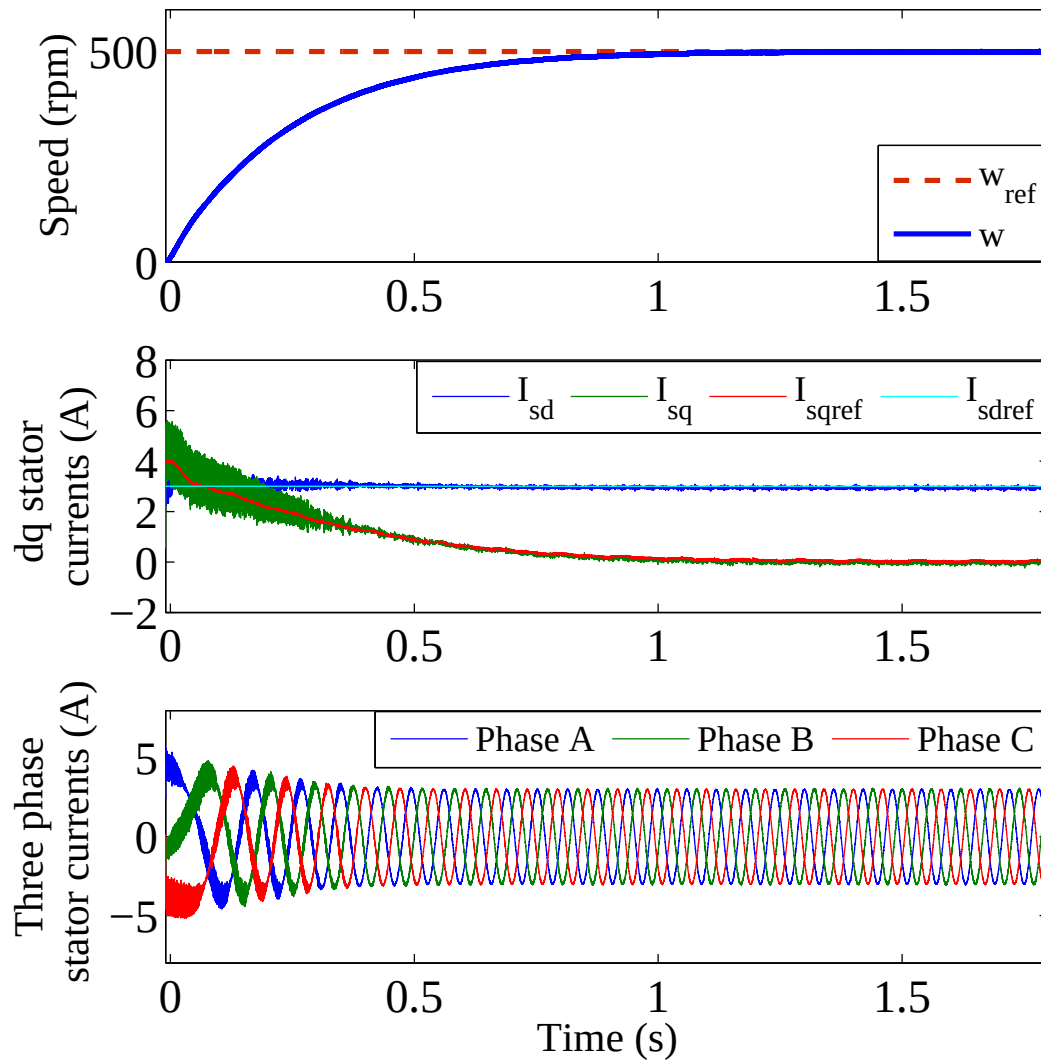


Figure 6.40: Speed and q-axis stator current reference tracking during machine start up from 0-500rpm at no load

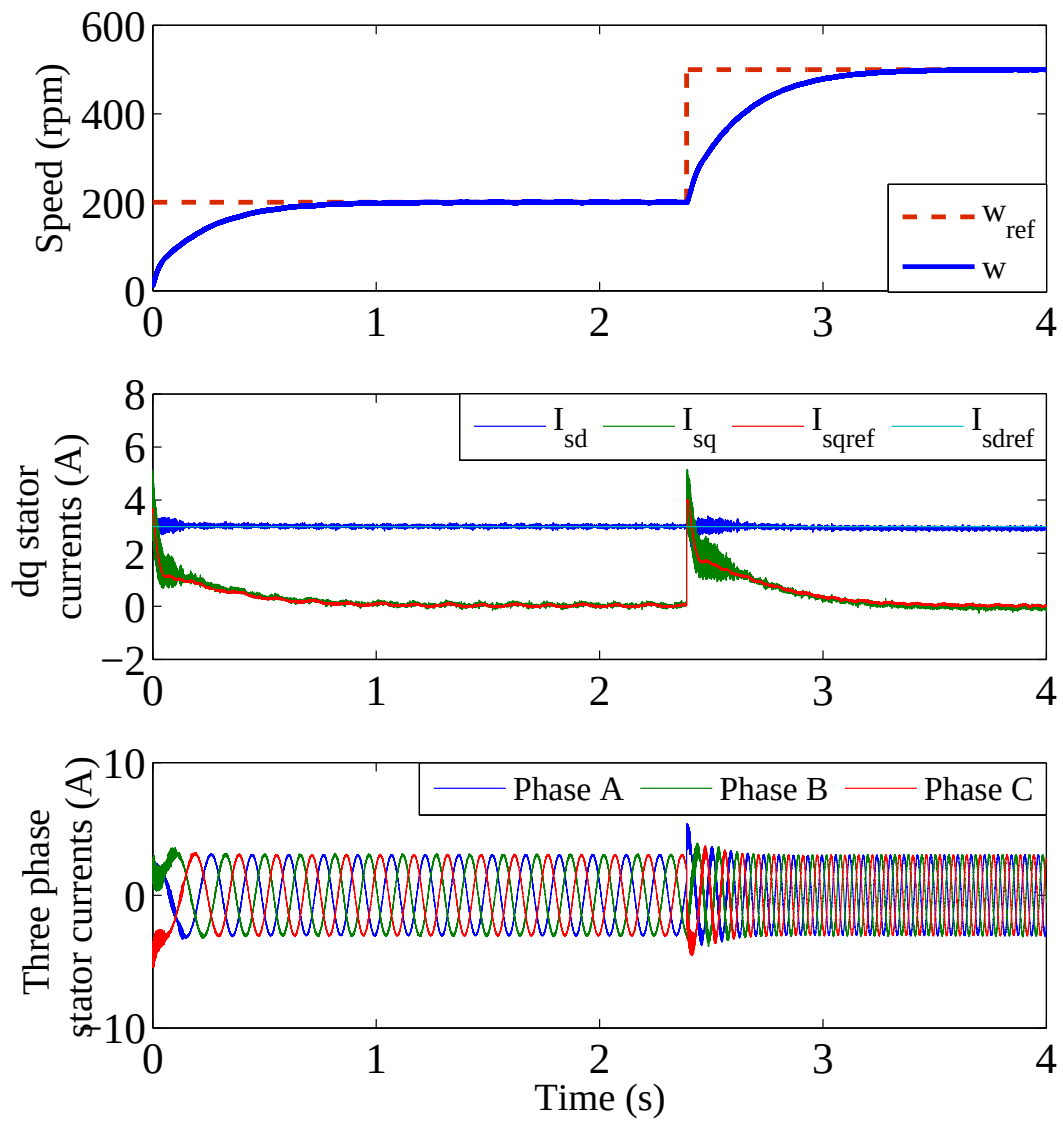


Figure 6.41: Speed, dq axis and three phase stator currents during step change in speed from 200-500rpm under no load condition

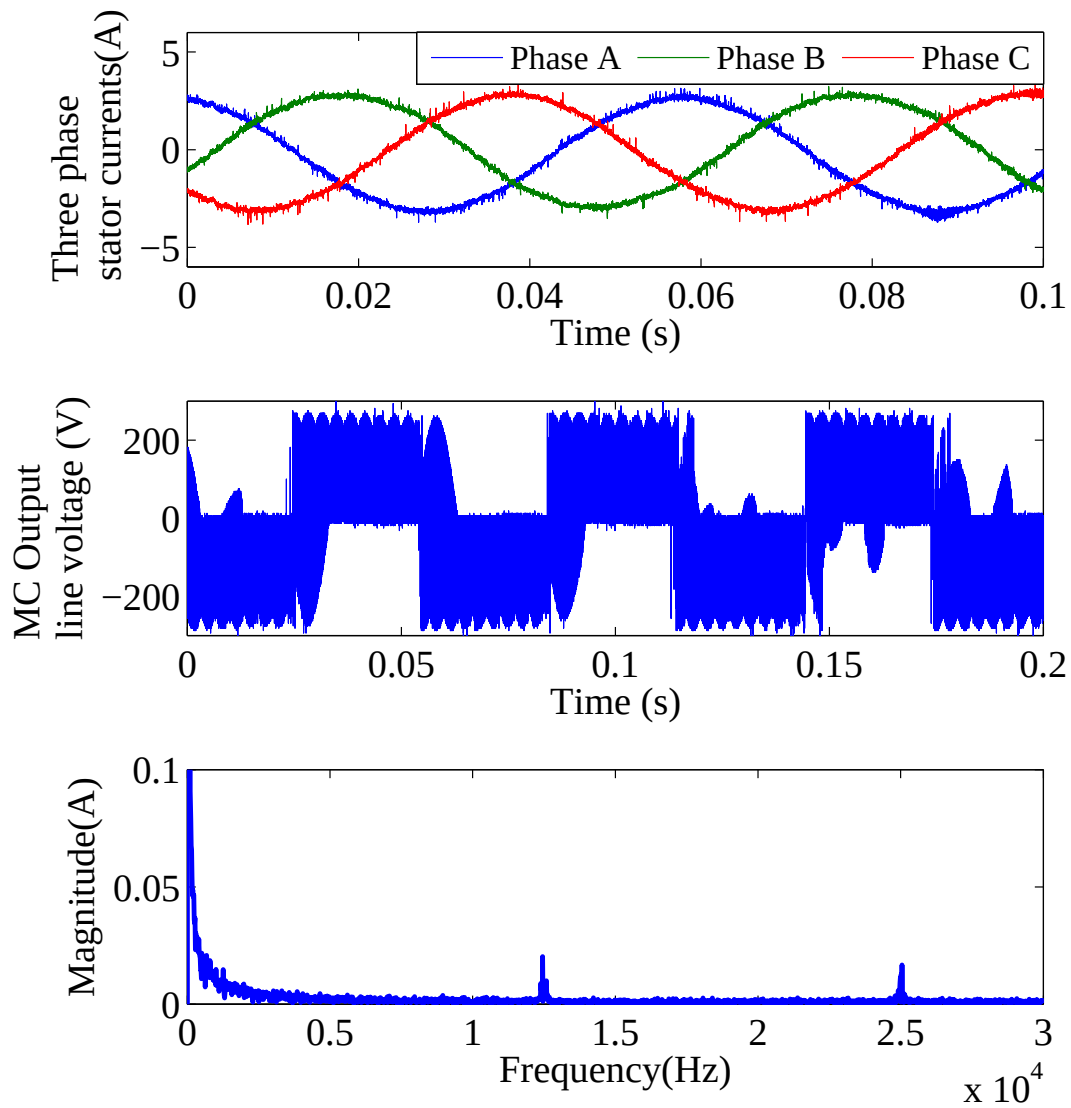


Figure 6.42: Three phase stator currents, MC output line voltage at steady state and FFT of Phase A load current at no load

6.4 Effect of Sampling Interval on Waveform Quality

To analyse the effect of sampling interval on the quality of the controlled currents, the MPC is applied to a direct matrix converter feeding and RL load using three different sampling intervals, for instance $50\mu s$, $80\mu s$ and $100\mu s$. The THDs of the controlled waveforms from the simulation and experimental tests are tabulated as shown in Table 6.4. It can be noted that for MPC, quality of the controlled currents are improved as the sampling interval gets smaller.

Table 6.4: Comparison of THDs of load currents controlled by MPC with different sampling intervals

Sampling time	Total Harmonic Distortion	
	<i>Simulation Results</i>	<i>Experimental Results</i>
$50\mu s$	6.7%	9.8%
$80\mu s$	8.09%	14.09%
$100\mu s$	10.48%	16.7%

Since the performance of MPC is largely dependant on the sampling interval, it is worth studying if this applies to M2PC. To analyse the effect of sampling time on the quality of the controlled waveforms, the M2PC strategy is implemented to control a direct matrix converter feeding and RL load for three different sampling times such as $50\mu s$, $80\mu s$ and $100\mu s$. The quality of the controlled currents are assessed by comparing the THDs for different sampling times. Both simulation and experimental tests are conducted for three different sampling intervals or switching frequencies and the THDs of the resulting waveforms are tabulated as shown in Table 6.5.

Table 6.5: Comparison of THDs of load currents controlled by M2PC with different sampling intervals

Sampling time	Total Harmonic Distortion	
	<i>Simulation Results</i>	<i>Experimental Results</i>
$50\mu s$	4.0%	7.33%
$80\mu s$	6.3%	10.8%
$100\mu s$	7.5%	12.07%

Table 6.5 indicates that there is a considerable effect on the quality of the controlled waveforms as the sampling interval is changed. The current quality was the best when a sampling time of $50\mu s$ is considered and the worst when the sampling time is $100\mu s$. It is interesting to note that the M2PC strategy is also dependant on the sampling interval considered which is expected as it is predominantly a predictive control based method.

6.5 Comparison of control strategies

To compare the performance of different control methods, a number of characteristics such as the steady state response, transient response and switching stresses can be considered. The stress on the switches depends on the switching frequency and the number of commutations required per sampling interval and SVM and DPCVC will have the minimum switching stress as the number of commutations are limited. In this case, the vectors applied within a sampling period changes only when there is a sector change at the input. However in M2PC, the switching sequence can change at every sample time resulting in a higher switching stress. Eventhough MPC results in a variable switching frequency operation, the it will result in less switching stress compared to M2PC. In this work, a quantitative analysis is made based on two aspects namely steady state and transient response. The THD of the currents controlled by the four methods discussed in this thesis can be considered as a measure of their

steady state performance. The simulation and experimental results for load current control of a direct MC with RL load are considered to compare the THDs of the currents controlled by PI Controller, MPC, M2PC and DPCVC. Table 6.6 shows the THD values of currents controlled by the different methods in simulation and experiment for the same system parameters.

Table 6.6: RL load: Comparison of THDs of load currents for different control strategies

<i>Control Strategy</i>	<i>Simulation result</i>	<i>Experimental result</i>
<i>PI Controller</i>	1.89%	3.85%
<i>MPC</i>	8.09%	14.09%
<i>M2PC</i>	6.3%	10.8%
<i>DPCVC</i>	2.3%	3.97%

From Table 6.6 it can be seen that in both simulation and experiment, PI Controller with SVM and DPCVC produced waveforms with least harmonic distortion followed by M2PC. MPC resulted in waveforms with the worst current quality as expected. Since DPCVC resulted in currents with very similar THDs to that of a PI controller with SVM, the control theory is validated. Even though the currents controlled by M2PC has less distortion compared to MPC, it is much higher than DPCVC. This suggests that out of the two methods proposed in this thesis for fixed switching frequency operation, DPCVC is superior in terms of quality of the controlled waveforms.

To analyse the transient response of the four controllers, stator current control of MC with an induction machine as load is considered. The simulation and experimental results during transients for the different controllers is utilised to conduct this analysis. Both the simulation and experimental tests are conducted for the same supply voltage and speed controller gains for all the control methods. The q-axis stator current of the induction machine during start up is considered to study the transient response. The step demand in the q-axis stator current reference corresponds to the step demand in speed of the machine. Figure 6.43 shows the simulation results for reference tracking

of q-axis stator current during start-up, for the four controllers. It is evident from Figure 6.43 that the q-axis current follows the reference fastest in the case of MPC, M2PC and DPCVC which are all predictive control based methods. Even though the PI controller is able to achieve fast response, it is not comparable to the predictive control based methods.

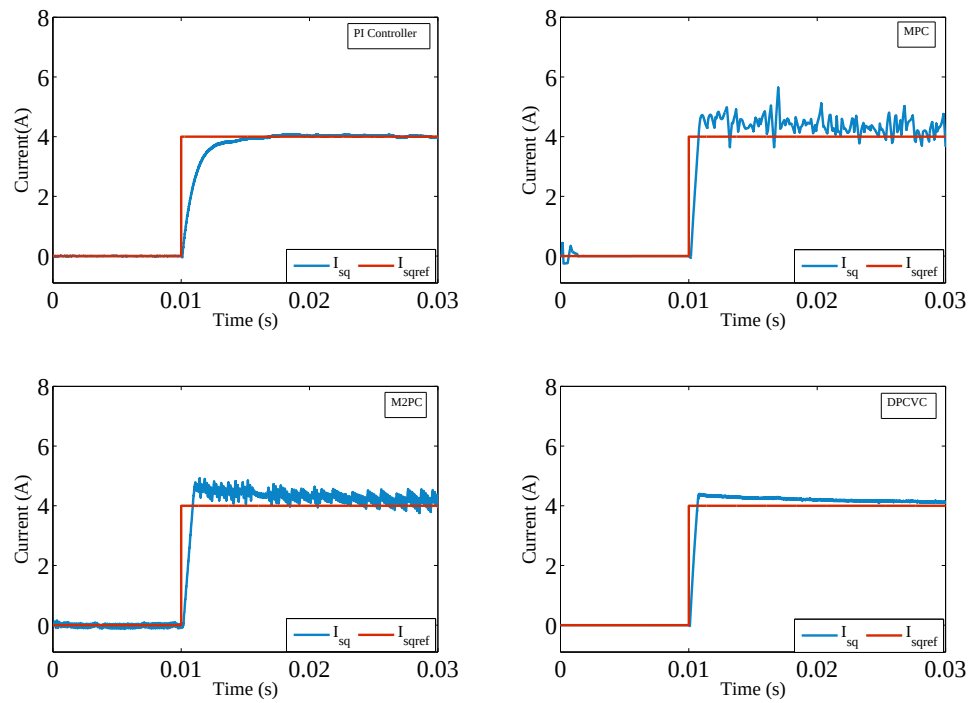


Figure 6.43: Simulation results: q-axis stator current reference tracking when IM is controlled by PI, MPC, M2PC and DPCVC

Figure 6.44 shows the experimental results for reference tracking of q-axis stator currents of an IM controlled by PI controller, MPC, M2PC and DPCVC. The steady state performance of the controllers can be studied from the THD of the controlled waveforms. To analyse the transient performance of the control strategies, the rise time of the controlled currents is considered. Rise time is the time taken by the control variable to reach from 10% to 90% of the steady state value. A comparison of the THDs and rise times during the start up of the machine for different controllers such as PI controller, MPC, M2PC and DPCVC from simulation and experimental results is tabulated in Table 6.7. It can be noted that PI controller and DPCVC resulted

in waveforms with the least distortion with THDs of 3.3% and 3.7% respectively. It can be noted that the simulation and experimental tests for M2PC in this comparison study resulted in waveforms with more distortion than MPC and had stability problems for higher speeds. This could be due to the machine parameter variation, inaccurate modelling of the machine or due to inaccurate duty cycles.

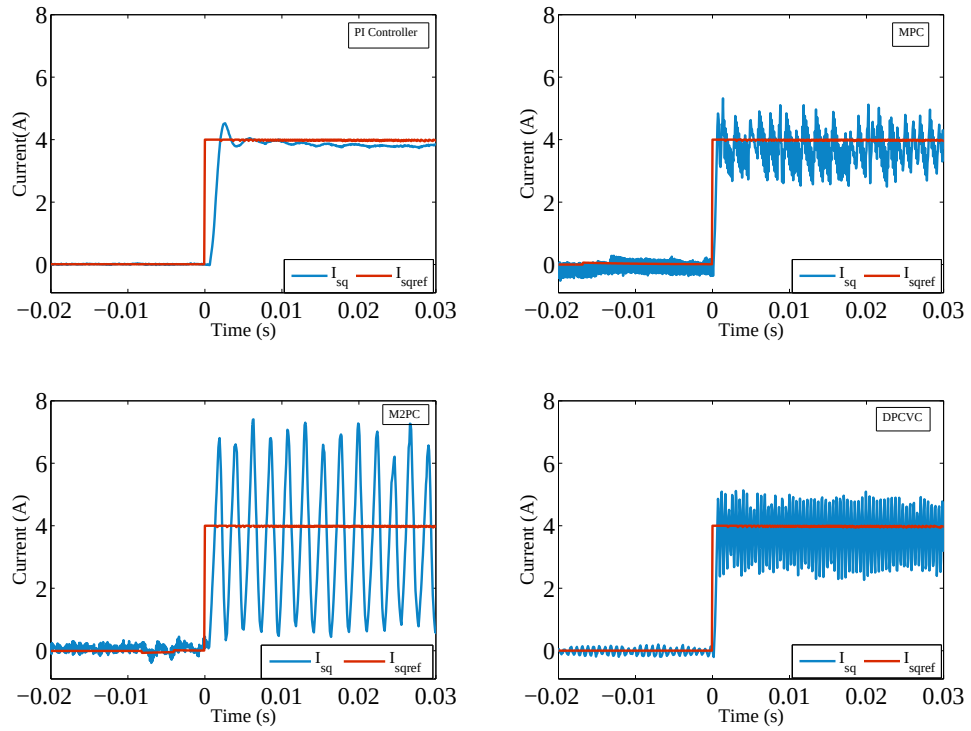


Figure 6.44: Experimental results: q-axis stator current reference tracking when IM is controlled by PI, MPC, M2PC and DPCVC

To conduct a fair analysis of the transient response of the controllers considered, the PI controller is tuned to achieve very fast response to set a benchmark for the comparison. Table 6.7 shows that out of all the controllers tested, the PI controller had the slowest response with a rise time of 1.2ms.

Table 6.7: Induction Machine: Comparison of the steady state and transient performance of different control strategies

Control Strategy	Steady State		Transient Performance	
	Performance (THD)		(Rise time)	
	Simulation	Experiment	Simulation	Experiment
<i>PI Controller</i>	2.2%	3.3%	1.9ms	1.2ms
<i>MPC</i>	5.1%	5.8%	0.46ms	0.36ms
<i>M2PC</i>	5.8%	12.1%	0.60ms	0.62ms
<i>DPCVC</i>	2.4%	3.7%	0.48ms	0.38ms

Predictive control based algorithms like MPC, M2PC and the DPCVC had faster dynamic response with rise times of 0.36ms, 0.62ms and 0.38ms respectively in the experimental results. These results indicate that DPCVC is capable of fast dynamic response very similar to that of MPC. The simulation and experimental results for stator current control of IM proves that DPCVC is capable of producing good quality waveforms similar to PI controller with SVM and fast transient response similar to MPC.

From the above discussion, it can be inferred that out of the two methods that aims at incorporating a modulation scheme with a predictive based control, the DPCVC yields better results compared to the M2PC.

6.6 Summary

Control of a direct matrix converter feeding an RL load and an induction machine using various control strategies and the simulation results for each of them are discussed in chapter 3, 4 and 5. To validate the simulation results of these control methods, experimental tests are conducted in the laboratory and the results are discussed in this chapter.

PI controllers together with SVM strategy for a direct matrix converter is commonly used to get good quality input and output current waveforms. PI controllers can be used to get desired dynamic and steady state performances by selecting the appropriate controller gains. The experimental results for load current control of matrix converter using PI for an RL load and an induction machine are included in this chapter. A cascaded current and speed control using multiple PI controllers is implemented for control of induction motor. The experimental results are comparable to the simulation results for the same control strategy thus validating the controller design.

Model Predictive control has been widely used to control power converters in the last decade and is gaining popularity. The attractive features like fast transient response, multi-objective control capability and high bandwidth have resulted in its application in various applications including electrical drives. The application of MPC for a direct matrix converter feeding and RL load and an induction machine and the corresponding simulation results are discussed in detail in chapter 4. The experimental tests for the same conditions are performed in the laboratory and the resulting waveforms validates the simulation results. This proves that the system model used for the predictive control algorithm is accurate.

Modulated Model Predictive Control is aimed at combining the positive features of MPC and a modulation scheme. Since MPC lacks a modulator, the resulting switching frequency is variable and M2PC attempts to deliver a fixed switching frequency operation. The control strategy is developed for a direct matrix converter and is tested in simulation with an RL load and an induction machine. To validate the control algorithm, M2PC is tested experimentally for a direct matrix converter feeding both RL load and an induction motor. The results for RL load are similar to the simulation results which proves the concept. To demonstrate multi-objective control capability of the control strategy, speed and current of an induction motor is controlled using M2PC and the experimental results does not correspond to the simulation results in terms of waveform quality. There are unresolved issues in the implementation of M2PC related to waveform quality when it comes to an induction motor load. This

needs to be investigated further. A comparison of transients response of the four control strategies discussed in this thesis reveals that M2PC has fast dynamic response similar to MPC. It can be noted that M2PC delivers good transient response and an improved current quality compared to MPC for MC with an RL load.

Another control strategy called Direct Predictive Current-error Vector Control is applied to the MC feeding both RL load and induction machine. This method has an inbuilt modulator to generate the desired voltage vectors in association with the predictive control of load current and will result in fixed switching frequency operation. The duty cycles for the switches of MC are calculated by solving a set of linear equations based on the load current error vectors. The error vectors dictate the amplitude and direction of the control variable from the reference. This control strategy results in good quality waveforms comparable to that of a PI Controller with SVM. The experimental results for this approach conforms to the simulation results mentioned in chapter 5. The transient behaviour of DPCVC is also tested experimentally and shows that it has fast dynamic response similar to MPC.

To summarise, the objective of the thesis was to investigate control strategies for direct MC with a focus on developing a fixed switching frequency predictive based controller. M2PC satisfies all the requirements for MC with RL load and will be a good candidate for applications that requires fast dynamic response and does not demand high power quality. The DPCVC strategy on the other hand produces good quality waveforms with fast dynamic response and results in a fixed switching frequency operation in the case of both RL load and induction machine.

Chapter 7

Conclusions and Future Work

This thesis investigates the performance of a direct matrix converter under different control strategies. A direct matrix converter is an AC/AC converter capable of four quadrant operation in a single stage without any DC link. It consists of nine bidirectional switches which enable power flow in both directions. The absence of a DC link capacitor makes the converter suitable for applications where converter sizing is critical like integrated drives [36]. However, due the absence of an intermediate energy storage element, it is difficult to control a direct matrix converter. An LC input filter with a damping resistor is required to filter the higher order harmonics at the supply side of the converter. To accurately size the input filter, the switching frequency of the converter has to be known. A control strategy involving PI controllers with space vector modulation will have a fixed switching frequency. However, a control strategy like predictive control which offers various advantages like fast dynamic response and multi-objective control results in a variable switching frequency operation. In order to overcome this restriction, two control strategies that incorporates the good features of predictive control and a modulation method are investigated and presented in this thesis.

As a starting point to control a direct MC, the standard controller for instance PI controller is used together with a standard modulation technique such as SVM. The

load currents in dq reference frame are controlled using PI controllers in the case of RL load. When an induction machine load is considered, both speed and stator currents were controlled by PI controllers forming cascaded loops. Simulation tests were carried out to verify the controller design. A modulation method such as SVM ensured that the input and output current of the MC were sinusoidal.

Model predictive control has been used in conjunction with matrix converters to control different variables like current, active and reactive power for various applications as discussed in chapter 4. MPC is a control strategy that executes an optimization algorithm to select the switching states that will produce the least error compared to the reference variable in the next sampling interval. This strategy results in very fast transients and is capable of controlling multiple variables without cascaded control loops. The major drawback of this strategy is the variable switching frequency as the controller can select the same switching state for multiple sampling intervals if it results in minimum error. The advantages of this strategy have encouraged researchers to extend the control strategy to drives and has been successful as discussed in chapter 4. The application of MPC to control the load current of a direct matrix converter feeding an RL load and an induction machine is presented in chapter 4. Simulation tests for load current control of a direct MC in Matlab Simulink validates the model of the system and controller design.

One of the main objectives of this thesis has been to develop the M2PC strategy for a direct matrix converter. M2PC algorithm makes use of the optimization algorithm of a standard MPC method using cost function minimization. The difference from MPC is in the application of the voltage vectors or the number of switching states selected per sampling interval. In M2PC, the reference current is achieved by selecting more than one switching state as in the case of a modulator. The switching states selected are applied for their respective duty cycles within the sampling interval. This results in a fixed switching frequency operation. The calculation of duty cycles depend on the cost functions formed by the voltage vectors resulting from available switching states. Chapter 5 focused on the theory of M2PC for a three phase two-level inverter and a direct matrix converter. The control strategy is verified by simulation tests for

load current control of a direct matrix converter in Matlab Simulink environment. The simulation results indicated that the load current quality is improved compared to MPC for an RL load. The transient performance of the controller is comparable to that of MPC. This verified the theory behind M2PC.

To validate the simulation results discussed in chapters 3, 4 and 5, a laboratory prototype of a direct matrix converter is utilized. The description of the experimental setup is discussed in detail in chapter 6. Load current control of a direct matrix converter is tested with four different control methods namely PI controller, MPC, M2PC and DPCVC. The performance of these control strategies for a RL load and induction machine at no load are examined in detail. The PI controller with SVM gives the best steady state performance and a fast transient response can be achieved by careful tuning of the PI controller gains. Even though MPC strategy falls behind in the steady state performance, a fast transient response and multi-objective control capability was observed. For an RL load, M2PC strategy produced results with a steady state performance superior to that of MPC while maintaining a fast transient response and ability to control multiple variables simultaneously. Even though this was true for matrix converter feeding an RL load, for an induction machine the quality of the stator currents controlled by M2PC were not improved compared to MPC at speeds greater than 300 rpm . This can be attributed to machine parameter variation or the inherent approximation introduced by the calculation of duty cycles in M2PC using the magnitude of the cost functions.

Another control method that achieves fixed switching frequency operation called the DPCVC is discussed in chapter 5 and implemented for load current control of MC in the laboratory. This strategy considers the cost functions/errors as vectors which is used to determine the voltage vectors required for modulation. The duty cycles for the selected voltage vectors are also calculated by solving a set of linear equations in alpha-beta co-ordinates formed by the predicted errors. Since the magnitude and direction of the errors are considered, the resulting duty cycles will be accurate and will be the only existing solution for the set of equations under consideration at any time. The simulation results for load current control of a direct MC feeding an RL load and an

induction machine were presented in chapter 5. The experimental results for RL load and induction machine shown in chapter 6 validates the simulation results presented in chapter 5. The experimental results for both RL load and induction machine verifies that DPCVC is capable of good steady state and fast transient response.

A comparison of the steady state and transient response of all the control strategies discussed in this thesis such as PI control, MPC, M2PC and DPCVC is presented in section 6.5. An analysis of the simulation and experimental results for each case shows that the steady state behaviour is the most improved when the load currents were controlled by PI controllers with SVM to generate the voltage vectors and their respective duty cycles. The MPC resulted in load currents with the worst THD compared to the other control methods. The load currents controlled by DPCVC had improved THD comparable to that of PI Controller with SVM as opposed to MPC and M2PC. A study of the transient behaviour of the currents controlled by the different control methods shows that the PI controller resulted in the slowest response and MPC and DPCVC have the fastest response. The rise times of M2PC was seen to be comparable to that of MPC which is expected considering the predictive control algorithm used in M2PC. Thus an analysis of the experimental results for load current control of a direct matrix converter using different control strategies indicate that the DPCVC is superior to the other control strategies such as MPC and M2PC in terms of the steady state and transient performance.

The achievements of this work can be summarised as follows.

- Proposed two fixed switching frequency model predictive control strategies for a direct matrix converter known as M2PC and DPCVC.
- Simulation tests were carried out showing the application of the M2PC and DPCVC for a direct matrix converter feeding an RL load and an induction machine under no load condition. The simulation results are verified by experimental tests for the same operating conditions and system parameters.
- Application of four different control methods such as PI controller, MPC, M2PC

and DPCVC to a direct matrix converter to demonstrate load current control was presented. The simulation and experimental results are utilised to compare the performance of each control strategy for the same system parameters.

- DPCVC proves to be superior to M2PC in terms of steady state and dynamic performance.

7.1 Future Work

Two strategies with the merits of a modulation scheme such as SVM and a high dynamic controller such as MPC proposed in this work has been tested for a direct matrix converter feeding an RL load and an induction machine. Both M2PC and DPCVC proved to satisfy the main design criterion which is the fixed switching frequency operation. However, further research to study the reliability, robustness and stability of these methods is required. The aspects of each control method that has the potential for further study can be summarised as follows:

- The performance of M2PC applied to a direct matrix converter can be improved by including control of the input current. For a direct matrix converter, due to the coupling between the input and output sides, control of the supply side will have significant effect on the load side waveforms. The supply side of the converter can be controlled by controlling the supply current directly or the input reactive power. Hence in future, the supply side control of a direct matrix converter using the M2PC strategy can be investigated.
- Another strategy known as DPCVC has been proposed for a direct matrix converter and experimental tests were carried out to validate the simulation results for load current control. Since this method considers the cost functions as vectors, an accurate value for the duty cycles can be obtained as opposed to M2PC. This method is verified in the simulation for a matrix converter feeding a RL load and presents better current quality compared to M2PC. In addition

to load current control, it is possible to extend this method for supply current control. This can be done by adding constraint equations to the calculation of duty cycles. The set of linear equations which dictates the constraints for both input current and output current control will have a unique solution set consisting of duty cycles which will ensure sinusoidal current in both input and output side of the converter. Another possible extension of this strategy is to investigate the multi-objective control capability. Since the load current is directly controlled in terms of the errors produced, it maybe possible to add more control variables to form a cumulative error equation.

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Appendix A

Published Papers

The following papers were published with material directly relating to this thesis:

- I Vijayagopal, M.; Zanchetta, P.; Empringham, L.; de Lillo, L.; Tarisciotti, L.; Wheeler, P., "Modulated model predictive current control for direct matrix converter with fixed switching frequency," in Power Electronics and Applications (EPE'15 ECCE-Europe), 2015 17th European Conference on , vol., no., pp.1-10, 8-10 Sept. 2015 doi: 10.1109/EPE.2015.7309347
- II Vijayagopal, M.; Empringham, L.; de Lillo, L.; Tarisciotti, L.; Zanchetta, P.; Wheeler, P., "Control of a direct matrix converter induction motor drive with modulated model predictive control," in Energy Conversion Congress and Exposition (ECCE), 2015 IEEE , vol., no., pp.4315-4321, 20-24 Sept. 2015 doi: 10.1109/ECCE.2015.7310270
- III M. Vijayagopal, L. Empringham, L. de Lillo, L. Tarisciotti, P. Zanchetta and P. Wheeler, "Current control and reactive power minimization of a direct matrix converter induction motor drive with Modulated Model Predictive Control," 2015 IEEE International Symposium on Predictive Control of Electrical Drives and Power Electronics (PRECEDE), Valparaiso, 2015, pp. 103-108. doi: 10.1109/PRECEDE.2015.7395591

Appendix B

Modelling of an induction motor

B.1 Introduction

A mathematical model of the induction motor is required to implement any control strategy. The stator and rotor equations of the machine can be derived from the three phase equivalent circuit of the induction machine and is discussed in the next section.

B.2 Modelling of IM in abc (stationary) reference frame

The stationary/abc frame mathematical model of a balanced three phase IM can be expressed as shown below[124]. The machine model can be derived from Figure.B.1.

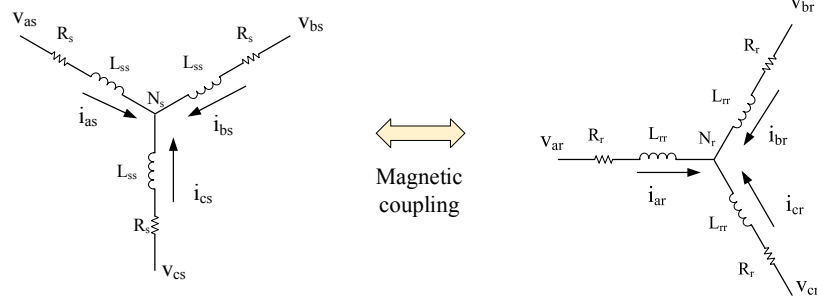


Figure B.1: Three phase IM model

The stator equations are shown in B.1 and the rotor equations are as shown in B.2.

$$v_{abcs} = R_s i_{abcs} + \frac{d\Psi_{abcs}}{dt} \quad (\text{B.1})$$

$$v_{abcr} = R_r i_{abcr} + \frac{d\Psi_{abcr}}{dt} \quad (\text{B.2})$$

where v_{abcs} are the three phase stator voltages, i_{abcs} the three phase stator currents, v_{abcr} three phase rotor voltages, i_{abcr} the three phase rotor currents and R_s and R_r the stator and rotor resistances. Ψ_{abcs} and Ψ_{abcr} are the stator and rotor fluxes respectively.

Using the self and mutual inductances between the stator and rotor windings, the flux linkages can be obtained as shown in B.3.

$$\begin{aligned} \Psi_{as} &= L_{asas}i_{as} + L_{asbs}i_{bs} + L_{ascs}i_{cs} + L_{asar}i_{ar} + L_{asbr}i_{br} + L_{ascr}i_{cr} \\ \Psi_{bs} &= L_{bsas}i_{as} + L_{bsbs}i_{bs} + L_{bscs}i_{cs} + L_{bsar}i_{ar} + L_{bsbr}i_{br} + L_{bscr}i_{cr} \\ \Psi_{cs} &= L_{csas}i_{as} + L_{csbs}i_{bs} + L_{cscs}i_{cs} + L_{csar}i_{ar} + L_{csbr}i_{br} + L_{cscr}i_{cr} \\ \Psi_{ar} &= L_{aras}i_{as} + L_{arbs}i_{bs} + L_{arcs}i_{cs} + L_{arar}i_{ar} + L_{arbr}i_{br} + L_{arcr}i_{cr} \\ \Psi_{br} &= L_{bras}i_{as} + L_{brbs}i_{bs} + L_{brcs}i_{cs} + L_{brar}i_{ar} + L_{brbr}i_{br} + L_{brcr}i_{cr} \\ \Psi_{cr} &= L_{cras}i_{as} + L_{crbs}i_{bs} + L_{crcs}i_{cs} + L_{crar}i_{ar} + L_{crbr}i_{br} + L_{cr cr}i_{cr} \end{aligned} \quad (\text{B.3})$$

where L_{asas} , L_{bsbs} , L_{cscs} , L_{arar} , L_{brar} and L_{crar} are the stator and rotor self-inductances and L_{asbs} , L_{csbs} , ..., L_{crar} are the mutual inductances between the stator and rotor windings.

The self and mutual inductances for the stator and rotor is given by B.4.

$$\begin{aligned} L_{asas} &= L_{bsbs} = L_{cscs} = L_{ss} = L_{ls} + L_{ms} \\ L_{arar} &= L_{brbr} = L_{crar} = L_{rr} = L_{lr} + L_{mr} \end{aligned} \quad (\text{B.4})$$

where L_{ls} and L_{lr} are the stator and rotor leakage inductances respectively. The stator and rotor magnetizing inductances L_{ms} and L_{mr} are equal and will be referred as L_m .

The mutual inductances between the stator and rotor windings can be simplified as follows.

$$\begin{aligned} L_{asbs} &= L_{ascs} = L_{bscs} = \frac{1}{2}L_m \\ L_{arbr} &= L_{arcr} = L_{brcr} = \frac{1}{2}L_m \end{aligned} \quad (\text{B.5})$$

The self and mutual inductances L_s and L_r can be expressed in the form of a matrix as shown in B.6 and B.7 respectively.

$$L_s = \begin{bmatrix} L_{ls} + L_m & -\frac{1}{2}L_m & -\frac{1}{2}L_m \\ -\frac{1}{2}L_m & L_{ls} + L_m & -\frac{1}{2}L_m \\ -\frac{1}{2}L_m & -\frac{1}{2}L_m & L_{ls} + L_m \end{bmatrix} \quad (\text{B.6})$$

$$L_r = \begin{bmatrix} L_{lr} + L_m & -\frac{1}{2}L_m & -\frac{1}{2}L_m \\ -\frac{1}{2}L_m & L_{lr} + L_m & -\frac{1}{2}L_m \\ -\frac{1}{2}L_m & -\frac{1}{2}L_m & L_{lr} + L_m \end{bmatrix} \quad (\text{B.7})$$

The mutual inductances between the stator and rotor windings are given in terms of

the electrical rotor angle (θ_r) with a period of 2π .

$$\begin{aligned}
 L_{asar} &= L_{aras} = L_{bsbr} = L_{brbs} = L_{cscr} = L_{crcs} = L_m \cos(\theta_r) \\
 L_{asbr} &= L_{bras} = L_{bscr} = L_{crbs} = L_{csar} = L_{arcs} = L_m \cos\left(\theta_r + \frac{2\pi}{3}\right) \\
 L_{ascr} &= L_{cras} = L_{bsar} = L_{arbs} = L_{csbr} = L_{brcs} = L_m \cos\left(\theta_r - \frac{2\pi}{3}\right)
 \end{aligned} \tag{B.8}$$

To determine the torque of an induction machine with P number of poles, the following equation can be used.

$$\begin{aligned}
 T_e = -\frac{P}{2} &\left\{ \left[i_{as}\left(i_{ar} - \frac{1}{2}i_{br} - \frac{1}{2}i_{cr}\right) + i_{bs}\left(i_{br} - \frac{1}{2}i_{ar} - \frac{1}{2}i_{cr}\right) + i_{cs}\left(i_{cr} - \frac{1}{2}i_{br} - \frac{1}{2}i_{ar}\right) \sin(\theta_r) \right. \right. \\
 &\left. \left. + \frac{\sqrt{3}}{2} (i_{as}(i_{br} - i_{cr}) + i_{bs}(i_{cr} - i_{ar}) + i_{cs}(i_{ar} - i_{br})) \cos(\theta_r) \right] \right\}
 \end{aligned} \tag{B.9}$$

The mechanical model of the induction machine is given by

$$\begin{aligned}
 J \frac{dw_r}{dt} &= \frac{P}{2} (T_e - T_{load}) - B_m w_r \\
 \frac{d\theta_r}{dt} &= w_r
 \end{aligned} \tag{B.10}$$

where J is the moment of inertia of the motor, B_m the frictional coefficient, w_r and θ_r the electrical speed and position of the motor and T_{load} is the load torque. From the electrical parameters, the mechanical speed and position of the motor can be calculated from the number of poles as shown in B.11.

$$\begin{aligned}
 w_m &= \frac{2}{P} w_r \\
 \theta_m &= \frac{2}{P} \theta_r
 \end{aligned} \tag{B.11}$$

B.3 Model of IM in synchronous rotating frame

The equations of the machine in abc reference frame is computationally demanding as seen from section B.2. This can be simplified by considering the machine model in a rotating reference frame as in the case of indirect rotor flux orientation. In this case, the reference frame is aligned to the rotor flux and rotating at a frequency ω_e which is the frequency of rotor flux. The direct and quadrature axis stator and rotor equations of the machine at a synchronously rotating reference frame is shown in B.12.

$$\begin{aligned}
 V_{sd} &= R_s I_{sd} + \frac{d\Psi_{sd}}{dt} - \omega_e \Psi_{sq} \\
 V_{sq} &= R_s I_{sq} + \frac{d\Psi_{sq}}{dt} + \omega_e \Psi_{sd} \\
 V_{rd} &= R_r I_{rd} + \frac{d\Psi_{rd}}{dt} - (\omega_e - \omega_r) \Psi_{rq} \\
 V_{rq} &= R_r I_{rq} + \frac{d\Psi_{rq}}{dt} + (\omega_e - \omega_r) \Psi_{rd}
 \end{aligned} \tag{B.12}$$

where Ψ_{sdq} and Ψ_{rdq} are the stator and rotor flux linkages respectively, ω_r is the electric rotor angular speed. For rotor flux orientation, $\Psi_{rq} = 0$.

The rotor speed dynamics can be expressed as shown in B.13.

$$\frac{d\omega_r}{dt} = \frac{P}{2J}(T_e - T_l) \tag{B.13}$$

where T_e is the electromagnetic torque and T_l the load torque of the machine.

The electromagnetic torque can be written in terms of the stator and rotor flux linkages and currents as shown in the set of equations in

$$\begin{aligned}
 T_e &= \frac{3P}{4}(\Psi_{sd} I_{sq} - \Psi_{sq} I_{sd}) \\
 &= \frac{3P}{4} \frac{L_m}{L_r} (\Psi_{rd} I_{sq} - \Psi_{rq} I_{sd}) \\
 &= \frac{3P}{4} \frac{L_m}{L_s} (I_{rd} I_{sq} - I_{rq} I_{sd})
 \end{aligned} \tag{B.14}$$

Appendix C

PI Controller Design to control an IM

For vector control of induction machine discussed in Chapter.3, PI controllers are used to control the stator currents and speed of the motor. This section describes the design procedure for both current and speed PI controllers.

C.1 Design of current controller

A block diagram of the current loop to control the stator currents of an induction machine are shown in Figure C.1.

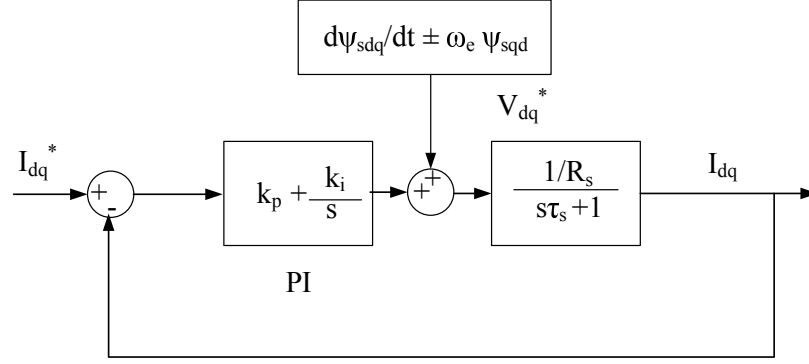


Figure C.1: Block diagram of the current loop

The forward transfer function for the PI controller is given by

$$G_c(s) = k_p + \frac{k_i}{s} \quad (\text{C.1})$$

where s is the Laplace operator, k_p the proportional gain and k_i the integral gain of the PI controller. The compensation terms shown in Figure C.1 are added once the controller gains are calculated using the plant model. From Figure C.1, the characteristic equation for the control loop can be formulated as follows.

$$s^2 + as + b = 0 \quad (\text{C.2})$$

where $a = \frac{k_p + R_s}{\tau_s R_s}$ and $b = \frac{k_i}{\tau_s R_s}$.

For a second order system, the general form of characteristic equation is given by

$$s^2 + 2\xi w_n s + w_n^2 = 0 \quad (\text{C.3})$$

where w_n is the natural frequency and ξ is the damping factor.

Comparing C.2 with C.3, and rearranging the terms, the controller gains k_p and k_i can be obtained.

$$\begin{aligned} k_p &= 2\xi w_n \tau_s R_s - R_s \\ k_i &= w_n^2 \tau_s R_s \end{aligned} \quad (\text{C.4})$$

The stator time constant, τ_s is defined as

$$\tau_s = \frac{\sigma L_s}{R_s} \quad (\text{C.5})$$

σ is the total leakage inductance and is defined as

$$\sigma = 1 - \frac{L_m^2}{L_s L_r} \quad (\text{C.6})$$

With a damping factor of 0.707 and a bandwidth of 100Hz and using the machine parameters given in Table.6.3, the current controller gains can be calculated. The current controller gains obtained from the above calculations are shown in C.7.

$$\begin{aligned} k_p &= 17.1 \\ k_i &= 2000 \end{aligned} \quad (\text{C.7})$$

C.2 Design of speed controller

To design the speed controller for the induction motor, the block diagram in Figure.C.2 is considered.

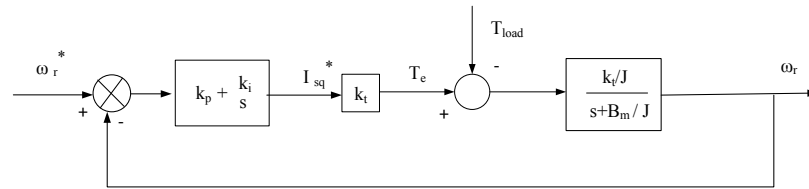


Figure C.2: Block diagram of the speed loop

From Figure.C.2 the characteristic equation for speed controller can be expressed as

$$s^2 + \left(\frac{B_m}{J} + \frac{k_p k_t P}{2J} \right) s + \frac{k_i k_t P}{2J} = 0 \quad (\text{C.8})$$

Comparing C.8 with C.3 and rearranging the terms, the controller gains can be cal-

culated as shown in

$$\begin{aligned} k_p &= \left(2\xi w_n - \frac{B_m}{J} \right) \frac{2J}{k_t P} \\ k_i &= w_n^2 \frac{2J}{k_t P} \end{aligned} \tag{C.9}$$

where $k_t = \frac{3PL_m^2 i_d^*}{4L_r}$ is the torque constant. By selecting a ξ of 0.707 and a bandwidth of 10Hz, the speed controller parameters can be determined. The resulting speed controller gains calculated empirically are as shown in C.10.

$$\begin{aligned} k_p &= 0.2 \\ k_i &= 0.6 \end{aligned} \tag{C.10}$$

Appendix D

Disturbance Analysis of Supply currents of a 30kW integrated drive

The following study was conducted for an integrated matrix converter drive mentioned in [36]. In the integrated drive, the matrix converter is mounted on the motor end plate resulting in a single unit which includes the drive and the motor. The absence of a DC link capacitor in a matrix converter enabled it to be used for applications like this where the size of the converter is crucial.

It was observed that the integrated drive developed in PEMC group at The University of Nottingham had issues regarding the input filter and stability of the system. The problem arises when there is a disturbance from the load/motor. This could be in the form of unbalanced back emf. When subjected to disturbance, the input current becomes distorted and starts resonating. This increases the temperature of the damping resistor and is not an ideal situation. To study this further, the same set up and conditions in the lab were simulated in Matlab Simulink environment. The disturbance is simulated by adding a voltage of frequency $24 * F_o$ to the reference voltage to the MC where F_o is the output frequency. The magnitude of the disturbance

is selected such that the effect on the output current is similar to what was seen in the lab. The test was then conducted for output frequencies ranging from 50-100Hz. The system was run at constant power in all these cases. The distortion in the input current was seen to be increasing with increasing output frequencies as shown in Figure.D.1.

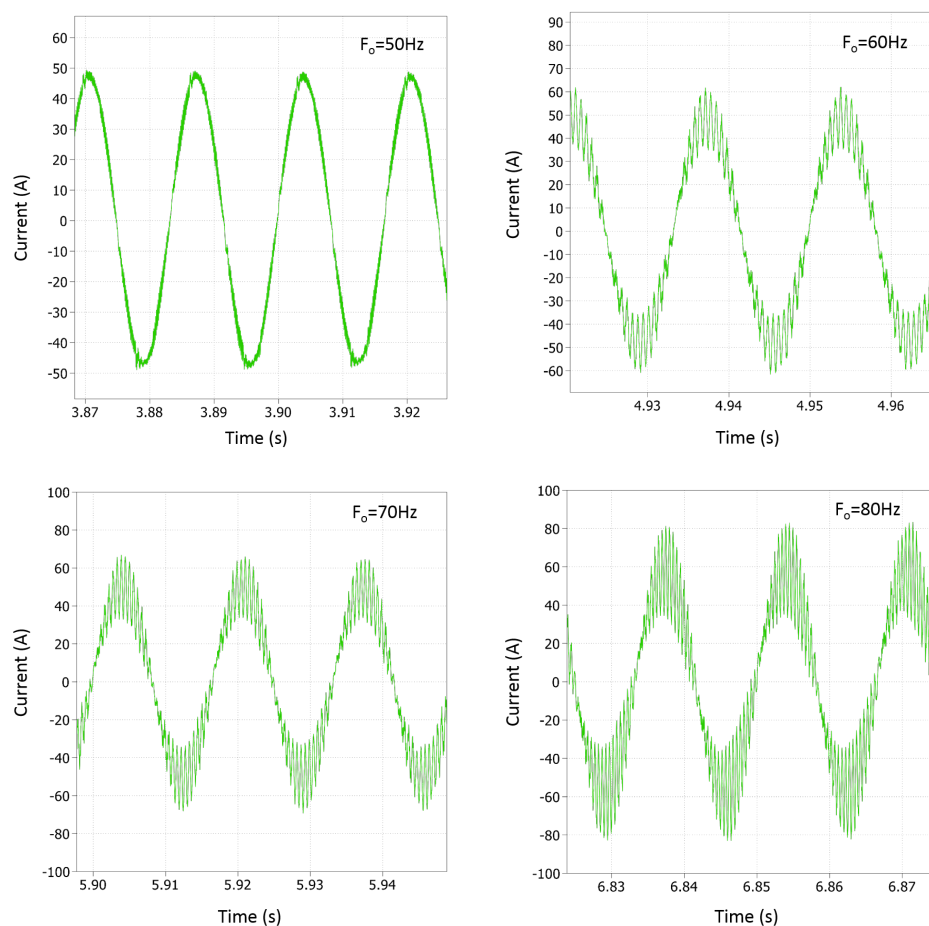


Table D.1: Supply currents for different output frequencies

The next step was to do a harmonic analysis of these input current waveforms and determine the dominant harmonics in each case. The findings of the harmonic analysis is summarised in Table.D.2.

<i>Output Frequency</i>	<i>Dominant Harmonic 1</i>	<i>Dominant Harmonic 2</i>	<i>Disturbance Frequency</i>	<i>Offset</i>
40	940	1060	960	20
50	1190	1310	1200	10
60	1440	1560	1440	0
70	1690	1810	1680	10
80	1940	2060	1920	20
90	2190	2310	2160	30

Table D.2: Disturbance analysis of supply current

When these harmonics are observed carefully it is noticed that there is a relation between the harmonics generated in addition to the fundamental supply frequency, F_i and the disturbance as shown in D.1. (In practice the disturbance is unknown).

$$\begin{aligned}
 \text{Dominant harmonic 1} &= \text{Disturbance freq} + (F_o - F_i) \\
 \text{Dominant harmonic 2} &= \text{Dominant harmonic 1} + 2 * F_i
 \end{aligned}
 \tag{D.1}$$

In order to eliminate these harmonics, one or both of the dominant harmonics were injected into the reference voltage waveform antiphase to the reference voltage. However this resulted in more harmonic bands together with dominant harmonics. Therefore it was concluded that a different approach is required to tackle this problem. It was observed that the SVM strategy for the matrix converter that calculates the input and voltage sectors will not be affected with the addition of harmonics in antiphase. Including the distorted input current for sector calculations is required to see its effect on the modulation. To test this, the dominant harmonics in anti phase was added to the input voltage for the input current angle calculation so that the input current angle will also include the disturbance. This resulted in an additional harmonic in the input current together with the dominant ones. Thus it was concluded that any type of harmonic injection will only result is more harmonics which is not desirable.

This analysis has led to the conclusion that the distortions in back emf of the in-

duction machine which results in distortions in the supply side waveforms cannot be eliminated by harmonic injection. Since the distorted back emf is a result of poor design of the commercial machine used for integrated drive, it is concluded that this can only be eliminated by designing the machine specifically for integrated drive which incorporates the desired requirements.