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ADDIS ABABA UNIVERSITY
SCHOOL OF GRADUATE STUDIES
FUCULTY OF TECHNOLOGY
DEPARTMANT OF
ELECTRICAL AND COMPUTER ENGINEERING

DSP BASED VECTOR CONTROL OF INDUCTION MOTOR

BY

ABRAHAM BELAY

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ABSTRACT

In this thesis Vector control of Induction Motor using DSP board, type TMS320LF2407 from Wintech digital instruments has been studied and implemented. This new family of DSP controllers enables cost-effective design of intelligent controllers for Induction motors, which can yield enhanced operation, fewer system components, lower system cost and increased efficiency. The vector control Algorithm maintains efficiency in a wide speed range and processes the dynamic model of the Induction Motor.

The DSP board is successfully programmed by using C and assembly language in a code composer studio working environment, interfaced with an inverter of type DL2646 from DELORENZO and the Induction motor has been driven by V/f control using SVPWM technique. Experimental and simulation results has been presented and discussed.

Chapter One

INTRODUCTION

1.1 General Introduction

For many years d.c motors were used extensively in areas where variable speed operation was required since their flux and torque could be controlled independently and easily by the field and armature current. Particularly, the separately excited d.c motor has been used mainly for applications where there was a requirement of fast response and four-quadrant operation with high performance near zero speed. However, due to the existence of the commutator and the brushes d.c motors have certain disadvantage. That is, they required periodic maintenance; they cannot be used in explosive or corrosive environments. These problems can be overcome by the application of a.c motors, because of its simplicity, ruggedness, efficiency, low cost, compactness, and economical and volume manufacturing advantages.

Some of the reasons for dealing with AC induction motor drives rather than DC motor drives are:

- ✧ Commutator and Brushes do not exist in Induction motors, which can result in higher speed operations.
- ✧ Induction motors have comparable and frequently better efficiency than the equivalent dc motor.
- ✧ Power Factor; the power factor of a dc drive operates in the range from 0 to about 0.9 and increases approximately in proportion to the motor speed. However almost all induction motors run at lagging power factor, typically 0.95.
- ✧ A dc motor must be regularly taken out of service to check or replace brushes and at less frequent interval to resurface the commutator. Induction motor is essentially maintenance free, except for the bearings.
- ✧ Protection of an induction motor is simpler than for dc motor. For large dc machines protection of a dc circuit breaker is required which are expensive.

Simpler stator thermal protection for induction motors gives comprehensive thermal protection for the entire machine.

- ✧ Ruggedness; Induction motor is more robust than the dc machine.
- ✧ Because of the commutator assembly, a dc machine of the same torque capacity is significantly larger than the equivalent induction machine.

Hence induction machine is superior to d.c machine with respect to size, weight, rotor inertia, efficiency, maximum speed, reliability, cost e.t.c, but because of its highly non-linear dynamic structure with strong dynamic interaction, it requires more complex control schemes than, a separately excited d.c machine.

The induction motor had been considered to be essentially a constant speed motor control of its speed has been possible, but the method and equipment been awkward and its control performance inferior, the principal reason being that the variable-frequency ac power supply was expensive. It is because of the awkward ac power supplies that the induction motor has been limited in its control performance, although the induction motor itself has a very high control capability. Where first used, control of the induction motor was restricted to speed control, which was by means of frequency and/or voltage control. By changing the frequency of the power supply synchronous speed of the induction motor could be changed. By changing the terminal voltage of the induction motor, its speed under load could be changed. If the frequency and voltage were changed proportionally to each other, the magnetic flux density could be kept almost constant, resulting in better control performance.

It should be pointed out that control of the electrical motor is actually control of the power supply. The motor can be controlled only through the power supply. If the power supply is controlled properly, the motor responds properly. Controllability of the power supply is more important, because that of the motor is not a limiting factor. For the classical period of induction motor control, controllability of the ac power supply was very inferior to that of the dc power

supply. This was only one reason that the dc motor was used predominantly for quick response purposes, inspite of the fact that the dc motor itself was so awkward in its structure and capability.

The general dynamic model of the induction machine can be represented by a sixth-order state-space equation, where the inputs to the stator are voltage and frequency and the outputs can be rotor speed, rotor position, electromagnetic torque, stator or rotor flux linkages, magnetizing current or a combination of these. Furthermore the cost of a.c power converters is higher for the variable-speed induction motor drive than for the converters, which can supply d.c machines. Until recently the cost of the variable-speed induction motor has been prohibitive and the complexity of control has made its developments in the field power electronics, whereby better and more powerful semiconductor devices are available and where the power devices and circuits are packaged into the modular form, and the existence of powerful and inexpensive microprocessors, which allow the complex control functions of the a.c drive to be performed by utilizing software instead of expensive hardware, means that a.c drives employing induction machines can be considered as economical alternatives to adjustable –speed d.c drives.

Vector control techniques incorporating fast μ Ps have made possible the application of induction-motor drives for high performance application where traditionally only d.c drives were applied. In the past such control techniques could have not been possible because of the complex hardware and software required to solve the complex control problem. As for d.c machines, torque control in a.c machines is achieved by controlling the motor currents. However, in contrast to a d.c machine, in an a.c machine, both the phase angle and the modulus of the current has to be controlled, or in other words, the current vector has to be controlled. Furthermore, in d.c machines, the orientation of the field flux and armature mmf is fixed by the commutator and the brushes, while in a.c

machines the field flux and the angle of the armature mmf require external control.³ With vector control of a.c machines, the torque and flux-producing stator current components are decoupled and transient response characteristics are similar to those of a separately excited d.c machine and the system will adapt to any load disturbances and/or reference value variation as fast as a d.c machine.

It is the ²real time-processing properties of silicon, such as the TMS320LF2407 DSP controller also called Wintech TDS2407EA,² and the accurate asynchronous motor model that have resulted in the development of a highly reliable drive with highly accurate and variable speed controls. Application of the vector control to the a.c induction drive results in the instance control of a high performance drive. The ²ability to achieve such control renders the asynchronous drive a high advantageous system for both home appliances and for industrial or automotive applications [15]. Key advantages are the robust nature of the drive, its reliability and efficiency, the cost effectiveness of both the motor and the drive, the high torque and zero speed, the speed variation capacity, the extended speed range, the direct torque and flux control and the excellent dynamic behavior.

This thesis describes simulation,² complete integration of the software, and the theoretical and practical aspects of the application.

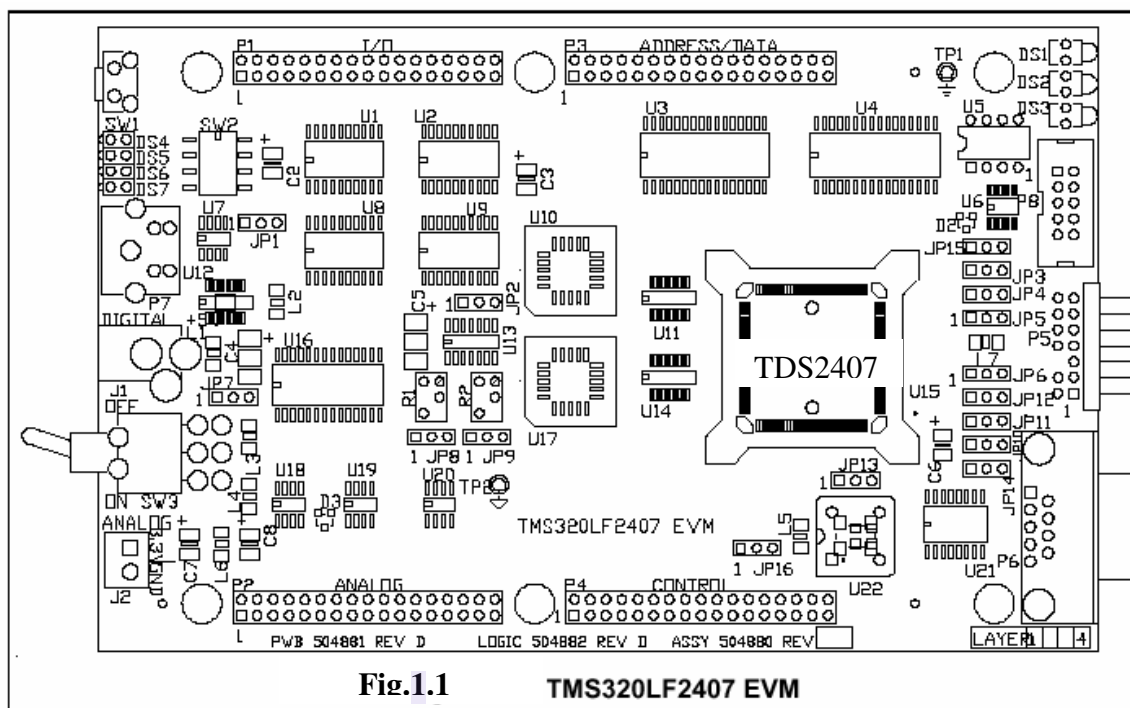
1.2 Advantages of using the DSP techniques in motor control:

- ✧ ³Cost reduction in control electronics;
- ✧ Improved reliability due to the reduction of the number of components;
- ✧ Standard universal hardware is required and the only changes are to the software, which is the very flexible and can easily modified;
- ✧ Digital transmission requires a minimal amount of cabling and is very tolerant to noise; it eliminates drift and electromagnetic interference problems;
- ✧ ³Very high accuracy, excellent repeatability, linearity and stability with different setting ranges;

- ❖ ³ Complex, high-speed arithmetic and capability of decision making e.t.c
- ❖ ³ Only a few standard modules are required without any special adjustment;
- ❖ There is the possibility of plug-in memory modules for user programs, parameters and modifications to functions, e.t.c;
- ❖ Powerful system software for on-line measurements, control parameter setting (example: current control parameter setting, speed control parameter setting e.t.c) and testing;
- ❖ Automatic location of hardware faults with the help of system and user software.

The control hardware is the TMS320LF24X evaluation Module introduced by Wintech Digital Instruments. This module is a ³⁹ member of the TMS320C28x™ DSP generation, are highly integrated, high-performance solutions for demanding control applications.

The board contains a ³³ DSP controller TMS320LF2407 and its oscillator, a JTAG Emulator TDS510USB, and an RS232 link with the necessary output connectors. The figure below depicts the EVM board:



The ²EVM board provide access to any signal from the DSP controller and contains test LED's and Digital to Analog converters. These characteristics are particularly interesting during the development stages [15].

The pin out diagram is shown below:

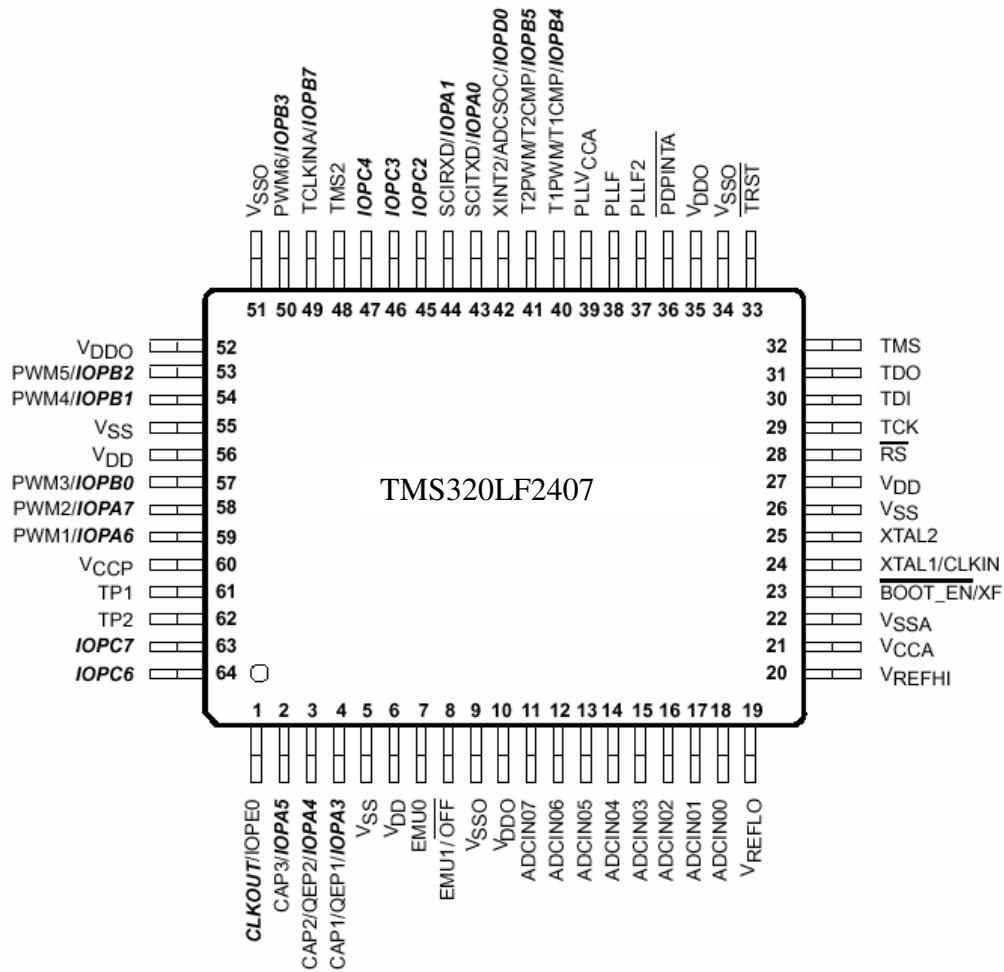


Fig.1.2 pin out diagram for TMS320LF 2407

Here are the main Key Features of TDS2407 EA:

- TMS320LF2401A Digital Signal Processor
- JTAG emulator
- 30²⁷ MIPS operating speed
- 1K words on-chip RAM
- 8K words on-chip Flash memory

- Onboard 7.28 MHz. CPU clock from gate array
- 2 Expansion Connectors (analog, I/O)
- 27-volt only operation with supplied AC adapter
- TI Code Composer tools driver

The TDS2407 EA⁴⁰ consists of four major blocks of logic:

- Analog Interface Connector
- I/O Interface Connector
- Parallel Port JTAG Controller Interface
- Voltage Control

1.3 The Inverter

Also I have used the inverter for adjustable speed; the¹⁰ process of converting dc to ac power is called inversion and it is the inverter, which creates the variable frequency from the dc source, which is used to drive an induction motor at a variable speed. In general, two basic types of inverters exist which are totally different in behavior.

VSI (Voltage Source Inverter) , which is more commonly used one and this type of inverter creates a relatively well defined switched voltage waveform at the terminals of the motor. The dc bus is maintained as voltage stiff by the use of a large capacitor in the dc link. In my thesis work I have used DE LORENZO VSI DL 2646[13].

CSI (³⁰Current Source Inverter), provides a switched current waveform at the motor terminals. The dc bus is maintained as current stiff by use of a large inductor in the dc link.

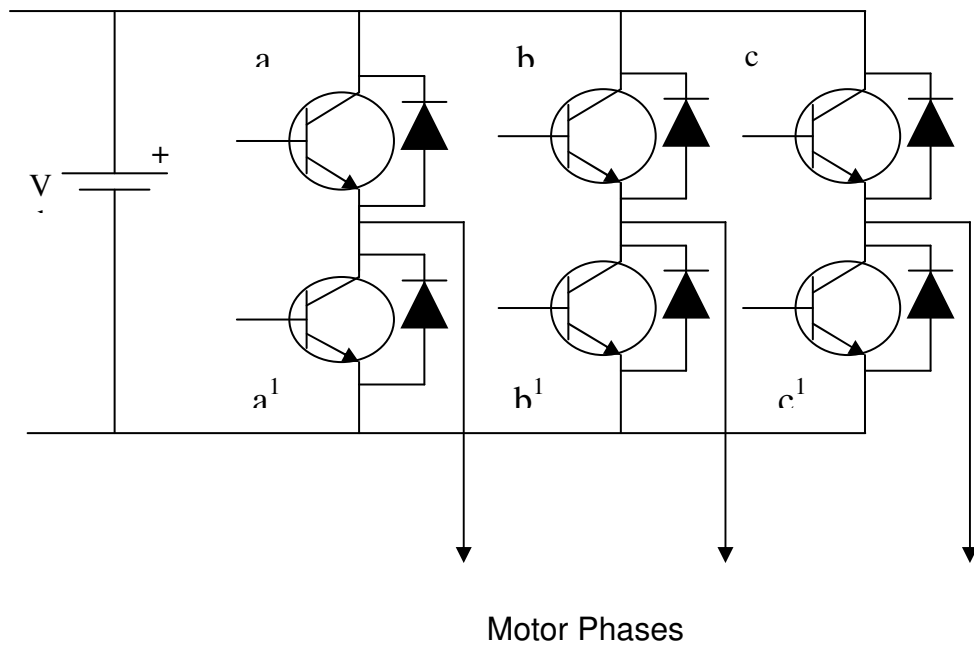


Fig1.3 Voltage source PWM Inverter

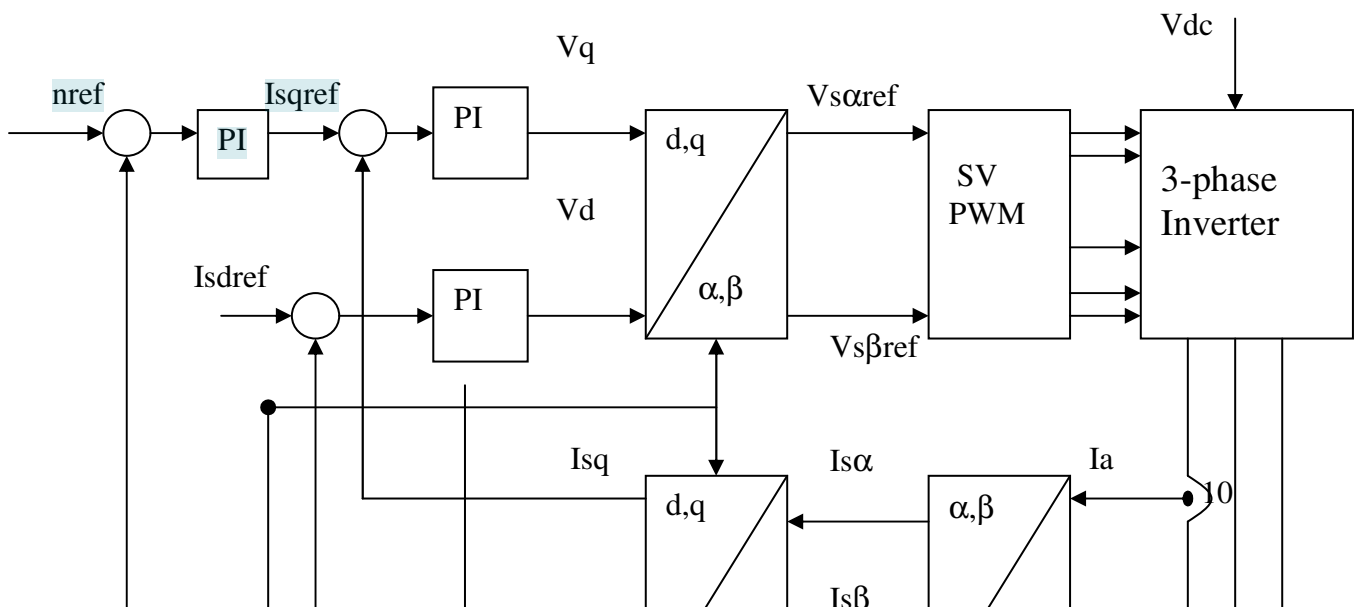
CHAPTER TWO

17 Vector control of Induction Motor

2.1 Complete Vector Control of Induction Motor Structure

The speed control of the AC induction drive is often split into two ranges: the constant torque, where the motor speed is below the nominal speed, and the constant power, where the motor speed is higher than the nominal speed. Above the nominal speed the effective back electromotive force (which depends on both the motor speed and on the rotor flux) is high enough, given the d.c bus voltage limitation, to limit the current in the winding[12]. As such, this limits both the torque production and the drive efficiency (due to problems with magnetic saturation and heat dissipation). Where the rotor flux has been maintained as its nominal value during the low speed operation so as to achieve the highest mutual torque production, it must be reduced in the high-speed operation in order to avoid magnetic saturation and generation of high back electromotive force.

These two induction motor control dedicated modules are added to the basic vector control structure which results in the following complete vector control AC induction drive structure:



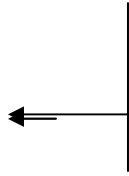


Fig.2.1 complete DSP based vector control of IM representation

12 Two phase currents feed the Clarke transformation module, these projection outputs are indicated $i_{s\alpha}$ and $i_{s\beta}$. These two components of the current provide the input of the park transformation that gives the current in the d, q rotating reference frame. The i_{sd} and i_{sq} components are compared to the references i_{sdref} (the flux references) and i_{sqref} (the torque reference). The torque command i_{sdref} indicates the right rotor flux command for every speed reference. The current regulator outputs are V_{sdref} and V_{sqref} ; they are applied to the inverse park transformation. The outputs of this projection are $V_{s\alpha ref}$ and $V_{s\beta ref}$, the components of the stator vector voltage in the α, β orthogonal reference frame. These are the inputs of the space vector PWM. The outputs of this block are the signals that drive the inverter.

Note that both park and inverse park transformation require the rotor to be in flux position, which is given by the current model block.

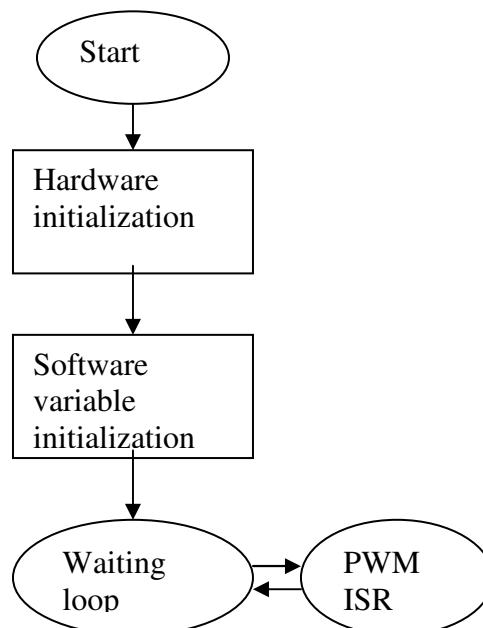


Fig 2.2 General Software Flow Chart

2.2 DYNAMIC MODEL OF INDUCTION MOTORS FOR VECTOR CONTROL

Although traditional per-phase equivalent circuit shown in fig, 2.3 has been widely used in steady-state analysis and design of induction motors, it is not appropriate to predict dynamic performance of the motor. In order to understand and analyze vector control of induction motors, the dynamic model is necessary.

The induction motor, which is the most widely used motor type in the industry, has been favored because of its good self-starting capability, simple and rugged structure, low cost and reliability, etc. Along with variable frequency AC inverters, induction motors are used in many adjustable speed applications, which do not require fast dynamic response. The concept of vector control has opened up a new possibility that induction motors can be controlled to achieve dynamic performance as good as that of DC or brushless DC motors. In order to understand and analyze vector control, the dynamic model of the induction motor is necessary. It has been found that the dynamic model equations developed on a rotating reference frame is easier to describe the characteristics of induction motors. It is the objective of the article to derive and explain induction motor model in relatively simple terms by using the concept of space vectors and d-q variables. It will be shown that when we choose a synchronous reference frame in which rotor flux lies on the d-axis, dynamic equations of the induction motor is simplified and analogous to a DC motor.

Throughout the article, all vectors are denoted as boldface and complex conjugates are denoted by *. Vectors on a rotating reference frame is followed by

a superscript letter which designates the frame used as in V_s^s (V_s in stationary frame). The derivative operator is denoted by p while P is the number of poles. For notational convenience, let Y (scalar) or \mathbf{Y} (vector) be the representative notation of any voltage, current or flux linkage variable. Real and Imaginary values of a space vector \mathbf{Y} is denoted by $\text{Re}(\mathbf{Y})$ and $\text{Im}(\mathbf{Y})$, respectively. Zero vectors are denoted by 0 regardless of the reference frame used.

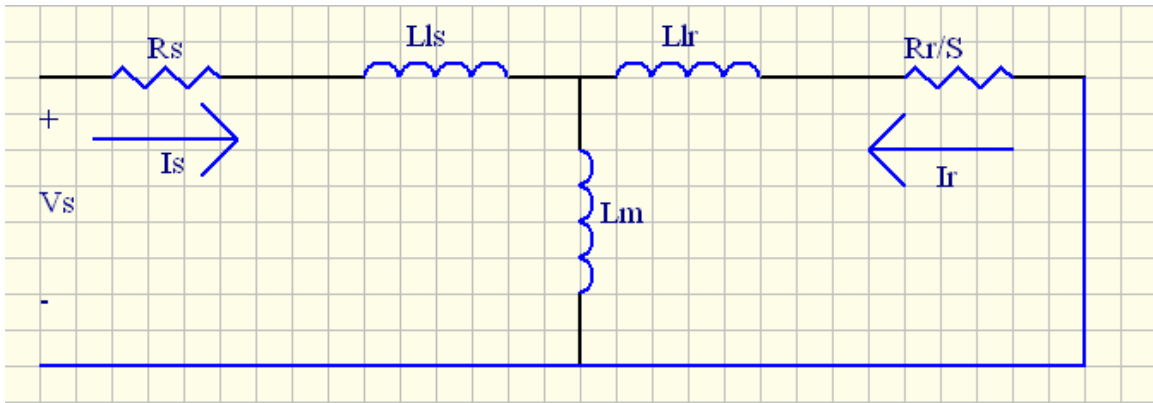


Fig. 2.3 Conventional Per-phase Equivalent Circuit

DYNAMIC MODEL IN SPACE VECTOR FORM

In an induction motor, the 3-phase stator windings are designed to produce sinusoidally distributed mmf in space along the air gap periphery. Assuming uniform air gap and neglecting the effects of slot harmonics, distribution of magnetic flux will also be sinusoidal.

For a sinusoidal 3-phase quantity of constant rms value, the corresponding space vector is a constant-magnitude vector rotating at the frequency of the sinusoid with respect to the fixed (stationary) reference frame. Note that the space vector is at vector angle 0 when A-phase signal (Y_a) is at its sinusoidal peak value in steady state. With space vector notation, voltage equations on the stator and rotor circuits of induction motors are[9],

$$V_{ss} = R_s I_{ss} + p\lambda_{ss} \quad (2.1)$$

$$V_r' = R_r' I_r' + p \lambda_r' = 0 \quad (2.2)$$

It is very convenient to transform actual rotor variables (V_r' , I_r' , λ_r') from Eq. 2.2 on a rotor reference frame into a new variables (V_{rs} , I_{rs} , λ_{rs}) on a stator reference frame as in the derivation of conventional steady-state equivalent circuit. Let the stator to rotor winding turn ratio be n and the angular position of the rotor be θ , and define

$$I_{rs} = (1/n) \exp(j\theta) I_r', \lambda_{rs} = n \exp(j\theta) \lambda_r' \quad (2.3)$$

Also, by defining referred rotor impedances as $R_r = n^2 R_r'$, etc., we have

$$V_{ss} = R_s I_{ss} + p \lambda_{ss} \quad (2.4)$$

$$0 = R_r I_{rs} + (p - j\omega_0) \lambda_{rs} \quad (2.5)$$

where $\omega_0 = p \theta_0$, is the speed of the motor in electrical frequency unit and

$$\lambda_{ss} = L_{ss} I_{ss} + L_m I_{rs} \quad (2.6)$$

$$\lambda_{rs} = L_m I_{ss} + L_r I_{rs} \quad (2.7)$$

The above 4 equations (Eq. 2.4 - 2.7) constitute a dynamic model of the induction motor on a stationary (stator) reference frame in space vector form. These model equations may be simplified by eliminating flux linkages as

$$V_{ss} = (R_s + L_s p) I_{ss} + L_m p I_{rs} \quad (2.8)$$

$$0 = (R_r + L_r (p - j\omega_0)) I_{rs} + L_m (p - j\omega_0) I_{ss} \quad (2.9)$$

From Eqs. 2.8-2.9, The dynamic equivalent circuit model on a stationary reference frame can be drawn as in Fig.2.4. For steady-state operation with

excitation frequency ω_e , p in Eq. 2.8-2.9 may be replaced by $j\omega_e$ and after some algebraic manipulation, we get

$$V_{ss} = (R_s + j\omega_e L_s)I_{ss} + L_m p I_{rs} \quad (2.10)$$

$$0 = (R_r / s + j\omega_e L_r)I_{rs} + j\omega_e L_m I_{ss} \quad (2.11)$$

1 Which exactly describes the conventional steady-state equivalent circuit of Fig. 2.3.

Now, the previous procedure can be generalized so that the dynamic model is described on an arbitrary reference frame rotating at a speed ω_a , where Eq. 2.4 - 2.11 is a special case with $\omega_a=0$. To do that, define the new space vector on the arbitrary frame as

$$Y_a = \exp(-j\theta_a)Y_s \quad (2.12)$$

11 and reconstruct all the model equations in terms of the new space vectors. In the arbitrary reference frame, Eqs. 2.4-2.5 are modified to

$$V_{sa} = (R_s + L_s p)I_{sa} + L_m p I_{ra} + j\omega_a \lambda_{sa} \quad (2.13)$$

$$0 = (R_r + L_r p)I_{ra} + L_m p I_{sa} + j(\omega_a - \omega_0)\lambda_{ra} \quad (2.14)$$

with new flux linkage equations defined by,

$$\lambda_{sa} = L_s I_{sa} + L_m I_{ra} \quad (2.15)$$

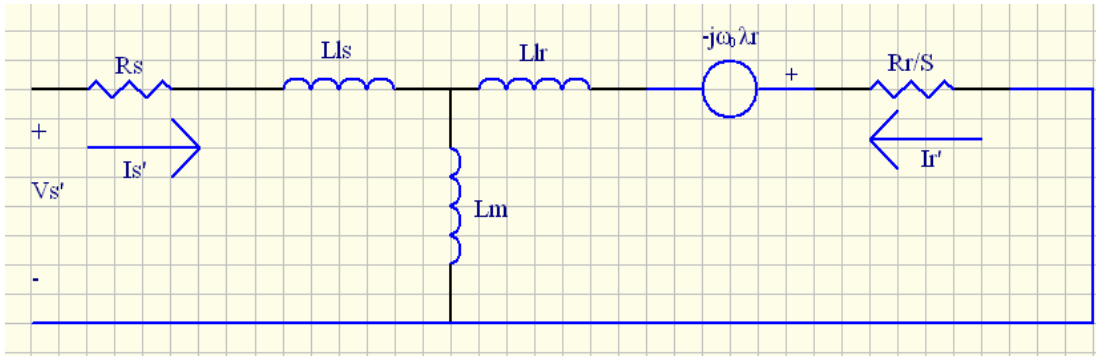
$$\lambda_{ra} = L_m I_{sa} + L_r I_{ra} \quad (2.16)$$

1 As before, by substituting Eqs. 2.14-2.15 into Eqs. 2.12-2.13, we have

$$V_{sa} = (R_s + L_s(p + j\omega_a))I_{sa} + L_m(p + j\omega_a)I_{ra} \quad (2.17)$$

$$1 \quad 0 = (R_r + L_r(p + j\omega_a - j\omega_0))I_{ra} + L_m(p + j\omega_a - j\omega_0)I_{sa} \quad (2.18)$$

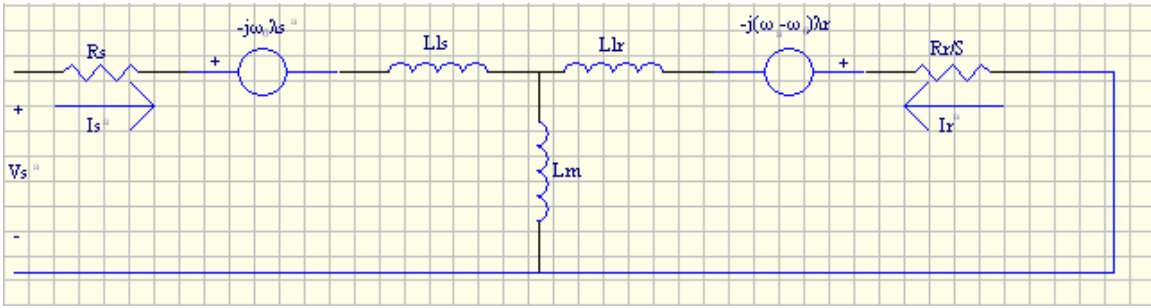
where flux linkage variables are eliminated.



1 Fig. 2.4 Dynamic Equivalent Circuit on a Stationary Reference Frame

The generalized equivalent circuit on an arbitrarily rotating frame based on Eq. 2.17-2.18 is shown in Fig. 2.5. Now, depending on a specific choice of ω_a , many forms of dynamic equivalent circuit can be established. Among them, the synchronous frame form can be obtained by choosing $\omega_a = \omega_e$. This form is very useful in describing the concept of vector control of induction motors as well as of PM synchronous motors because at this rotating frame, space vector is not rotating, but fixed and have a constant magnitude in steady-state. Since space vectors in the synchronous frame will frequently be used, they are denoted without any superscript indicating the type of frame.

Another possible reference frame used in vector control is the rotor reference frame by choosing $\omega_c = \omega_0$ which is in fact, the reverse step of Eq. 2.3 with $n = 1$.



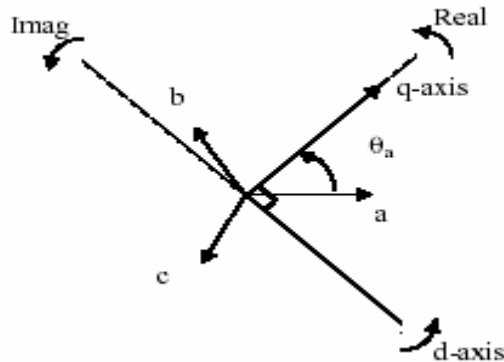
1 Fig. 2.5 Dynamic Equivalent Circuit on an Arbitrary Reference Frame Rotating at ω_a .

D-Q EQUIVALENT CIRCUIT

In many cases, analysis of induction motors with space vector model is complicated due to the fact that we have to deal with variables of complex numbers. For any space vector Y , define two real quantities S_q and S_d as,

$$S = S_q - jS_d \quad (2.19)$$

1 In other words, $S_q = \text{Re}(S)$ and $S_d = -\text{Im}(S)$. Fig. 2.6 illustrates the relationship between d-q axis and complex plane on a rotating frame with respect to stationary a-b-c frame. Note that d- and q-axes are defined on a rotating reference frame at the speed of $\omega_a = p \theta_a$ with respect to fixed a-b-c frame.



11 Fig. 2.6 Definition of d-axis and q-axis on an arbitrary reference frame

With the above definition, Eq. 2.17-2.18 can be translated into the following 4 equations of real variables expressed in a matrix form.

$$\begin{bmatrix} V_{qs}^a \\ V_{ds}^a \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} R_s + L_{sp} & \omega_a L_s & L_{mp} & \omega_a L_m \\ -\omega_a L_s & R_s + L_{sp} & -\omega_a L_m & L_{mp} \\ L_{mp} & (\omega_a - \omega_0)L_m & R_r + L_{rp} & (\omega_a - \omega_0)L_r \\ -(\omega_a - \omega_0)L_m & L_{mp} & -(\omega_a - \omega_0)L_r & R_r + L_{rp} \end{bmatrix} \begin{bmatrix} I_{qs}^a \\ I_{ds}^a \\ I_{qr}^a \\ I_{dr}^a \end{bmatrix} \quad (a)$$

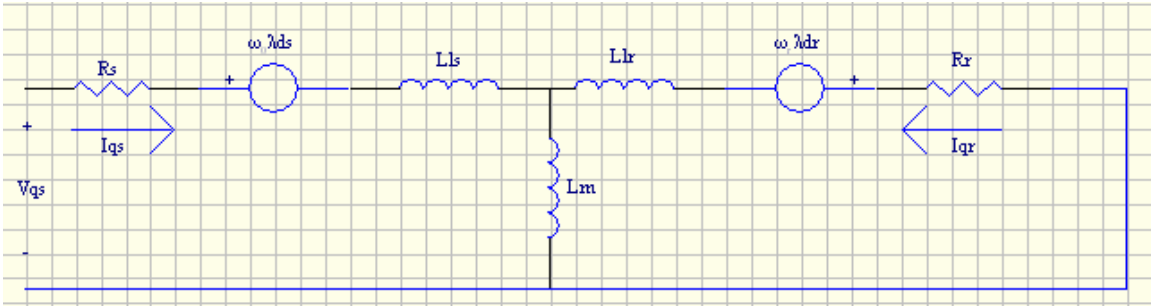
For future reference, the above matrix equation simplified for popular reference frames in analysis and design of vector control will be introduced. For stationary reference frame, by substituting $\omega_a = 0$, the above equation is reduced to

$$\begin{bmatrix} V_{qs}^s \\ V_{ds}^s \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} R_s + L_{sp} & 0 & L_{mp} & 0 \\ 0 & R_s + L_{sp} & 0 & L_{mp} \\ L_{mp} & -\omega_0 L_m & R_r + L_{rp} & -\omega_0 L_r \\ \omega_0 L_m & L_{mp} & -(\omega_a - \omega_0)L_r & R_r + L_{rp} \end{bmatrix} \begin{bmatrix} I_{qs}^s \\ I_{ds}^s \\ I_{qr}^s \\ I_{dr}^s \end{bmatrix} \quad (b)$$

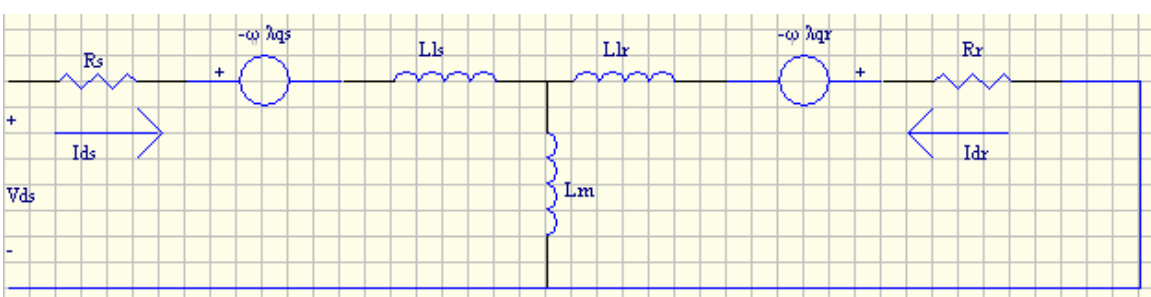
Some implementation of vector drive includes calculation in rotor reference frame (frame is attached to the rotor rotating at ω_0). In this case, we can substitute all ω_a by ω_0 , which makes simplified rotor voltage equations. Moreover, for synchronous frame, we have

$$\begin{bmatrix} V_{qs} \\ V_{ds} \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} R_s + L_{sp} & \omega_0 L_s & L_{mp} & \omega_0 L_m \\ -\omega_0 L_s & R_s + L_{sp} & -\omega_0 L_m & L_{mp} \\ L_{mp} & -\omega_0 L_m & R_r + L_{rp} & -\omega_0 L_r \\ -\omega_0 L_m & L_{mp} & -\omega_0 L_r & R_r + L_{rp} \end{bmatrix} \begin{bmatrix} I_{qs}^s \\ I_{ds}^s \\ I_{qr}^s \\ I_{dr}^s \end{bmatrix} \quad (c)$$

As mentioned before, each variable (voltage, current or flux linkage) in the synchronous frame is stationary and fixed to a constant magnitude in steady state. Based on Eq. (c), dynamic d-q equivalent circuit is shown in Fig. 2.7.



(A) Q-axis Circuit



(B) D-axis Circuit

Fig. 2.7 D-Q Equivalent Circuit on a Synchronous Frame

For dynamic simulation of induction motors, Eq. (b) or Eq. (c) may be used. In this case, one may prefer to use the standard form of differential equation as

$$pX = AX + BU \tag{2.20}$$

For Eq. (c), matrix quantities on the above equation are as follows.

$$X = \begin{bmatrix} Iqs \\ Ids \\ Iqr \\ Idr \end{bmatrix} \quad U = \begin{bmatrix} Vqs \\ Vds \\ 0 \\ 0 \end{bmatrix} \quad B = 1/\Delta \begin{bmatrix} Lr & 0 & -Lm & 0 \\ 42 & Lr & 0 & -Lm \\ -Lm & 0 & Ls & 0 \\ 0 & -Lm & 0 & Ls \end{bmatrix}$$

And

$$A = 1/\Delta \begin{bmatrix} R_s L_r & \omega_0 L_s L_r - \omega_r L_m^2 & -R_r L_m & \omega_0 L_m L_r \\ \omega_r L_m^2 - \omega_0 L_s L_r & R_s L_r & -\omega_0 L_r L_m & -R_r L_m \\ -R_s L_m & -\omega_0 L_s L_m & R_r L_s & \omega_r L_s L_r - \omega_0 L_m^2 \\ \omega_0 L_s L_m & -R_s L_m & \omega_0 L_m^2 - \omega_r L_s L_r & R_r L_s \end{bmatrix}$$

Although both Eq. (c) and Eq. 2.20¹ are frequently used to describe the induction motor on a synchronous frame, we need another set of equations that include flux linkage variables to explain the concept of vector control. By translating Eq. 2.23 - 2.26 in d-q coordinate on a synchronous frame, we have the following 8 equations. Both stator and rotor voltage equations are,

$$V_{qs} = R_s I_{qs} + p\lambda_{qs} + \omega_s \lambda_{ds} \quad (2.21)$$

$$V_{ds} = R_s I_{ds} + p\lambda_{ds} - \omega_s \lambda_{qs} \quad (2.22)$$

$$0 = R_r I_{qr} + p\lambda_{qr} + \omega_r \lambda_{dr} \quad (2.23)$$

$$0 = R_r I_{dr} + p\lambda_{dr} - \omega_r \lambda_{qr} \quad (2.24)$$

¹¹ Where flux linkage variables are defined by

$$\lambda_{qs} = L_s I_{qs} + L_m I_{qr} \quad (2.25)$$

$$\lambda_{ds} = L_s I_{ds} + L_m I_{dr} \quad (2.26)$$

$$\lambda_{qr} = L_m I_{qs} + L_r I_{qr} \quad (2.27)$$

$$\lambda_{dr} = L_m I_{ds} + L_r I_{dr} \quad (2.28)$$

The above equations¹ are very useful in explaining the dynamic structure of the motor and the concept of vector control.

When induction motors are controlled by a vector drive, control computation is often done in the synchronous frame. Since actual stator variables either to be generated or to be measured are all in stationary a-b-c frame, frame transform should be executed in the control. The most popular transform is between stationary a-b-c frame quantities to synchronously rotating d-q quantities. We have

$$S_{qs} = (2/3) \operatorname{Re}\{\exp(-j\theta_a)(S_a + \alpha S_b + \alpha^2 S_c)\} \quad (2.29)$$

$$S_{ds} = -(2/3) \operatorname{Im}\{\exp(-j\theta_a)(S_a + \alpha S_b + \alpha^2 S_c)\} \quad (2.30)$$

Or in a simpler form,

$$\begin{bmatrix} Yq \\ Yd \\ 0 \end{bmatrix} = (2/3) \begin{bmatrix} \cos\theta & \cos(\theta - 2\pi/3) & \cos(\theta + 2\pi/3) \\ \sin\theta & \sin(\theta - 2\pi/3) & \sin(\theta + 2\pi/3) \\ 0.5 & 0.5 & 0.5 \end{bmatrix} \begin{bmatrix} Ya \\ Yb \\ Yc \end{bmatrix} \quad (a)$$

and its inverse transform is given by

$$\begin{bmatrix} Ya \\ Yb \\ Yc \end{bmatrix} = \begin{bmatrix} \cos\theta & \sin\theta & 1 \\ \cos(\theta - 2\pi/3) & \sin(\theta - 2\pi/3) & 1 \\ \cos(\theta + 2\pi/3) & \sin(\theta + 2\pi/3) & 1 \end{bmatrix} \begin{bmatrix} Yq \\ Yd \\ 0 \end{bmatrix} \quad (b)$$

In vector control drives, Eq. (a) is frequently used to convert measured currents and voltages to d-q quantities while Eq. (b) may be used to feed command signals to the amplifier. In many modern drives, Eq. (b) can be accomplished in a slightly different mechanism such as the space vector modulation.

Regardless of reference frame, instantaneous input power can be expressed, in terms of space vectors, by

$$P_i = (3/2) \operatorname{Re}(V_s I_s^*) \quad (2.31)$$

Or in terms of d-q variables as

$$P_i = (3/2)[V_{ds}I_{ds} + V_{qs}I_{qs}] \quad (2.32)$$

The reactive power Q_i can also be defined as

$$Q_i = (3/2)\text{Im}(V_s I_s^*) \quad (2.33)$$

or in terms of d-q variables as

$$Q_i = (3/2)[V_{qs}I_{ds} - V_{ds}I_{qs}] \quad (2.34)$$

This reactive power can be used in some parameter adaptation methods which automatically corrects the rotor time constant parameter (T_r) during steady-state operation.

Now, one simple way of obtaining the output torque is to consider the power associated with speed voltage term on Fig. 2.4 as

$$P_o = (3/2)\omega_0 \text{Im}(\lambda_r I_r^*) \quad (2.35)$$

Since torque is the above power divided by the rotor speed,

$$T_0 = (3/4)P \text{Im}(\lambda_r I_r^*) \quad (2.36)$$

where P is the number of poles. In terms of d-q variables, Eq. 2.36 is

$$T_0 = (3/4)P\{\lambda_{qr}I_{dr} - \lambda_{dr}I_{qr}\} \quad (2.37)$$

Although the torque expression on the above is derived from stationary reference frame, it is true for any other reference frames. Many other forms of torque equations are possible. For example, by substituting flux linkage relation of Eq. 2.17 into Eq.2.36, we have

$$T_0 = (3/4)PL_m \text{Im}(I_s I_r^*) \quad (2.38)$$

$$T_0 = (3/4)P L_m \{ I_{qs} I_{dr} - I_{ds} I_{qr} \} \quad (2.39)$$

Again, by using Eq. 2.16, we can eliminate I_r on Eq. 2.38 to get

$$T_0 = (3/4)P(L_m / L_r) \text{Im}(I_s \lambda_r^*) \quad (2.40)$$

$$T_0 = (3/4)P(L_m / L_r) \{ I_{qs} \lambda_{dr} - I_{ds} \lambda_{qr} \} \quad (2.41)$$

It will be shown later that Eqs. 2.40 – 2.41 are particularly important in vector control because output torque is expressed in terms of stator current and rotor flux linkage.

2.3 SPACE VECTOR MODULATION FOR VOLTAGE SOURCE INVERTERS

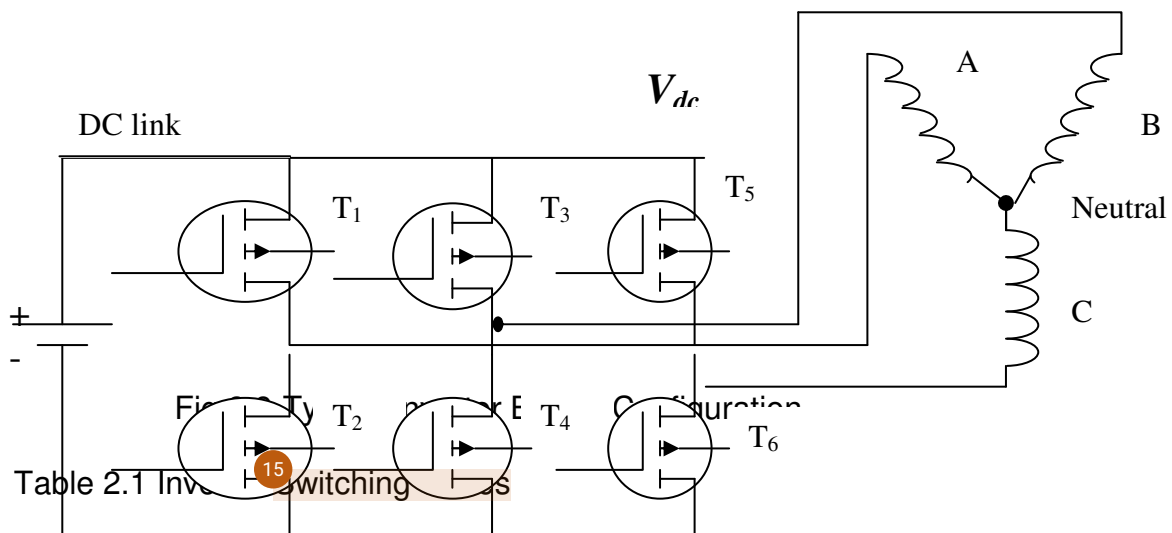
Sinusoidal PWM has been a very popular technique used in AC motor control. This relatively unsophisticated method employs a triangular carrier wave modulated by a sine wave and the points of intersection determine the switching points of the power devices in the inverter. However, this method is unable to make full use of the inverter's supply voltage and the asymmetrical nature of the PWM switching characteristics produces relatively high harmonic distortion in the supply.

Space Vector PWM (SVPWM) is a more sophisticated technique for generating a fundamental sine wave that provides a higher voltage to the motor and lower total harmonic distortion, it is also compatible for use in vector control (Field orientation) of AC motors. This thesis work describes the theory of SVPWM and applies it to a practical example using a TMS320LF2407 DSP kit.

2.4 VOLTAGE SOURCE INVERTER

The topology of a three-leg voltage source inverter is shown in Fig. 2.8. Because of the constraint that the input lines must never be shorted and the output current must always be continuous a voltage source inverter can assume only eight distinct topologies [1]. These topologies are shown on Fig. 2.9. Six out of these eight topologies produce a nonzero output voltage and are known as non-zero switching states and the remaining two topologies produce zero output voltage and are known as zero switching states [9].

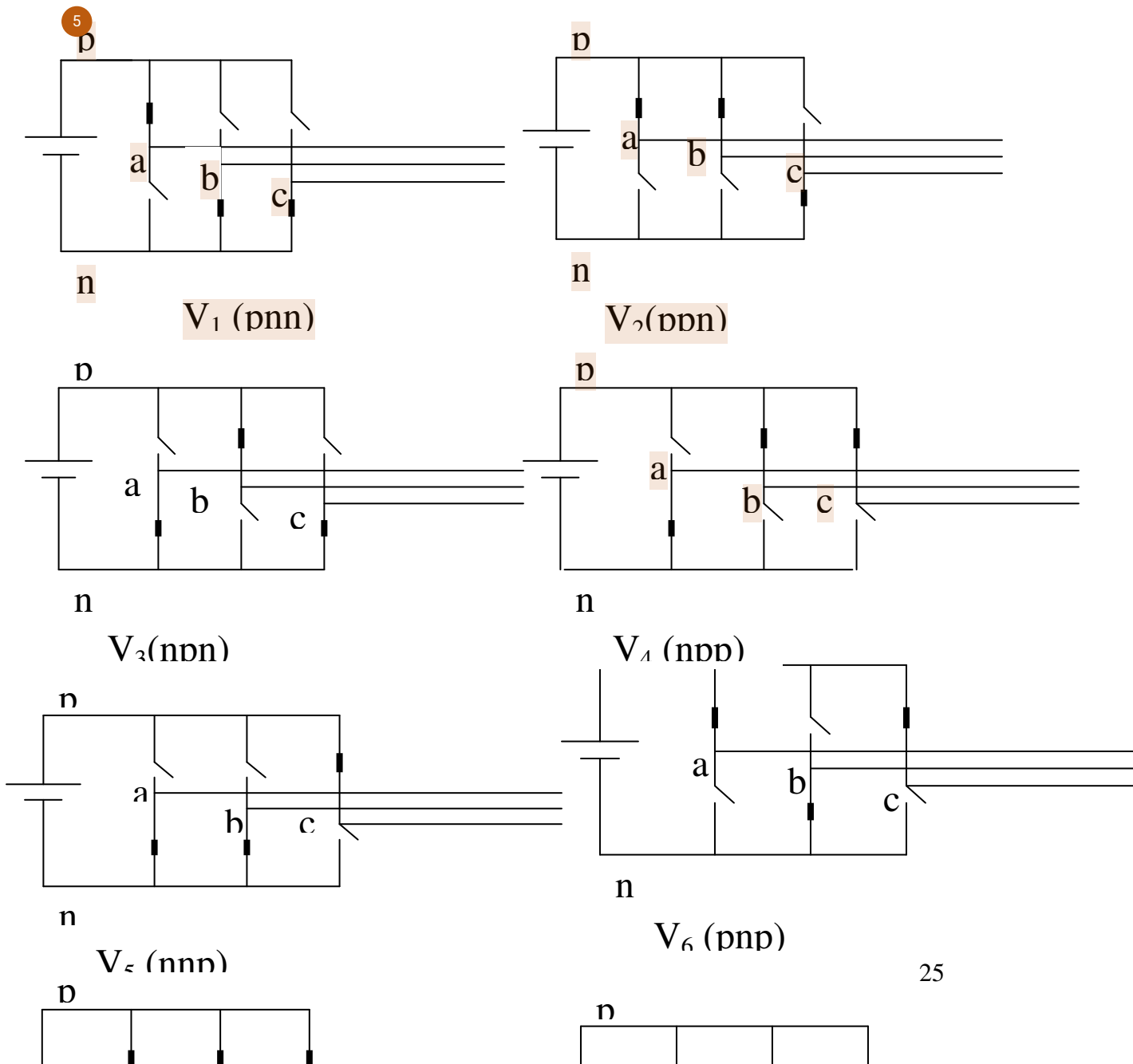
7 Consider first the 3-phase inverter shown below (Figure 2-8). The 7 six-transistor combination in the inverter has eight permissible switching states. Table 2-1 summarizes these states along with the corresponding line to neutral voltage applied to the motor.



State	On Device	V_{an}	V_{bn}	V_{cn}	Space Voltage Vector
0	T_2, T_4, T_6	0	0	0	$V_0(000)$
1	T_1, T_4, T_6	$(2/3)V_{dc}$	$(-1/3)V_{dc}$	$(-1/3)V_{dc}$	$V_1(100)$
2	T_1, T_3, T_6	$(1/3)V_{dc}$	$(1/3)V_{dc}$	$(-2/3)V_{dc}$	$V_2(110)$
3	T_3, T_2, T_6	$(-1/3)V_{dc}$	$(2/3)V_{dc}$	$(-1/3)V_{dc}$	$V_3(010)$
4	T_2, T_3, T_5	$(-2/3)V_{dc}$	$(1/3)V_{dc}$	$(1/3)V_{dc}$	$V_4(011)$
5	T_2, T_4, T_5	$(-1/3)V_{dc}$	$(-1/3)V_{dc}$	$(2/3)V_{dc}$	$V_5(001)$

6	T_1, T_4, T_5	$(1/3)V_{dc}$	$(-2/3)V_{dc}$	$(1/3)V_{dc}$	$V_6(101)$
7	T_1, T_3, T_5	0	0	0	$V_7(111)$

The inverter has six states when a voltage is applied to the motor and two states (0 and 7) when the motor is shorted through the upper or lower transistors resulting in zero volts being applied to the motor. ⁷ It is useful to express these states as vectors. V_0-7 expresses the three voltages V_{an} , V_{bn} and V_{cn} , that are spatially separated 120° apart, as a space vector (V) for each of the switching states 0-7. The six vectors including the zero voltage vectors can be expressed geometrically as shown in Figure 2-11.



6 Fig 2.9 Eight switching state topologies of a voltage source inverter.

2.5 VOLTAGE SPACE VECTORS

Space vector modulation (SVM) for three-leg VSI is based on the representation of the three phase quantities as vectors in a two-dimensional (α, β) plane. Considering topology 1 of Fig. 2.9, which is repeated in Fig. 2.10 (a) we see that the line voltages V_{ab} , V_{bc} , and V_{ca} are given by

$$\begin{aligned} V_{ab} &= V_g \\ V_{bc} &= 0 \\ V_{ca} &= -V_g \end{aligned} \quad \dots(2.42)$$

This can be represented in the α, β plane as shown in Fig. 2.10(b), where voltages V_{ab} , V_{bc} , and V_{ca} are three line voltage vectors displaced 120° in space. The effective voltage vector generated by this topology is represented as $V_1(pnn)$ in Fig. 2.10(b). Here the notation 'pnn' refers to the three legs/phases a,b,c being either connected to the positive dc rail (p) or to the negative dc rail (n). Thus 'pnn' corresponds to 'phase a' being connected to the positive dc rail and phases b and c being connected to the negative dc rail.[10]

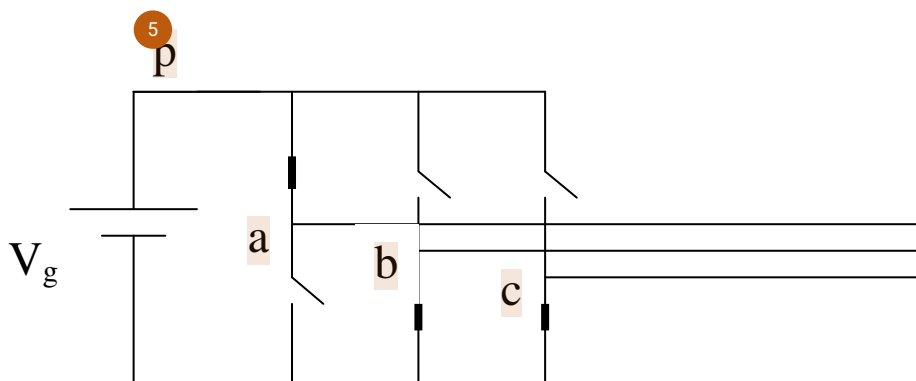
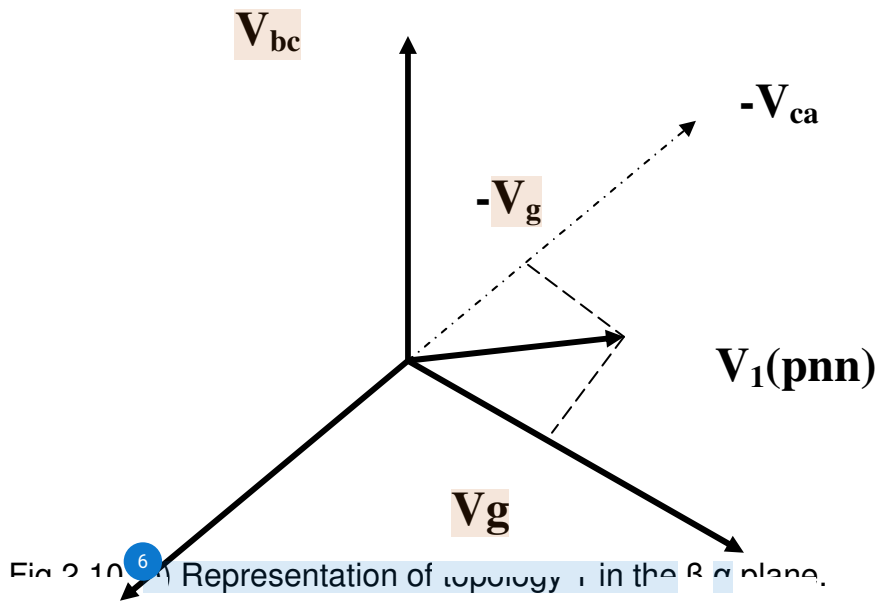


Fig 2.10(a) Topology 1-V1 (pnn) of a voltage source inverter.



Proceeding in a similar manner, the six non-zero voltage vectors V_{ab} (V1 - V6) can be shown to assume the positions shown in Fig. 2.11. The tips of these vectors form a regular hexagon (dotted line in Fig. 2.11). We define the area enclosed by two adjacent vectors, within the hexagon, as a sector. Thus there are six sectors numbered 1 - 6 in Fig. 2.11.

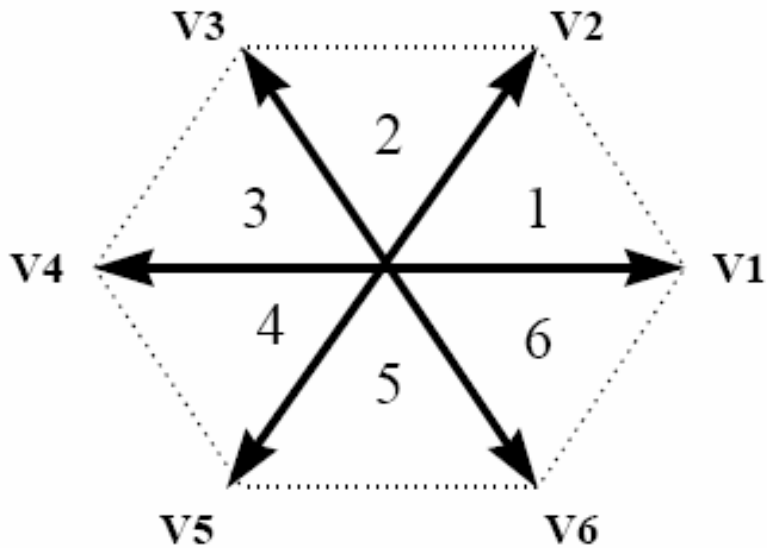


Fig. 2.11 Non-zero voltage vectors in the α, β plane.

Considering the last two topologies of Fig. 2.9 which are repeated in Fig. 2.12(a) for the sake of convenience we see that the output line voltages generated by this topology are given by

$$V_{ab}=0$$

$$V_{bc}=0$$

.....(2.43)

$$V_{ca}=0$$

These are represented as vectors which have zero magnitude and hence are referred to as zero-switching state vectors or zero voltage vectors. They assume the position at origin in the α, β , plane as shown in Fig. 2.12(b). The vectors V1-V8 are called the switching state vectors (SSVs).

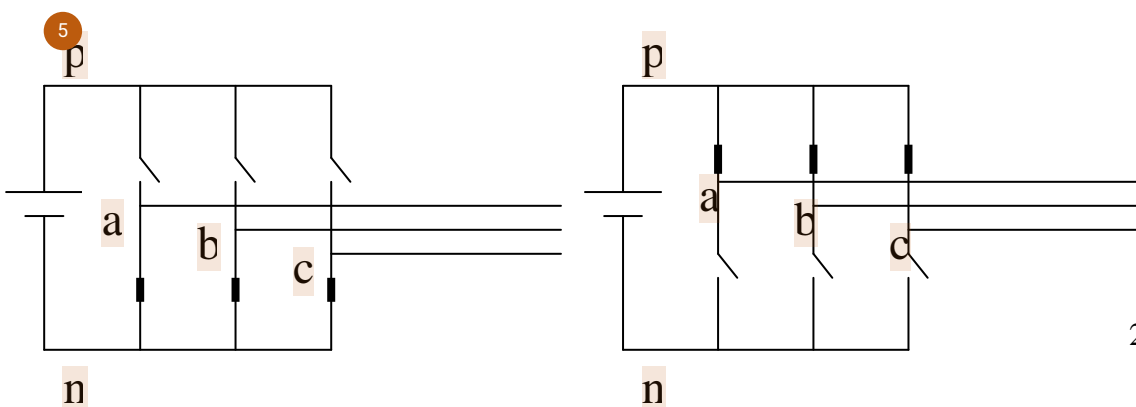
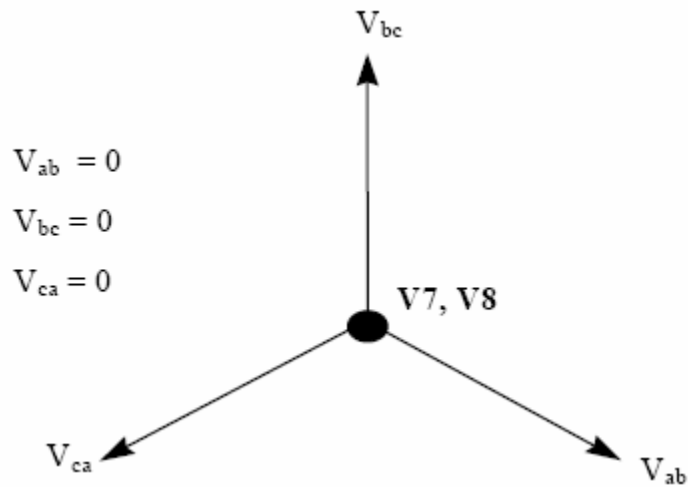


Fig. 2.12(a). Zero output voltage topologies.



5 Fig. 2.12(b) Representation of the zero voltage vectors in the α, β plane.

2.6 SPACE VECTOR MODULATION

The desired three phase voltages at the output of the inverter could be represented by an equivalent vector V rotating in the counter clock wise direction as shown in Fig. 2.13(a). The magnitude of this vector is related to the magnitude of the output voltage (Fig. 2.13(b)) and the time this vector takes to complete one

revolution is the same as the fundamental time period of the output voltage.

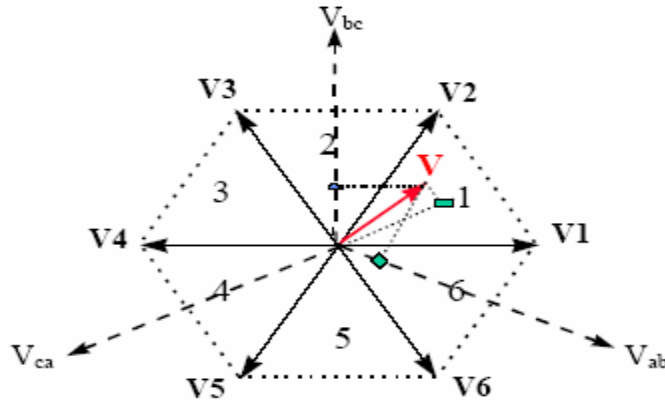


Fig. 2.13(a). Output voltage vector in the α, β plane

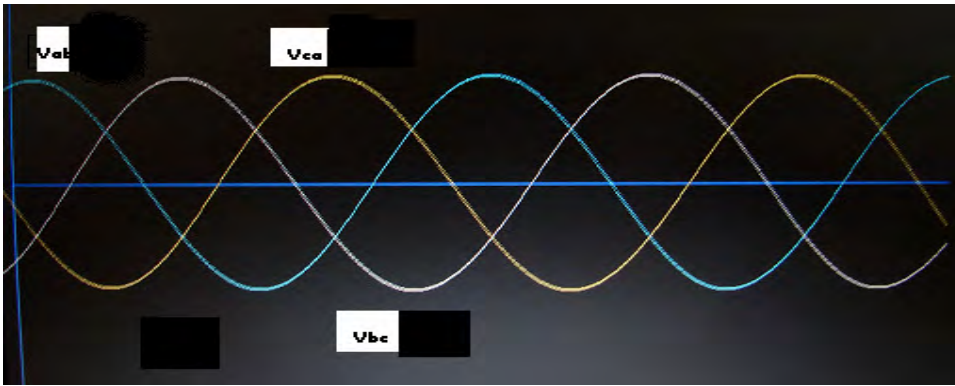


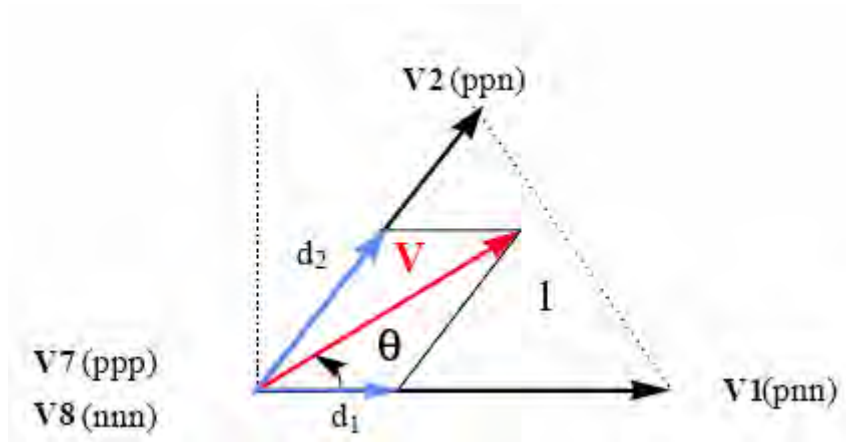
Fig. 2.13(b). Output line voltages in time domain.

Let us consider the situation when the desired line-to-line output voltage vector V is in sector 1 as shown in Fig. 2.14. This vector could be synthesized by the pulse-width-modulation (PWM) of the two adjacent SSV's V_1 (pnn) and V_2 (ppn), the duty cycle of each being d_1 and d_2 , respectively, and the zero vector (V_7 (nnn) / V_8 (ppp)) of duty cycle d_0 :

$$d_1 V_1 + d_2 V_2 = V = m V_g e^{j\theta} \quad \dots (2.44)$$

$$d_1 + d_2 + d_0 = 1 \quad \dots (2.45)$$

where, $0 < m < 0.866$, is the modulation index. This would correspond to a maximum line-to-line voltage of $1.0V_g$, which is 15% more than conventional sinusoidal PWM.



5 Fig. 2.14. Synthesis of the required output voltage vector in sector 1.

All SVM schemes and most of the other PWM algorithms, use (2.44), (2.45) for the output voltage synthesis. The modulation algorithms that use non-adjacent SSV's have been shown to produce higher THD and/or switching losses and are not analyzed here, although some of them, e.g. hysteresis, can be very simple to implement and can provide faster transient response. The duty cycles d_1 , d_2 , and d_0 , are uniquely determined from Fig. 2.14, and (2.44), (2.45), the only difference between PWM schemes that use adjacent vectors is the choice of the zero vector(s) and the sequence in which the vectors are applied within the switching cycle.

2.7 Clarke and Parke transformation (Rotating transformation)

14 Vector control of an AC motor is analogous to the control of separately excited DC motor. In a DC motor the field flux ϕ_f produced by the field current I_f is perpendicular to the armature flux ϕ_a produced by armature current I_a . These fields are decoupled and stationary with respect to each other. Therefore, when

the armature current is controlled to control the torque the field flux remains unaffected enabling a fast transient response.

Vector control seeks to recreate these orthogonal components in the AC machine in order to control the torque producing current separately from the magnetic flux producing current so as to achieve the responsiveness of the DC machine. ⁴⁴ In order to create the perpendicular components of the stator current of an Induction Motor in this thesis work, we have to have the concept of rotating transformation. In this subsection The (a,b)->(α-β) projection (Clarke transformation) and The (α,β)->(d,q) projection (Park transformation) will be discussed. [5]

The Clarke and Park ¹⁷ transformation is a transformation of coordinates from the three-phase stationary coordinate system to the d-q rotating coordinate system. This transformation is made in two steps:

A transformation from the three-phase stationary coordinate system to the two phase, so-called α-β, stationary coordinate system is called Clarke Transformation.

In figure 2.15 assume that there is a balanced three phase stator currents (acb phase sequence)

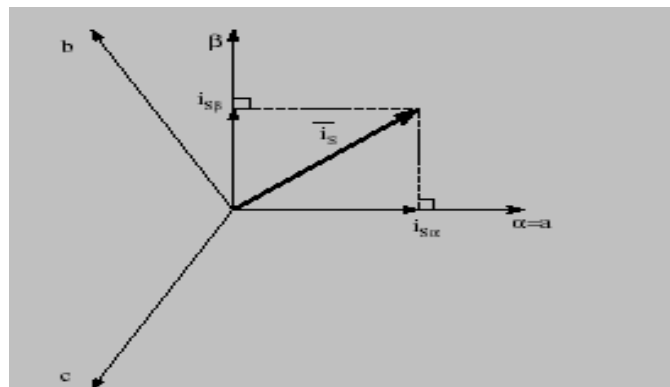


Figure 2.15 (a,b,c)->(α,β) Projection (Clarke Transformation)

The projection of the stator current/voltage vector from three phase into two phase in this fixed frame gives:

$$\begin{aligned} x_{s\alpha} &= x_a \\ x_{s\beta} &= \frac{1}{\sqrt{3}}x_a + \frac{2}{\sqrt{3}}x_b \\ x_a + x_b + x_c &= 0 \end{aligned} \quad (2.46)$$

In matrix form of a three phase to two phase transformation of the stator current/voltage ((a,b,c) to (α,β))

$$\begin{bmatrix} x_{s\alpha} \\ x_{s\beta} \end{bmatrix} = \frac{2}{3} \begin{bmatrix} 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} \quad (2.47)$$

Where:

x=is stator voltage or current

A transformation from the αβ stationary coordinate system to the dq rotating coordinate system is called Park Transformation.[13]

Here θ_e is the angle between the d-axis and the stationary α-axis.

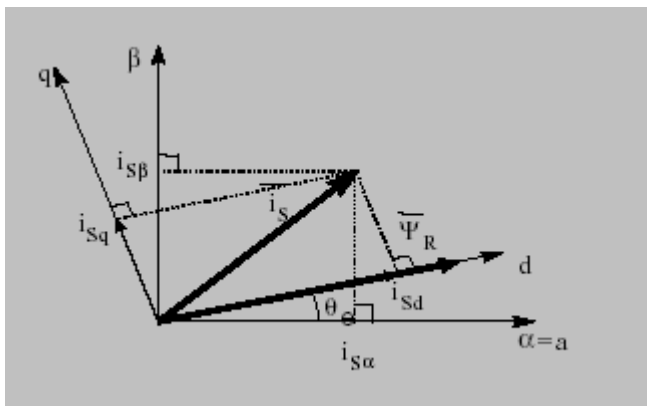


Figure 2.16- (a,b)->(d,q) Projection (Park Transformation)

The equations corresponding to this transformation are given by

$$\begin{aligned} i_{sd} &= i_{s\alpha} \cos(\theta_e) + i_{s\beta} \sin(\theta_e) \\ i_{sq} &= -i_{s\alpha} \sin(\theta_e) + i_{s\beta} \cos(\theta_e) \end{aligned} \quad (2.48)$$

From Park transformation i_{sd} and i_{sq} are obtained directly from i_a , i_b and i_c . Here is the matrix equation corresponding to this transformation.

$$\begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} = \begin{bmatrix} \cos \theta_e & \cos(\theta_e - 2\pi/3) & \cos(\theta_e + 2\pi/3) \\ -\sin \theta_e & -\sin(\theta_e - 2\pi/3) & -\sin(\theta_e + 2\pi/3) \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \quad (2.49)$$

Inverse park transformation has the following equations

$$\begin{aligned} i_a &= i_d \cos \theta_e - i_q \sin \theta_e \\ i_b &= i_d \cos(\theta_e - 2\pi/3) - i_q \sin(\theta_e - 2\pi/3) \\ i_c &= i_d \cos(\theta_e - 4\pi/3) - i_q \sin(\theta_e - 4\pi/3) \end{aligned} \quad (2.50)$$

2.8 Controllers

The controller compares a measured value from a process with a reference setpoint value. The difference (or "error" signal) is then used to calculate a new value for a manipulatable input to the process that brings the process-measured value back to its desired set point. It can be shown mathematically that a PI loop will produce accurate, stable control in cases where a simple proportional control would either have a steady-state error or would cause the process to oscillate. PI controllers do not require advanced mathematics to design and can be easily adjusted (or "tuned") to the desired application.

PID controllers are normally implemented with the K_p gain applied to the I_{contrib} term as well in the following form[3][10];

$$\text{output} = K_p (e(t) + K_i \int_{-\infty}^t e(t) dt + K_d de(t) / d(t)) \quad (2.51)$$

Often, one deals with discrete time intervals instead of the continuous.

Here, the first term is integral, the second proportional, and the third derivative. Note that in this form, K_i must be identically 1, otherwise the controller will not even come close to converging to the set-point. This isn't quite the same integral as in the continuous form, but it's analogous.

"Tuning" a control loop is the adjustment of its control parameters (gain/proportional band, integral gain/reset, derivative gain/rate) to the optimum values for the desired control response. The optimum behavior on a process change or set-point change varies depending on the application. Some processes must not allow an overshoot of the process variable from the set-point. Other processes must minimize the energy expended in reaching a new set-point. Generally stability of response is required and the process must not oscillate for any combination of process conditions and set-points. Tuning of loops is made more complicated by the response time of the process; it may take minutes or hours for a set-point change to produce a stable effect. Some processes have a degree of non-linearity and so parameters that work well at full-load conditions don't work when the process is starting up from no-load. This section describes some traditional manual methods for loop tuning.

There are several methods for tuning a PID loop. The choice of method will depend largely on whether or not the loop can be taken "offline" for tuning, and the response speed of the system. If the system can be taken offline, the best tuning method often involves subjecting the system to a step change in input,

measuring the output as a function of time, and using this response to determine the control parameters.

If the system must remain online, one tuning method is to first set the I and D values to zero. Increase the P until the output of the loop oscillates. Then increase I until oscillation stops. Finally, increase D until the loop is acceptably quick to reach its reference. A fast PID loop tuning usually overshoots slightly to reach the set-point more quickly; however, some systems cannot accept overshoot.

Effects of increasing parameters				
Parameter	Rise Time	Overshoot	Settling Time	S.S. Error
P	Decrease	Increase	Small Change	Decrease
I	Decrease	Increase	Increase	Eliminate
D	Small Change	Decrease	Decrease	Small Change

PI controllers, ¹³ A control loop consists of three parts:

- ✓ Measurement by a sensor connected to the process (or the "plant"),
- ✓ Decision in a controller element,
- ✓ Action through an output device ("actuator") such as a control valve.

As the controller reads a sensor, it subtracts this measurement from the "set-point" to determine the "error". It then uses the error to calculate a correction to the process's input variable (the "action") so that this correction will remove the error from the process's output measurement.

In a PI loop, correction is calculated from the error in two ways: cancel out the current error directly (Proportional), the amount of time the error has continued uncorrected (Integral).

24 More technically, a PI loop can be characterized as a filter applied to a complex frequency-domain system. This is useful in order to calculate whether it will actually reach a stable value. If the values are chosen incorrectly, the controlled process input can oscillate, and the process output may never stay at the set point.

Traditionally

$$output = P_{contrib} + I_{contrib} \quad (2.52)$$

13 where $P_{contrib}$, $I_{contrib}$, and $I_{contrib}$ are the feedback contributions from the PI controller, defined below:

$$P_{contrib} = K_p e(t) \quad (2.53)$$

$$I_{contrib} = K_i \int_{-\infty}^t e(t') dt' \quad (2.54)$$

4 Where K_p , and K_i are constants that are used to tune the PID control loop:

- ✓ K_p : Proportional Gain - Larger K_p typically means faster response since the larger the error, the larger the feedback to compensate.
- ✓ K_i : Integral Gain - Larger K_i implies steady state errors are eliminated quicker. The tradeoff is larger overshoot: any negative error integrated during transient response must be integrated away by positive error before we reach steady state.

Chapter Three

Experimentation and simulation results

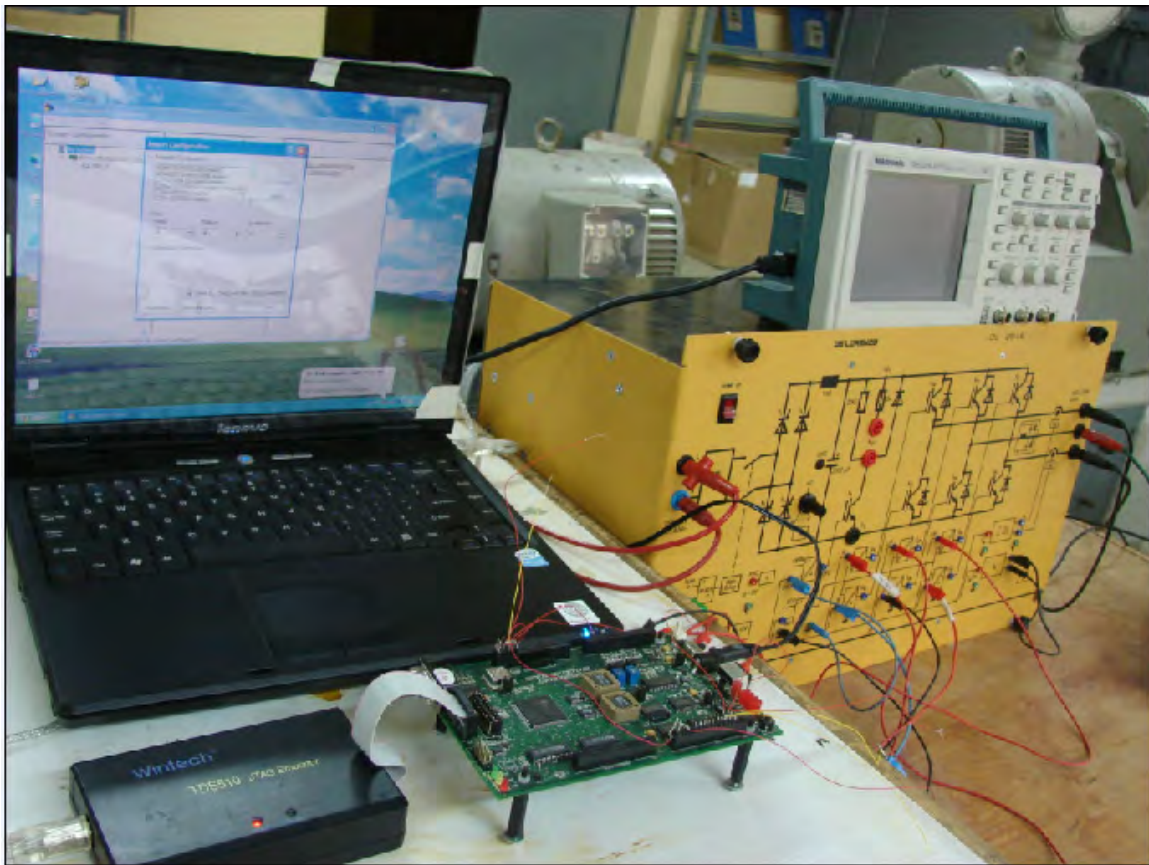
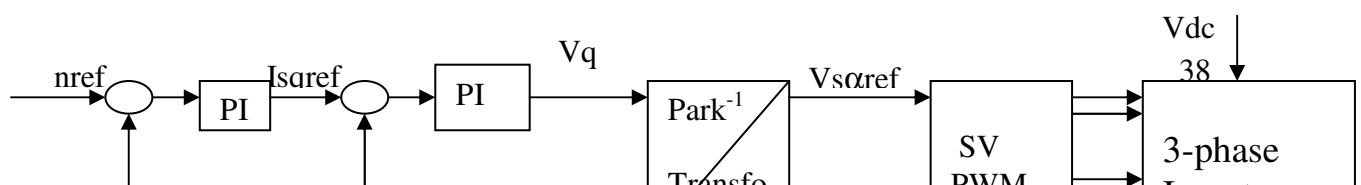


Fig 3.1 Experimental Setup for an open loop AC induction motor control system

3.1 Block Diagram Representation:

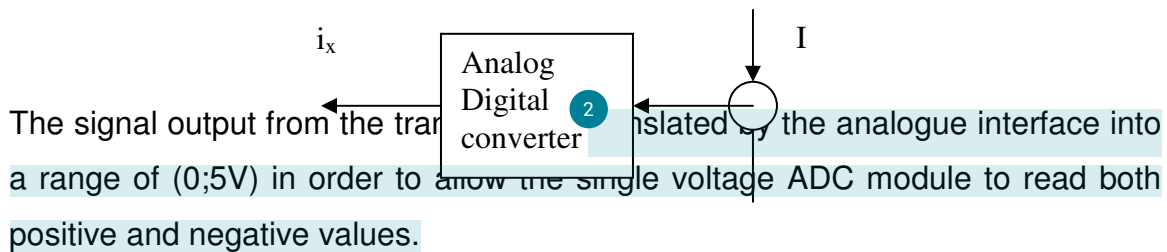


Vd

Fig. 3.2 Block diagram representation for Vector control of IM

✧ **Current Sensing:**

Since two phase currents which feed the Clarke transformation are needed, current Voltage transducers are used in the Inverter to sense these two currents.



The two phase currents are sampled simultaneously by means of the DSP controller by using one channel of each ADC module per current. In this case P2 Analog pins ADCIN6 and ADCIN7 can be used to sample the phase currents.

✧ **The PI Regulator:**

The PI regulators are implemented with output saturation and with integral component correction.

The constants K_{pi} , K_i , K_{cor} (proportional, integral and integral correction components) are selected depending on the sampling period and on the motor parameters.

$$K_{cor} = K_i / K_{pi}$$

✧ **Clarke and park transformation:**

Transformation from three phase into two phase vectors is called rotational transformation. Clarke and Parke transformation are used to transform the three phase current outputs from the inverter into two rotating phases (d and q) which are perpendicular to each other.

The (a,b) \longrightarrow (α ; β) projection (Clarke transformation)

$$i_{s\alpha} = i_a$$

$$i_{s\beta} = (2i_b + i_a) / \sqrt{3}$$

In this case balanced three phase system is shown below when transformed into (α ; β) orthogonal frame.

The (α ; β) \longrightarrow (d,q) projection (Park transformation)

$$i_{sd} = i_{s\alpha} * \cos(\theta) + i_{s\beta} * \sin(\theta)$$

$$i_{sq} = -i_{s\alpha} * \sin(\theta) + i_{s\beta} * \cos(\theta)$$

Where θ represents the rotor flux position calculated by the current model.

✧ **The Current Model:**

This module takes as input i_{sd} , i_{sq} plus the rotor electrical speed. The current module implements the following two equations of the motor in d,q reference frame:

$$i_{sd} = T_R (di_{mR} / dt) + i_{mR}$$

$$f_s = (1/\omega_b) d\theta / dt = n + (i_{qs} / T_R i_{mr} \omega_b)$$

where θ is the rotor flux position, i_{mR} magnetizing current, and where

$$T_R = L_R / R_R$$

is the rotor time constant. Knowledge of this constant is critical to the correct functioning of the current model as it is this system that outputs the rotor flux speed that will be integrated to get the rotor flux position. Assuming that $i_{qsk+1} = i_{qsk}$

The above equation can be discretized as follows:

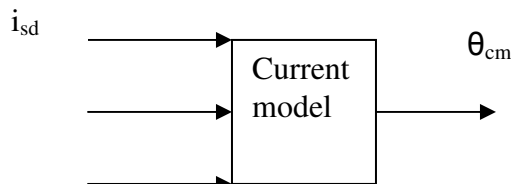
$$i_{mrk+1} = i_{mrk} + T / T_R * (i_{dsk} - i_{mRk})$$

$$f_{sk+1} = n_{k+1} + (1/T_R \omega_b) * (i_{qsk} / i_{mRk+1})$$

Once the rotor flux speed (f_s) has been calculated, the necessary rotor flux position (θ_{cm}) is computed by the integration formula:

$$\theta_{cmk+1} = \theta_{cmk} + \omega_b f_{sk} T$$

where $\omega_b f_s T$ is the angle variation with in one sample period. At nominal operation ($f_s=1$, mechanical speed is 1500rpm) $\omega_b f_s T = 0.031415 \text{ rad}$. In one mechanical revolution performed at nominal speed there are $2\pi / 0.031415 = 200$ increments of the rotor flux position.



✧ **Space Vector Modulation:**

The SVM needs the reference voltages V_{sref} , $V_{s\beta ref}$ as input, the DC bus voltage as parameter and gives the three PWM patterns as output.

3.2 Software-Determined Switching Pattern

Figure 3.2 below shows the waveform for each sector of a symmetric switching scheme.

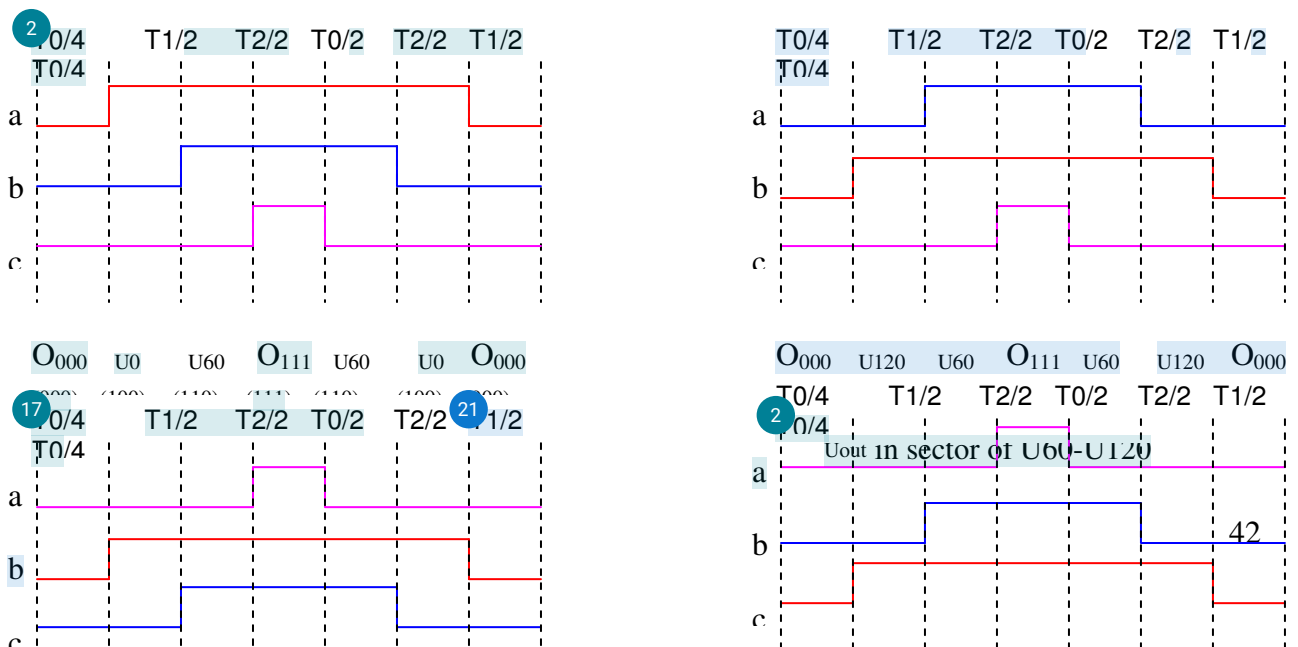
This scheme can easily be implemented with the TMS320C24x/F24x using software determined switching order for the three PWM channels. Figure 5 is another illustration of the switching scheme, where the arrows indicate for each sector the order of the first and second basic space vectors. This switching scheme can be represented by $(O_{000}, U_x, U_{x \pm 60}, O_{111}, O_{111}, U_{x \pm 60}, U_x, O_{000})$, where x can be 0, 120 and 240. It has the following properties:

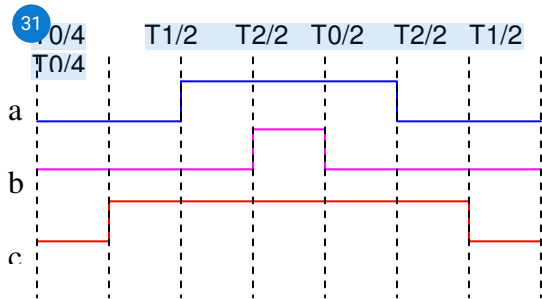
Each PWM channel switches twice per every PWM period except when the duty cycle is 0% or 100%.

There is a fixed switching order among the three PWM channels for each sector.

Every PWM period starts and ends with O000.

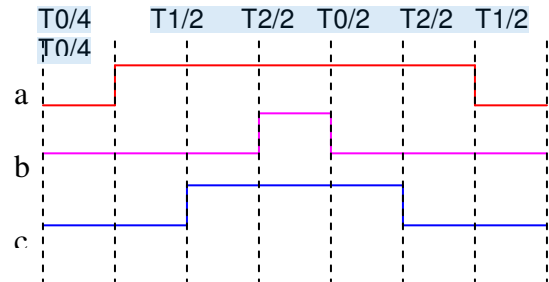
The amount of O000 inserted is the same as that of O111 in each PWM period.





O_{000} U_{240} U_{300} O_{111} U_{300} U_{240} O_{000}
 (000) (001) (101) (111) (101) (001) (000)

Uout in sector of U240-U300

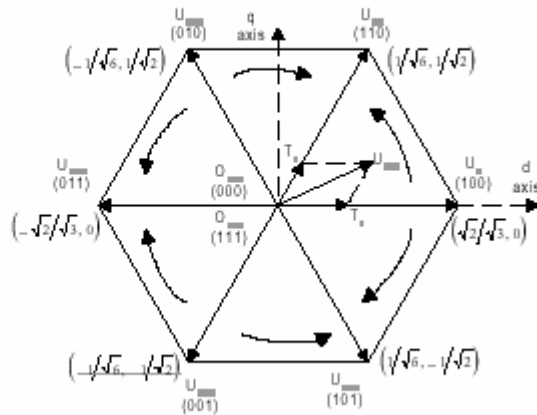


O_{000} U_0 U_{300} O_{111} U_{300} U_0 O_{000}
 (000) (100) (101) (111) (101) (100) (000)

Uout in sector of U0-U300

inverted SV PWM

software Determined SV PWM Pattern



Hence:

The ² space vector modulation is a highly efficient way to generate the six pulsed signals necessary at the power stage. SVM needs the reference voltages V_{sref} , $V_{s\beta ref}$ as input, the DC bus voltage as parameter and gives the three PWM patterns as output.

In my algorithm I have used the following ways to calculate the duty cycles:

$$V_{ref1} = V_{s\beta ref}$$

$$V_{ref2} = 1/2(\sqrt{3}V_{s\alpha ref} - V_{s\beta ref})$$

$$V_{ref3} = 1/2(-\sqrt{3}V_{s\alpha ref} - V_{s\beta ref})$$

$$X = \sqrt{3}V_{dcinv} V_{s\beta ref}$$

$$Y = \sqrt{3}/2V_{dcinv} V_{s\beta ref} + 3/2V_{dcinv} V_{s\alpha ref}$$

$$Z = \sqrt{3}/2V_{dcinv} V_{s\beta ref} - 3/2V_{dcinv} V_{s\alpha ref}$$

Where

$$V_{dcinv} = T_{PWM} / V_{dc}$$

1) Then the first step to determine is in which sector the voltage vector defined by $V_{s\alpha ref}$, $V_{s\beta ref}$ is found. The following few code lines give the sector as output:

Sector Determination

If $V_{ref1} > 0$ Then A=1; Else A=0

If $V_{ref2} > 0$ Then B=1, Else B=0

If $V_{ref3} > 0$ Then C=1, Else C=0

Sector= A+2B+4C

2) The second step to perform is to calculate and saturate the duration of the two sector boundary vectors application as shown below:

CASE sector OF

1 $t_1=Z$ $t_2=Y$

2 $t_1=Y$ $t_2=-X$

3 $t_1=-Z$ $t_2=X$

4 $t_1=-X$ $t_2=Z$

5 $t_1=X$ $t_2=-Y$

6 $t_1=-Y$ $t_2=-Z$

end times calculation

Saturations

If $(t_1 + t_2) > PWMPRD$ Then

$$t_{1sat} = t_1 (PWMPRD / (t_1 + t_2))$$

$$t_{2sat} = t_2 (PWMPRD / (t_1 + t_2))$$

3) The third step is to compute the three necessary duty cycles. This is shown below

$$t_{aon} = (PWMPRD - t_1 - t_2) / 2$$

$$t_{bon} = t_{aon} + t_1$$

$$t_{con} = t_{bon} + t_2$$

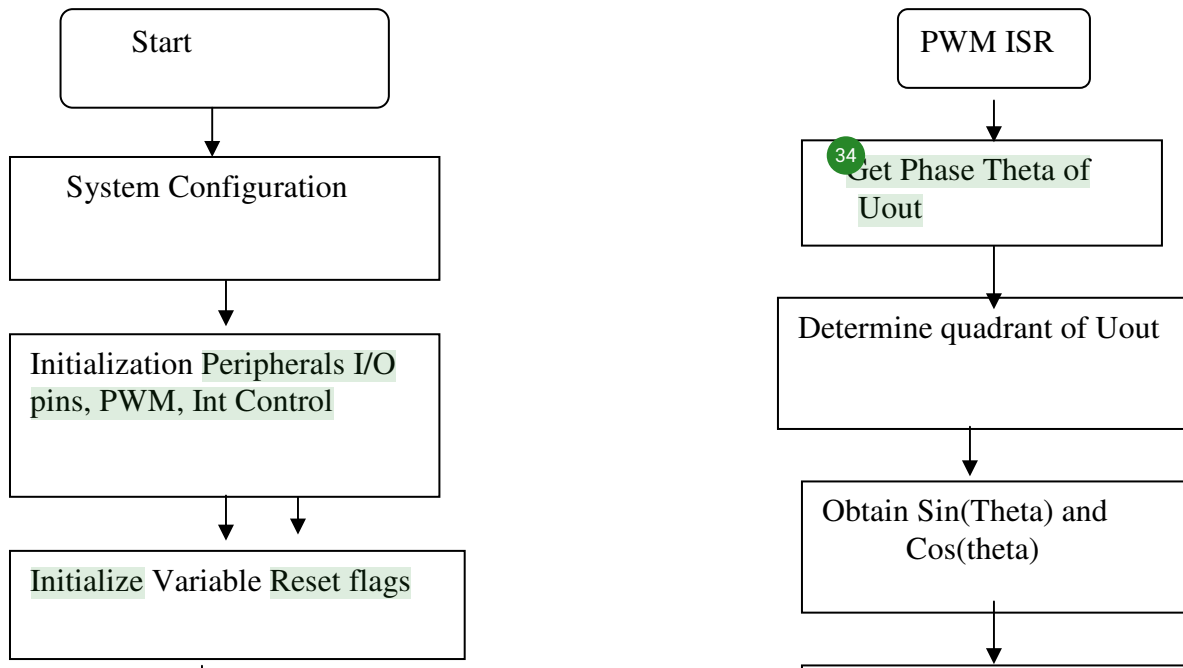
4) ²⁰ The last step is to assign the right duty cycle (t_{xon}) to the right motor phase (in my programming word to the right CMPRx) according to the sector

	1	2	3	²³ +	5	6
--	---	---	---	--------------------	---	---

Sector						
CMPR4	T_{bon}	T_{aon}	T_{aon}	T_{con}	t_{con}	t_{bon}
CMPR5	T_{aon}	T_{con}	T_{bon}	T_{bon}	t_{aon}	t_{con}
CMPR6	T_{con}	T_{bon}	T_{con}	T_{aon}	t_{bon}	t_{aon}

3.3 Software Algorithm

Fig3.5 a) SV PWM technique open loop three-phase AC IM Control



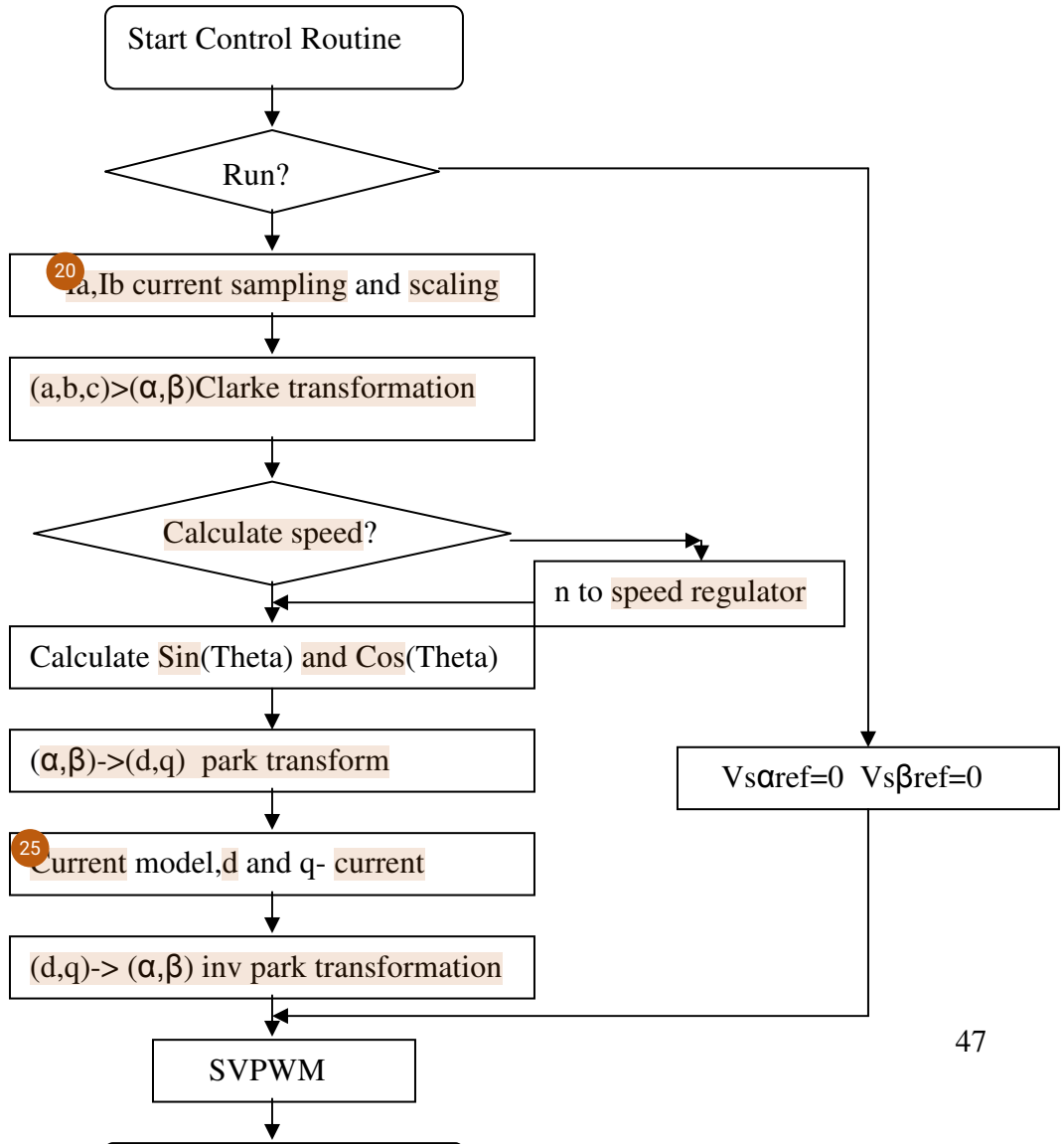


Fig3.5 b)DSP based closed loop Vector control of IM implementation flow chart

By using the code composer studio working environment it generates SVPWM with a frequency of 20KHz as shown below. These outputs are given to an frequency converter DL 2646 with technical specification of output voltage 3x0...230v; output current 3x8A max.

supply Voltages: power circuit 1x255Vmax, 50/60 Hz; control circuit single phase main.

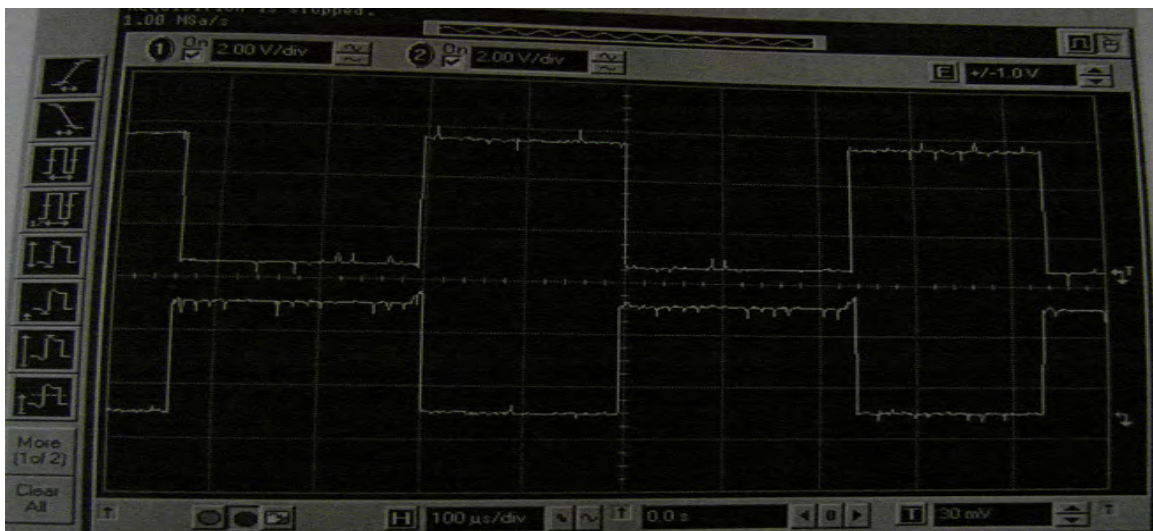


FIG3.6. 20KHz PWM wave output from the DSP board

The six drivers with galvanic isolation for controlling the power transistors of the invertors. Each of the six transistors can be switched on and off with the appropriate control signal L_{ij} ; an interlocking logic element prevents the transistors in a branch pair from both simultaneously switching through in order to avoid short-circuiting of the link capacitor.

The switching status G_{ij} of each individual transistor can be recognized at a relevant leds while at the sockets is present a high digital signal when the IGBT is on.

The drive signal L_{ij} is supplied by the PWM generated by the DSP control unit connecting a bridging plug L_{ij} -DGND.

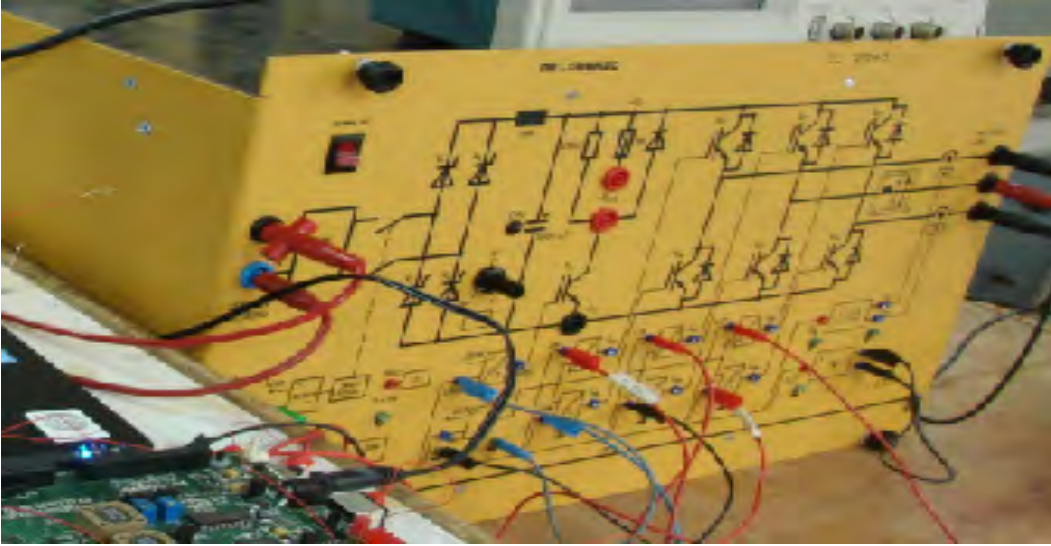


Fig3.7. DL2646 Frequency converter

The output from the frequency converters at the terminals U, V, W of the three phase motor 50HZ frequency as shown below. The output line voltage amount is to $3 \times 230V$ so a motor with rated voltage $230V_{\text{delta}}/400V_{\text{star}}$ must be connected in delta configuration while a $400V_{\text{delta}}/690V_{\text{star}}$ motor can never achieve the rated voltage. The three phase voltage outputs are as shown below.

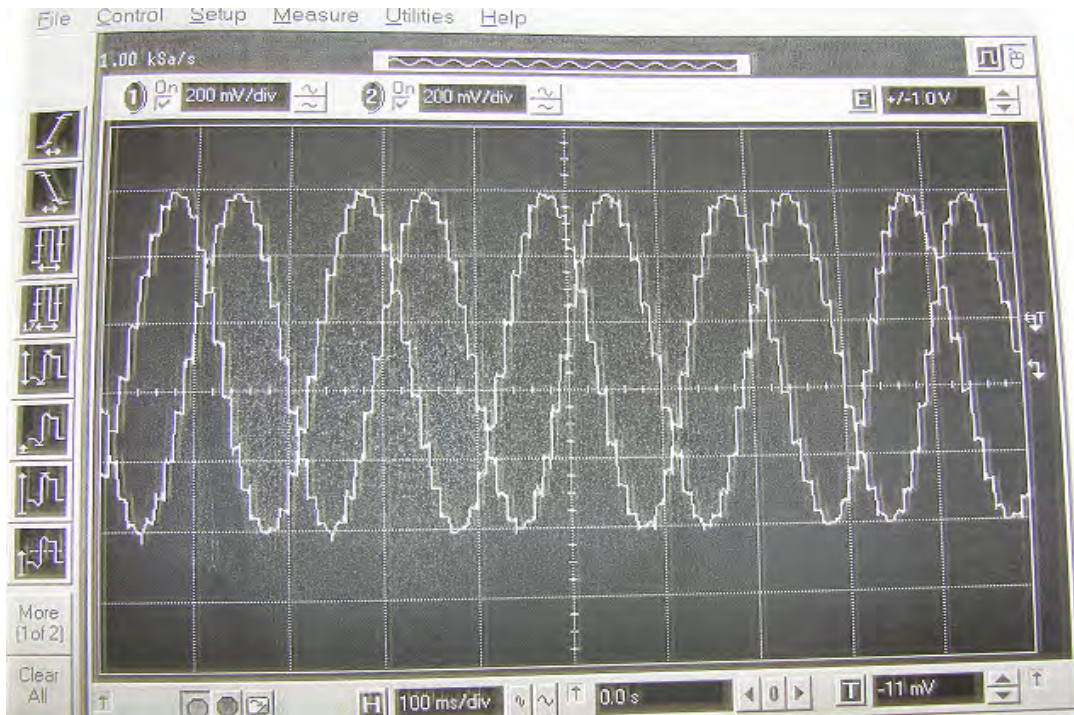


FIG 3.8. 50HZ Three phase sine wave generated from the frequency converter

Since the induction motor I have got is 0.8kw which my inverter can't drive up to synchronous speed.

3.4 Simulation result using MATLAB:

A three-phase induction motor rated at 50hp, 460V, 4pole, 60Hz is modeled in the common stator reference frame. The speed loop utilizes a PI controller to produce the quadrature-axis currents i_q^* which serves as the torque command. The derived direct-axis current i_d^* serves as the rotor flux command. These currents are then transformed via the `<dq2abc>` block into the current references

i_a^* , i_b^* , and i_c^* fed into the current regulator. The current-controlled PWM inverter is made up of three phase inverter. Rotor flux field orientation is obtained in the <theta calculator> block [12].

The machine is initially operating in steady state at no load ($T_l=0$) at the speed $w_{ref}=120\text{rad/s}$. Examine the transient response due to step changes in the command speed w_{ref} and load torque T_l . Repeat the simulation for a so-called detuned case when the estimated value of the rotor time constant T_r appearing in the <theta calculator> differs from the actual value used for the motor parameters.

Here under we can see the simulation result for w_{ref} torque response is not as fast as the vector control response. To compare with the dynamic response for vector control of induction motor you can see Fig.3.9 and Fig.3.10.

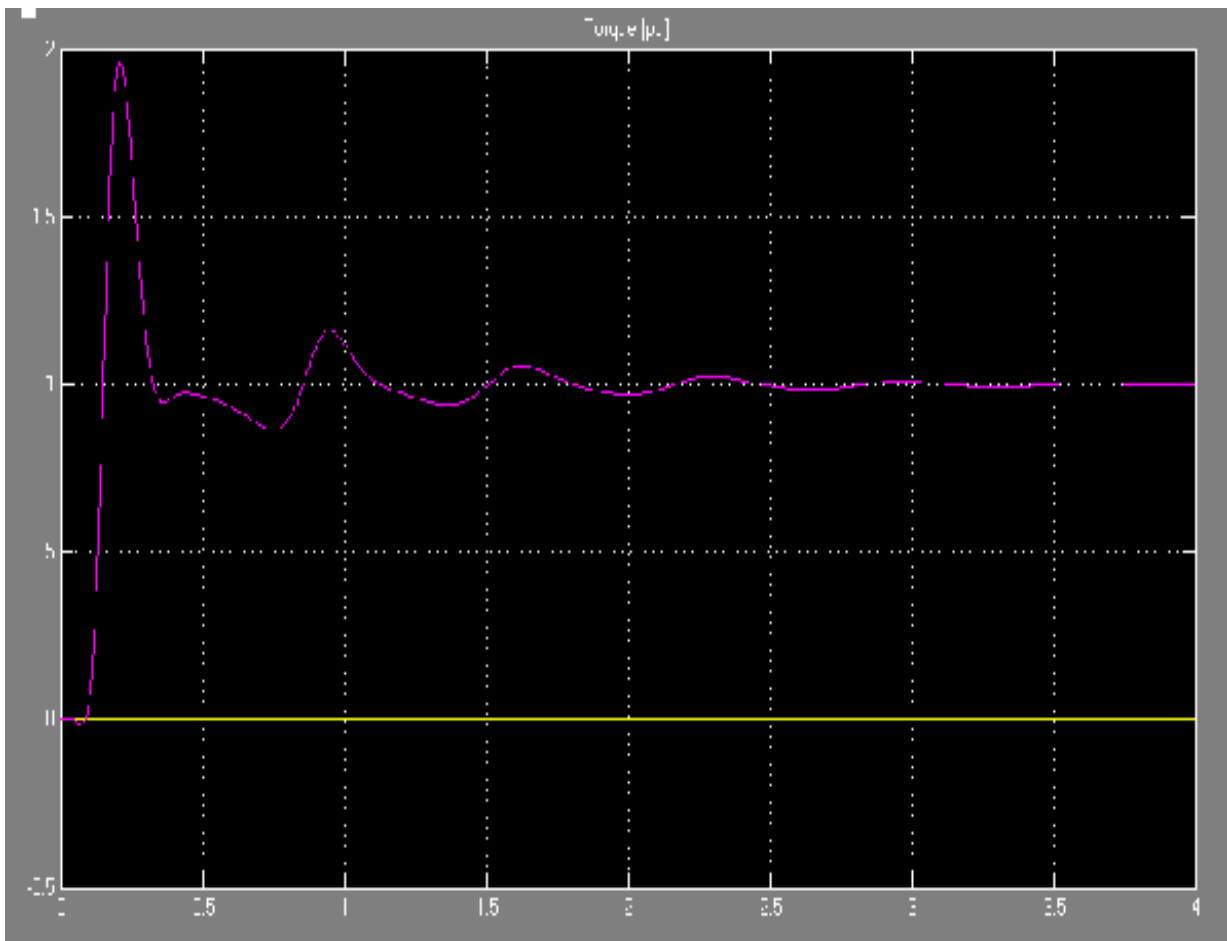


Fig 3.9 torque response for v/f speed control of IM

Here under it is shown the vector control of induction motor simulation results.

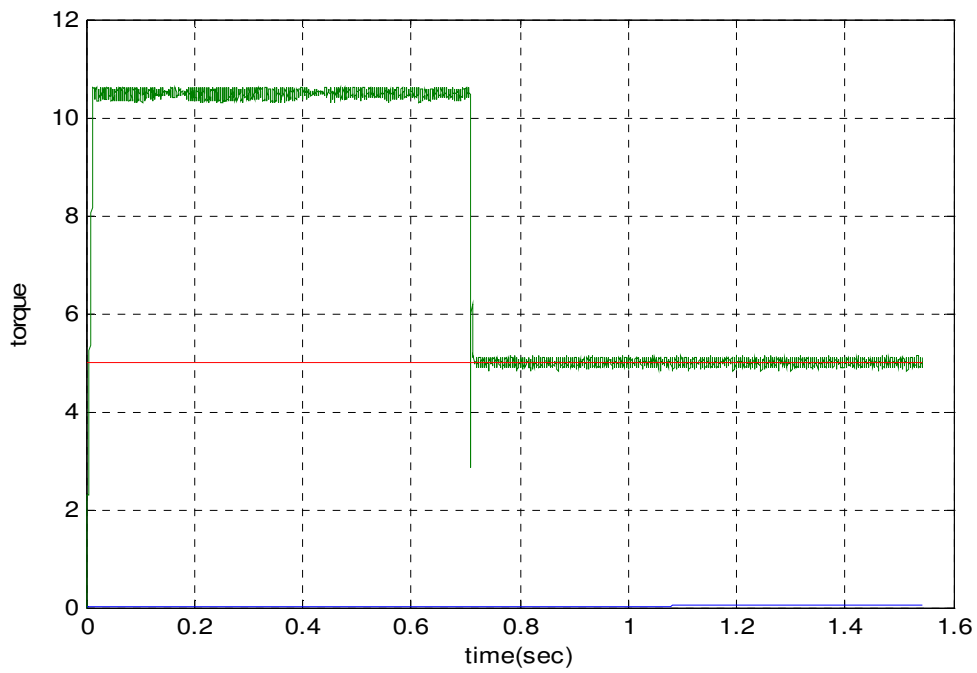


Fig3.10. torque response Vs time

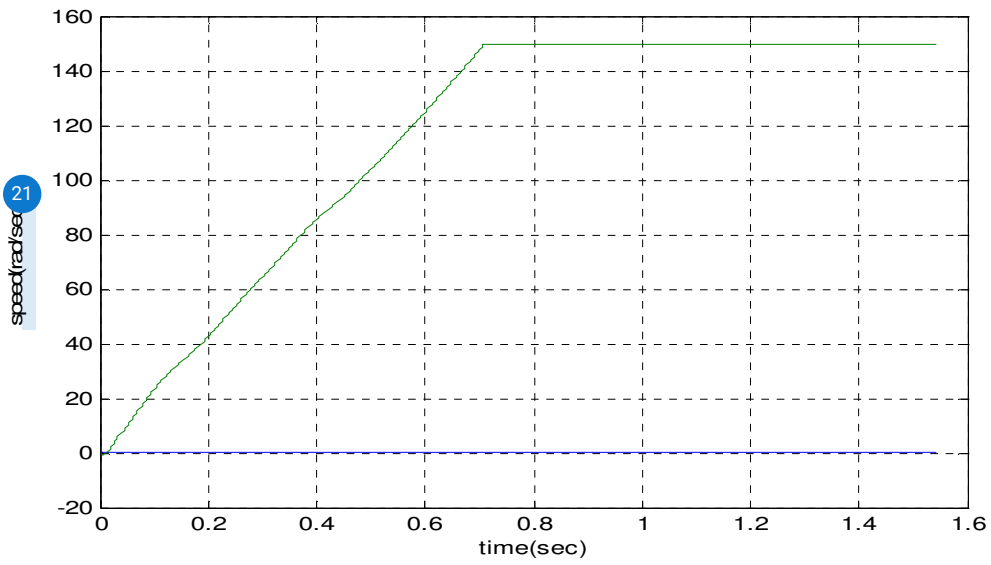


Fig3.11 Speed Vs time

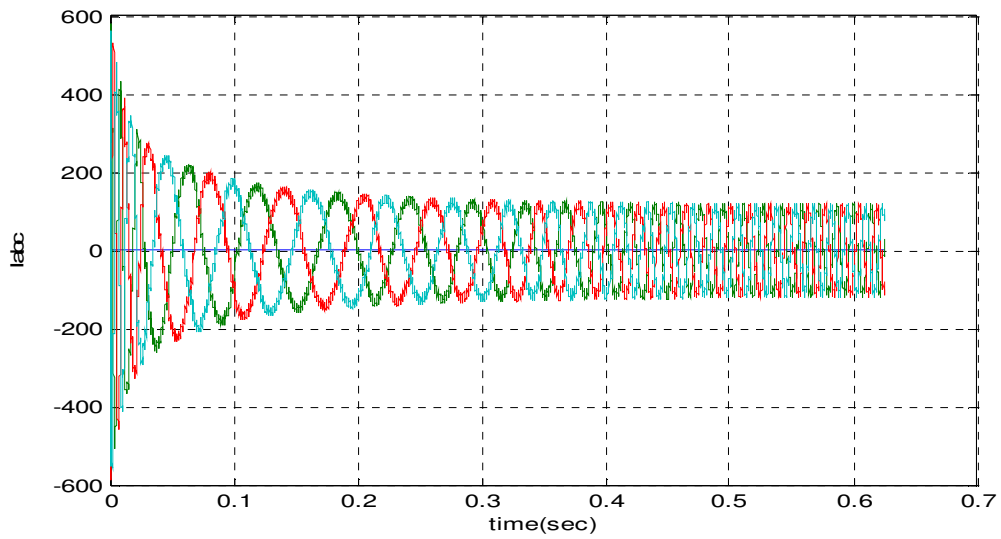


Fig3.12 armature Current Vs Time

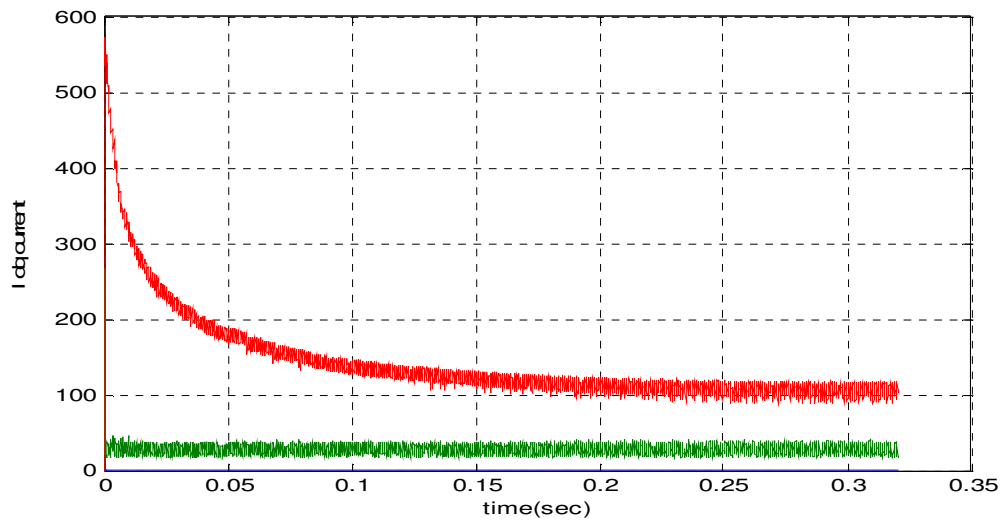


Fig 3.13 Idq current VS time

4. Conclusions and Recommendation

CONCLUSION

In this thesis induction motor dynamic model for vector control has been studied. DSP board type TMS320LF2407 which ²³ combines the performance of DSP architecture with the optimized peripherals of a Microcontroller in a single chip solution for the control of an induction motor has been programmed and interfaced with personal computer and voltage source inverter. ⁵¹ v/f control of induction motor has been implemented.

Due to limitations in time and shortage of some components like speed sensor for feed back, the vector control has not been realized experimentally, but simulated using Matlab.

The simulation result demonstrates that the dynamic performance of induction motor improves with vector control.

RECOMMENDATION

Induction motor vector control is one of the recent drive technologies to be familiarized in ECED lab.

Challenges in this thesis work were:

- DSP board configuration and initialization
- Implementation of space vector modulation
- Lack of some components

The two challenges have been overcome, and interested student can smoothly continue on this work for the realization of the induction motor vector control.

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6 □ Appendix

Software Code for the program:

```
#include "LF2407REGS.h" /*LF2407 register definition*/

#define MS_TIME_LOOP 0x0385

float taon,tbon,tcon;

int choise=30;

/** Wait state for the program**/

28 unsigned int wait_ms( volatile unsigned int delay_val )
{
    unsigned int i;
    unsigned int ms_ctr;
    for ( i = 0; i < delay_val; i++ )
    {
        ms_ctr = MS_TIME_LOOP;
        while ( ms_ctr )
```

```

        {
            ms_ctr--;
        }
    }
    return( delay_val );
}

/** SVPWM update for CMPR**/
void CalUpdateCMPRs()
{
    switch(choise)
    {
        case 30:
        {
            *CMPR4=0x300;
            *CMPR5=0x180;
            *CMPR6=0x120;
        }
        break;
        case 31:
        {
            *CMPR4=0x180;
            *CMPR5=0x300;
            *CMPR6=0x120;
        }
    }
}

```



```
}  
  
break;  
  
case 32:  
  
{  
  
*CMPR4=0x120;  
  
*CMPR5=0x300;  
  
*CMPR6=0x180;  
  
}  
  
break;  
  
case 33:  
  
{  
  
*CMPR4=0x120;  
  
*CMPR5=0x180;  
  
*CMPR6=0x300;  
  
}  
  
break;  
  
case 34:  
  
{  
  
*CMPR4=0x180;  
  
*CMPR5=0x120;  
  
*CMPR6=0x300;  
  
}  
  
break;
```

```

case 35:
{
    *CMPR4=0x300;
    *CMPR5=0x120;
    *CMPR6=0x180;
}
choise=30;
break;
}

float vd,vq,k;
float valpha,vbeta;
float x,y,z;
float t1,t2;
float taon,tbon,tcon;
float Ts=0x4000;
vd=0x10;
k=0x200;
vq=sqrt((k*k)-(vd*vd));
vbeta=vq*cos(theta)+vd*sin(theta);
valpha=-1.0*vq*sin(theta)+vd*cos(theta);
x=vbeta;
y=0.5*(sqrt(3)*valpha+vbeta);
z=0.5*(-1.0*sqrt(3)*valpha+vbeta);

```

```
switch(choise)
{
case 30:
if(theta=(pi/6))
{
t1=x;
t2=z*(-1.0);
} break;
case 90:
if(theta=(pi/2))
{
t1=z;
t2=y;
} break;
case 150:
if(theta=(5*pi/6))
{
t1=y*(-1.0);
t2=x;
} break;
case 210:
if (theta=(7*pi/6))
{
```

```

t1=x*(-1.0);
t2=z;
} break;
case 270:
if (theta=(3*pi/2))
{
t1=z*(-1.0);
t2=y*(-1.0);
} break;
case 330:
if (theta=(11*pi/6))
{
t1=y;
t2=x*(-1.0);
} break;
}
if(t1+t2<=Ts)
{
32taon=(Ts-t1-t2)/2;
tbon=taon+t1;
tcon=tbon+t2;
*CMPR4=taon;
*CMPR5=tbon;

```

```

    *CMPR6=tcon;

}

else

{
    32 t1=t1*Ts/(t1+t2);

    t2=t2*Ts/(t1+t2);
    32 taon=(Ts-t1-t2)/2;

    tbon=taon+t1;

    tcon=tbon+t2;

    *CMPR4=taon;

    *CMPR5=tbon;

    *CMPR6=tcon;

}

}

void initsys(){

    asm(" setc SXM ");

    asm(" clrc OVM ");

    asm(" clrc CNF ");

    /** configure the 34 system control and status registers**/

    *SCSR1=0x83fe;

    /** Disable the watchdog timer**/

    *WDCR =0x0e8;

    /** Setup the core interrupt**/

```

```

*IMR =0x000f;

*IFR =0xffff;

/* 8 Setup external memory interface for LF2407 EVM*/

*WSGR =0x00;

}

/** disable global interrupts**/

54 void inline disable(){
asm(" setc INTM ");
}

/** enable global interrupts**/

void inline enable(){
asm(" clrc INTM ");
}

void PWMinit(){

/** Setup shared I/O pins **/

*MCRC=*MCRC|0x007e;           //IOPE1~IOPE6 group C pins used as 6
PWM pins

*ACTRB=0x0666;               //PWM12□10□8 active low□PWM11□9□7 active
high

*DBTCONB=0;                  //Dead band unit off

*T3PER = 0x600;               //set timer 3 register□to send different on times to
CMPR4~CMPR6

*COMCONB=0x8200;             //compare operator

*T3CON = 0x1000;              // configure T3CON register

```

```

}

void interrupt nothing()

{
    return;
}

/** Initialization Module**/

void main()

{ disable();          //disable the interapt
  initsys();          //system initialization
  PWMinit();          //pwm initialization
  *T3CON=*T3CON|0x40;      //start timer 3

  do {
    CalUpdateCMPRs();
    wait_ms(15);
    choise=choise+1;
  }

  while(1);
}

```

Where :

/**¹⁶ Configure the System Control and Status registers ***/

*SCSR1 = 0x83Fe;

bit 15 1: reserved

bit 14 0: CLKOUT = CPUCLK

bit 13–12 00: IDLE1 selected for low–power mode

bit 11–9 001: PLL x4 mode

bit 8 1: reserved

bit 7 1: 1 = enable ADC module clock

bit 6 1: 1 = enable SCI module clock

bit 5 1: 1 = enable SPI module clock

bit 4 1: 1 = enable CAN module clock

bit 3 1: 1 = enable EVB module clock

bit 2 1: 1 = enable EVA module clock

bit 1 1: reserved

bit 0 0: clear the ILLADR bit

/** 8 Disable the watchdog timer ***/

*WDCR = 0x00E8;

bits 15–8 0's: reserved

bit 7 1: clear WD flag

bit 6 1: disable the dog

bit 5–3 101: must be written as 101

bit 2–0 000: WDCLK divider = 1

/** Setup external memory interface for LF2407 EVM ***/

WSGR = 0x0040;

bit 15–11 0's: reserved

bit 10–9 00: bus visibility off

bit 8–6 000: 0 wait–state for I/O space

bit 5–3 000: 0 wait–state for data space

bit 2–0 000: 0 wait state for program space620

/** Setup shared I/O pins */

MCRC=MCRC/0x007e

8 bit 15 0: reserved

* bit 14 0: 0=IOPF6, 1=IOPF6

* bit 13 0: 0=IOPF5, 1=TCLKINB

* bit 12 0: 0=IOPF4, 1=TDIRB

* bit 11 0: 0=IOPF3, 1=T4PWM/T4CMP

* bit 10 0: 0=IOPF2, 1=T3PWM/T3CMP

* bit 9 0: 0=IOPF1, 1=CAP6

* bit 8 0: 0=IOPF0, 1=CAP5/QEP4

* bit 7 0: 0=IOPE7, 1=CAP4/QEP3

* bit 6 0: 0=IOPE6, 1=PWM12

* bit 5 0: 0=IOPE5, 1=PWM11

* bit 4 0: 0=IOPE4, 1=PWM10

* bit 3 0: 0=IOPE3, 1=PWM9

* bit 2 0: 0=IOPE2, 1=PWM8

* bit 1 0: 0=IOPE1, 1=PWM7

* bit 0 0: 0=IOPE0, 1=CLKOUT

DBTCONA = 0x0000; / deadband units off */

Declaration

I, the undersigned, declare that this thesis work is my original work, has not been presented for a degree in this or any other universities, and all sources of materials used for the thesis work have been duly acknowledged.

Name: Abraham Belay

Signature: _____

Place: Addis Ababa

Date: _____

This thesis has been submitted for examination with my approval as a university advisor.

Name: Dr. Mengesha Mamo

Signature: _____

Place: Addis Ababa

Date:

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