

IMPROVEMENT IN THE PERFORMANCE OF DIRECT TORQUE CONTROL INDUCTION MOTOR DRIVES

Dissertation

submitted in partial fulfillment of the requirements for the award of degree of

Master of Engineering
in
Power Systems and Electric Drives

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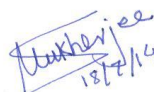
July, 2014

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CERTIFICATE

I hereby certify that the work which is being presented in the thesis entitled, "**Improvement in the Performance of Direct Torque Control Induction Motor Drives**", in partial fulfilment of the requirements for the award of degree of Master of Engineering in *Power Systems and Electric Drives* submitted in Electrical and Instrumentation Engineering Department of Thapar University, Patiala, is an authentic record of my own work carried out under the supervision of **Dr. Amrita Sinha**, Assistant Professor, EIED.

The matter presented in the thesis has not been submitted for award of any other degree of this or any other University.


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This is to certify that the above statement made by the candidate is correct and true to the best of my knowledge and belief.


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ABSTRACT

In this work, we have studied different control strategies of induction motor drives and with the technological advancement, various command approaches have been developed by the scientific community to master in real time, the flux and the torque of the electrical machines, the direct torque control (DTC) scheme being one of the most proven steps in this direction.

In principle, the motor terminal voltages and currents are sampled and used to estimate the motor flux and torque. Based on estimates of the flux position and the instantaneous errors in torque and stator flux magnitude, a voltage vector is selected to restrict the torque and the flux errors within their respective torque and flux hysteresis bands. In the conventional DTC, the selected voltage vector is applied for the whole switching period regardless of the magnitude of the torque error. This results in high torque ripple. A better drive performance can be achieved by varying the sector angle in modified DTC.

The space vector modulation technique (SVPWM) is applied to 2 level inverter with 12 sectors angular control in the DTC drive system to reduce the torque ripple. The results are simulated and compared using the MATLAB/ Simulink.

TABLE OF CONTENTS

	Page No
<i>Certificate</i>	<i>i</i>
<i>Acknowledgement</i>	<i>ii</i>
<i>Abstract</i>	<i>iii</i>
<i>Table of Contents</i>	<i>iv</i>
<i>Abbreviations</i>	<i>vi</i>
<i>List of Figures</i>	<i>vii</i>
<i>List of Tables</i>	<i>viii</i>

CHAPTER 1: INTRODUCTION

1.1	Overview	1
1.2	Features of Direct Torque Control (DTC)	3
1.3	Motivation	3
1.4	Objective	4
1.5	Review of Literature	4
1.6.	Organisation of the Thesis Work	9

CHAPTER 2: INDUCTION MOTOR MODEL AND ITS GENERALITIES

2.1	Equations of the Induction Machine Model	10
2.1.1	Introduction	10
2.1.2	Voltage Equations	11
2.1.3	Application of Park's transformation	12
2.1.4	Voltage matrix equations	14
2.1.4.1	Fixed to Stator	14
2.1.4.2	Fixed to Rotor	14
2.1.4.3	Fixed to Synchronism	14

CHAPTER 3: DIRECT TORQUE CONTROL OF INDUCTION MOTORS

3.1	Introduction	16
3.1.1	DC drives analogy	16
3.1.2	Principle of direct torque control of induction motor	17

3.2	DTC Controller	18
3.2.1	Voltage Source Inverter	18
3.3	DTC Schematic	21
3.3.1	Methods for Estimation of Stator Flux in DTC	22
3.3.2	Direct flux control	23
3.3.3	Torque hysteresis controller	24

CHAPTER 4: SPACE VECTOR PULSE WIDTH MODULATION FED INDUCTION MOTOR DRIVES

4.1	Introduction	25
4.2	Theory of Space vector pulse width modulation	25
4.2.1	Realization of voltage space phasor	25
4.2.2	Pulse pattern generation	27

CHAPTER 5: MODEL SIMULATION AND RESULTS

5.1	Improvement by Sector Change	29
5.2	Improvement through 12 Sectors SVPWM DTC	31
5.3	Induction Motor Parameters	33
5.4	Results for Conventional DTC	33
5.5	Results for Modified DTC	34
5.6	Results for 12 Sectors SVPWM DTC	36

CHAPTER 6: COMPARISON AND CONCLUSIONS

6.1	Comparison	38
6.2	Conclusion	39
6.3	Future Scope	39

REFERENCES

ABBREVIATIONS

AC	:	Alternating Current
ANN	:	Artificial Neural Network
D	:	Direct Axis
DC	:	Direct Current
DEG	:	Degree
DTC	:	Direct Torque Control
EMF	:	Electromotive Force
FOC	:	Flux Oriented Control
GA	:	Genetic Algorithm
HP	:	Horse Power
IGBT	:	Insulated Gate Bi Polar Transistor
IM	:	Induction Motor
PI	:	Proportional Integral
PSO	:	Particle Swarm Optimization
PWM	:	Pulse Width Modulation
Q	:	Quadrature Axis
SAT	:	Sector Advancing Technique
SVM	:	Space Vector Modulation
SVPWM	:	Space Vector Pulse Width Modulation
VSI	:	Voltage Source Inverter

LIST OF FIGURES

FIGURE NO.	NAME OF FIGURE	PAGE NO.
1.1	The Block Diagram for AC Motor Drive System	2
1.2	Basic Direct Torque Control (DTC) Scheme.	2
2.1	Cross section of Elementary Symmetrical 3 ϕ Induction Motor	11
2.2	Scheme of Equivalent Axis Transform	13
3.1	Basic Model of a Separately Excited DC Motor	16
3.2	Voltage Source Inverter	19
3.3	Stator Flux Vector Locus and Switching Voltage Vectors	20
3.4	Circular Trajectory of Stator Flux	23
3.5	Torque Hysteresis Controller	24
4.1	Inverter Switching State Vectors	26
4.2	Reference Vector in Sector 1	27
4.3	Leg Voltages and Space Vector Disposition in Sector 1	28
5.1	Simulink Model for conventional DTC	29
5.2	Implementation of Flux and Torque Error Through Voltage Vectors	30
5.3	Simulink Model for Modified DTC	31
5.4	Three Level Voltage Inverter Vectors	31
5.5	Implementation of 12 Sectors in the SVPWM DTC Simulink Model	32
5.6	Simulink Model for SVPWM based DTC	32
5.7	Stator Current Characteristics of Conventional DTC	33
5.8	Torque Characteristics of Conventional DTC	34
5.9	Rotor Speed Characteristics of Conventional DTC	34
5.10	Stator Current Characteristics of Modified DTC	35
5.11	Torque Characteristics of Modified DTC	35
5.12	Rotor Speed Characteristics of Conventional DTC	36
5.13	Stator Current Characteristics of 12 Sectors SVPWM DTC	36
5.14	Torque Characteristics of 12 Sectors SVPWM DTC	37
5.15	Rotor Speed Characteristics of 12 Sectors SVPWM DTC	37
6.1	Comparison of Stator Current Characteristics	38
6.2	Comparison of Torque Characteristics	39

LIST OF TABLES

TABLE NO.	NAME OF THE TABLE	PAGE NO.
3.1	General Selection Table for DTC, 'K' being the sector number	21
3.2	Conventional Look Up Table	21
5.1	Look Up Table for Modified DTC	30

CHAPTER 1

INTRODUCTION

1.1 Overview

Induction motors are the work horses in industries for variable speed applications starting from fractional horse power motors to multi megawatt motors. The application of Induction motor includes pumps, Paper and Textile mills, fans, locomotive propulsions, robotics, home appliances, air conditioning, rolling mills etc. Other than the process control, there is a lot of attention towards the energy saving aspects of variable frequency drives.

There is a considerable amount of increase in power handling capacities of the power electronic devices observed during the recent development in the semi-conductor technologies. However in the multi-level inverters, there is distortion in the output voltages due to lower order harmonics when compared with standard two-level inverter at same frequency. Control of induction motor is a challenge in itself owing to fast motor dynamics and non-linearity in mathematical models of the machines. Therefore the task is to derive a suitable mathematical model of the motor.

Two comparable control techniques in modern automated system with respect to ac motor drives are: (a) Field Oriented Control (FOC) popularly known as Vector Control and (b) Direct Torque Control (DTC). Almost 40 – 43 years back, F.Blaschke in year 1971 presented a paper on induction motor control through Field Oriented Control (FOC) strategy. Now this technique is considered to be a mature one from industrial point of view and several drives of this type are available in the market. The block diagram for AC motor drive system is given in figure 1.1. The Field Oriented Control techniques are very complex and highly sensitive to inaccuracy in parameters of the motor. A considerable amount of research is already devoted in this field. Now the aim is to develop a simpler technique in speed control. One method known as Direct Torque Control (DTC) is popular and was coined by I. Takahashi and M. Depenbrock for voltage fed PWM inverter drives during mid 1980s and today this control scheme is considered as the world's most advanced AC Drives control technology. This is a simple control

technique which does not require coordinate transformation, PI regulators, and Pulse width modulator and position encoders

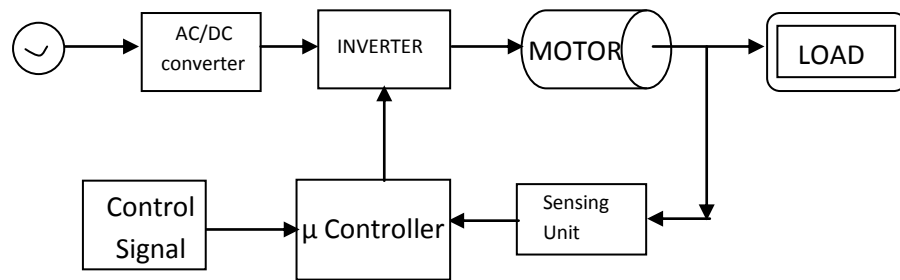


Figure 1.1 Block Diagram of AC Motor Drive System.

This technique displays results which are comparable in nature with vectored controlled drives. The scheme involves direct control of torque and stator flux of the drive by inverter voltage space vectored selection using look-up tables. In conventional DTC, the electromagnetic torque and flux are independently controlled through suitable selection of optimum inverter switching modes. This is made to limit the errors in electromagnetic torque and flux linkages within torque and flux hysteresis bands. The basic DTC technique shown in figure 1.2 consists of major blocks as two comparators with specified bandwidth, switching table, flux and torque estimator, voltage source inverter.

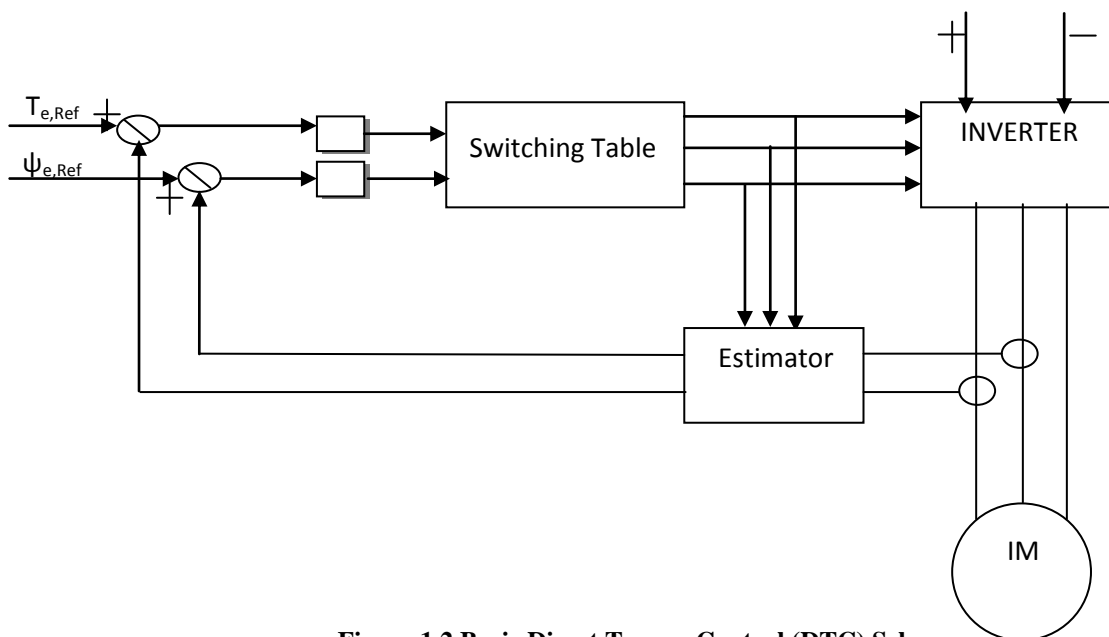


Figure 1.2 Basic Direct Torque Control (DTC) Scheme

1.2 Features of Direct Torque Control (DTC)

The main features of DTC are as follows

- (a) The flux and torque are directly controlled.
- (b) The stator voltages and currents are indirectly controlled.
- (c) Approximately sinusoidal stator fluxes.
- (d) At standstill condition high dynamic performance.

The main advantages of DTC are as follows

- (a) Absence of co-ordinate transforms.
- (b) Absence of voltage modulator block, as well as other controllers such as PID for motor flux and torque.
- (c) Minimal torque response time, even better than the vector controllers.

Some of the disadvantages associated with DTC are as follows

- (a) Possible problems during starting.
- (b) Requirement of torque and flux estimators, implying the consequent parameters identification.
- (c) Inherent torque and stator flux ripple.

1.3 Motivation

The electric drives used in industry are Adjustable Speed Drives and in most of these drives AC motors are applied. Induction motors are the standard in these drives. Induction motors are today the most widely used ac machines due to the advantageous mix of low cost, reliability and performance. So effective control of IM parameters e.g. speed, torque and flux is of utmost importance. From the investigation of the control methods it is known that torque control of IM can be achieved according to different techniques ranging from inexpensive Volts/Hz ratio strategy to sophisticated sensorless vector control scheme. But every method has its own disadvantages like losses, need of separate current control loop, coordinate transformation (thus increasing the complexity of the controller), torque and current ripple etc. So it is very much necessary to design a controller to obtain an ideal electric vehicle motor drive system which would have high efficiency, low torque ripple and minimum current distortion.

1.4 Objective

The objective was to analyse the limitations of the conventional direct torque control drives and to develop some modified technique for DTC in order to improve its performance. The application of space vector pulse width modulation technique on DTC for further improvement its performance. The above techniques are modelled using MATLAB/Simulink and the analysis was carried out through results.

1.5 Review of Literature

Joachim Bocker and Shashidhar Mathapati [1], have illustrated the historical and recent developments in the field of control of Induction motors and its major milestones. Subsequently they have proposed how in today's industrial standards, the results are translated and finally what are the recent trends in industries and research in respect of control of induction motor.

Dal Y. Ohm [2], have proposed that the dynamic model of induction motor is necessary for understanding and analysing vector control of induction motor. But the dynamic model equations are very complex and it has many different forms of model based on the choice of the reference frame. They have explained various forms in a concise way in order to make it understand clearly. Finally, the basic principles of vector control were discussed generally along with the development of fundamental dynamic mechanism of the motor in respect to synchronous frame.

Inoue.et.al [3], have explained a control method for permanent magnet synchronous motor drive system based on direct torque control technique suitable for limited armature voltage and current. Firstly, they proposed the flux-weakening and torque limiting control that is required for a direct torque control based drive system. They have used a mathematical model in rotating reference frame synchronized with stator flux linkage. Secondly, it utilizes an anti-windup scheme for torque controller used in direct torque system. The degradation in performance of torque limiting (current limiting) control is observed due to windup of the controller. Hence with application of anti-windup, there is an improvement in the performance of the proposed torque limiting method especially in transient state. The proposed system achieved stable control with direct torque control

based drives combined with a speed controller. The effectiveness of the model has been confirmed experimentally.

Hoang Le-Hung [4], has done a comparative study between two most important and popular control techniques for induction motor drives, i.e. Field Oriented Control (FOC) and Direct Torque Control (DTC). The comparison has been carried out involving various criteria like basic control characteristics, dynamic performance, implementation complexities and parameter sensitivity. It has been studied with the help of simulation using *Simulink Power System Block Set* which contains various representations of power system (inverter and induction motor) and the control system. Actual parameters of a 50 HP induction motor fed with IGBT PWM inverter used for simulation and evaluation of both the control techniques.

A. A. Ansari and D.M Deshpande [5], have proposed different models of Induction Motor and guidelines for their use along with their typical results. Firstly they developed a mathematical model of the small power induction motor and carried out dynamic simulation of the model. The dynamic simulation is a key tool for validation of the design process of the drive system of the motor. It is also required for prototype construction and testing in order to eliminate inadvertent design mistakes and errors. Secondly it has demonstrated the simulation of the steady state performance of induction motor through MATLAB program. The modelling of a 3-Phase induction motor and its simulation has been done through Simulink.

R J Lee et. al [6], have explained the equation of reference frames using 2 axis (d & q) theory for simulation of induction machines. The case studies have been considered in order to demonstrate that depending upon the problem and the type of computer available the reference frames are required to be chosen.

Wu Tao and Zhoo Liang [7], has described the theory behind the mathematical model of induction motor. The model of induction motor vector control has been developed with simulation tool-kit of MATLAB / Simulink. The details of various subsystems are also discussed in this paper. The simulation is used to verify the correctness of the model and effectiveness of the control system. The system gives good dynamic and static performance.

Praveen S Babu and Dr. S Ushakumari [8], have explained the main drawback of the conventional Direct Torque Control technique ie. high torque, flux ripples and variable switching frequency of the devices. The drawback has been rectified using space vector modulation (SVM) technique. They developed two modified direct torque control schemes using Proportional Integral (PI) controller and sliding mode controllers. The analysis of the same was confirmed by simulation using MATLAB / Simulink.

M. Lakshmi Swarupa. et. al [9] have explained new scheme of direct torque control of induction motor used in electric vehicles and a suitable torque control scheme. The drive used in electric vehicle comprises of rewound induction motor and a three level IGBT inverter. The schemes under consideration were (1) Field – Oriented Control, (2) Direct Torque Control and (3) Space Vector Modulation based DTC. The results were compared in Simulink and the best result was obtained for Space Vector Modulation based DTC.

Emesto Ruppert Filho and Ronaldo Martins de Souza [10] have presented the dynamic model of induction motor, frequently used by electrical machine researchers in motor dynamic studies, constituted with the help of four voltage differential equations and one mechanical differential equation. The three phase induction motor dynamic mathematical model being comprehensive and includes the skin effect, the temperature influences on the motor parameters and allows for the stator and rotor winding and core average temperature evaluation. This model has been useful for motor speed changing, intermittent loading analysis and for those motors which are fed from non-sinusoidal voltage. The model can be utilised for energy conservation and power quality subject analysis.

H.F. Abdul Wahab and H. Sanusi [11], have presented DTC as an alternative to the Field Oriented Control (FOC) or vector control technique. Though the two control schemes are different in their operating principle, but their objective are almost same. Their aim being effective control of torque and flux. Comprehensive studies of DTC strategy for torque control of induction machine has been presented in this paper. The performance of the proposed control technique has been demonstrated with simulations using Simulink / MATLAB. Also several numerical simulations were carried out in transient and steady state operation for speed control.

B.Karanayil and M.F.Rahan [12], have proposed and validated the complete dynamic model of induction motor, with an indirect rotor flux oriented vector controller with the help of Simulink. The Simulink model takes the dynamic model of three phase induction motor along with inverter and its control system. The model has been developed for the study of the effects of the parameter variations in a vector controlled drive, also development being carried out for parameter estimator using Artificial Neural Network (ANN) and Fuzzy logic.

Narasimamhan PVR. et.al [13], have represented the improvement in the dynamics associated with DTC drives by using estimators, observers and machine intelligence systems. The paper has proposed Sector Advancing Technique (SAT) for reduction in the response time of the drive for known torque. The validation of the proposed technique being carried using Simulink models and the performance of SAT with DTC has been found to be better than conventional DTC.

C.J.Bonanno et. al [14], have illustrated with the help of voltage – current mode algorithm or physical based algorithm the rotor flux can be estimated for a direct field oriented induction machine.

Hind Djeghloud and Hocine Benalla [15], have investigated the multilevel converters which has provided high safety voltages with less harmonic components compared to the two-level inverters. Employed for converter's gating signals generation, the space-vector pulse width modulation (SVPWM) strategy reduces the switching losses by limiting the switching to the two thirds of the pulse duty cycle. It has been explained, via simple and easy algebraic expressions, the application of the SVPWM on the three-level voltage inverter.

Manoj Kumar Sahu. et.al [16], have discussed the direct torque control (DTC) as a control technique in AC drive systems to obtain high performance torque control. The classical DTC drive contains a pair of hysteresis comparators and suffers from variable switching frequency and high torque ripple. This problem has been solved by the use of space vector depending on the reference torque and flux. The space vector modulation technique has been applied to the three-level neutral Point Clamped inverter control in the

proposed DTC-based induction motor drive system, resulting to a significant reduction of torque ripple.

K. Suman et.al [17], have explained the application of space vector modulation technique in (SVPWM) 2 level inverter control DTC-based induction motor drive system and has dramatically reduced the torque ripple. Thereafter the controller based on space vector modulation has been designed to be applied in the control of Induction Motor (IM) with a three-level Inverter which has resulted into further improvement.

Subrata K Mondal and Bimal K Bose [18] have explained the importance of multilevel voltage fed inverter with SVPWM in high power industrial drives applications. They have also explained the switching philosophy for both the under modulation and over modulation algorithm. The switching states could be designed in such a way that the neutral point voltage always remains balanced.

N. Venkataramana Naik and S.P.Singh[19] have presented the scalar control as well as direct torque control for speed control of induction motor. They have demonstrated that the SVPWM could be utilized for better voltage regulation and better dynamic response of induction motor through considerable reduction in harmonics in voltage as well as current of the inverter.

Mohamed.M.Ismail[20], has proposed GA and PSO algorithm for improvement of speed control for two level and three level inverter and thereby mitigating the problem of controlling the Π model induction motor with magnetic saturation.

Guo Qing and Gao Yun[21], have developed a asynchronous motor direct torque control system where in the size of flux determines the switching time of non zero voltage vector and thus affects the hexagonal flux trajectory. An open loop direct torque control system has been developed and simulated through MATLAB for validation of the proposed theory.

1.6. Organisation of the Thesis Work

Chapter II gives a review of the induction motor modelling. It discusses the mathematical model of three phase induction motor, the concept of reference frames and the induction motor dynamic equations in various reference frames, the equations obtained in stationary reference frame are required to be applied on further chapters during the course of work.

Chapter III covers the fundamental principles of DTC strategy of induction motors, the building up of the look up tables and the limitations of the conventional DTC drives. The voltage vector diagram is also discussed.

Chapter IV explains the Space Vector Pulse Width Modulation based technique, its basic principle and building up of lookup table from the voltage vector diagram and its application in Direct Torque Control (DTC).

Chapter V deals with the Simulation of the various Models developed and their results.

Chapter VI provides the comparison of the results obtained after simulation and conclusions.

CHAPTER 2

INDUCTION MOTOR MODEL AND ITS GENERALITIES

2.1 Equations of the Induction Machine Model

2.1.1 Introduction

A dynamic model of the machine subjected to control should be known before hand in order to understand and design vector controlled drives. Due to the fact that every good control has to face some possible change of the plant, it could be said that the dynamic model of the machine should be a good approximation of the real plant. Moreover, the model should incorporate all the important dynamic effects occurring during both transient and steady-state operations. Further, it should be valid for any changes in the inverter's supply such as voltages or currents [22] [23]. Such a model can be obtained by means of either the space vector phasors theory or two-axis theory of electrical machines. Despite of the compactness and the simplicity of the space phasors theory, both methods are actually close to each other. The cross section of elementary symmetrical three phase induction motor is given in figure 2.1. For simplicity, the induction motor under consideration will have the following assumptions:

- Three phase winding with symmetrical two-poles.
- Neglect slotting effects.
- The permeability of the iron parts is considered infinite.
- The flux density is radial along the air gap.
- Iron losses are neglected.
- The stator and the rotor windings can be simplified as a single or multi-turn full pitch coils situated in between two sides of the air gap.

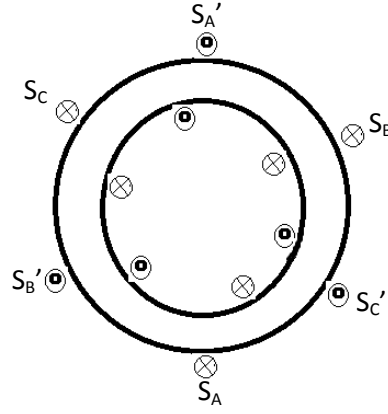


Figure 2.1. Cross section of elementary symmetrical three phase induction motor.

2.1.2 Voltage Equations

The stator voltages will be formulated in this section from the motor natural frame, known as the stationary reference frame which is fixed to the stator. In a similar way, the rotor voltages will be obtained with respect to the rotating frame fixed to the rotor. Thus in the stationary reference frame, the equations can be obtained as follows:

$$V_{sA} = R_s I_{sA} (t) + \frac{d\phi_{sA}(t)}{dt} \text{----- (2.1)}$$

$$V_{sB} = R_s I_{sB} (t) + \frac{d\phi_{sB}(t)}{dt} \text{----- (2.2)}$$

$$V_{sC} = R_s I_{sC} (t) + \frac{d\phi_{sC}(t)}{dt} \text{----- (2.3)}$$

Similar expressions can be obtained for the rotor:

$$V_{rA} = R_r I_{rA} (t) + \frac{d\phi_{rA}(t)}{dt} \text{----- (2.4)}$$

$$V_{rB} = R_r I_{rB} (t) + \frac{d\phi_{rB}(t)}{dt} \text{----- (2.5)}$$

$$V_{rC} = R_r I_{rC} (t) + \frac{d\phi_{rC}(t)}{dt} \text{----- (2.6)}$$

The instantaneous stator flux linkage per phase can be expressed as:

$$\begin{aligned}
\varphi_{sA} &= L_s I_{sA} + M_s I_{sB} + M_s I_{sC} + M_{sr} \cos \theta_m I_{ra} + M_{sr} \cos(\theta_m + \frac{2\pi}{3}) I_{rb} + M_{sr} \cos(\theta_m + \frac{4\pi}{3}) I_{rc} \\
\varphi_{sB} &= M_s I_{sA} + L_s I_{sB} + M_s I_{sC} + M_{sr} \cos(\theta_m + \frac{4\pi}{3}) I_{ra} + M_{sr} \cos \theta_m I_{rb} + M_{sr} \cos(\theta_m + \frac{2\pi}{3}) I_{rc} \\
\varphi_{sC} &= M_s I_{sA} + M_s I_{sB} + L_s I_{sC} + M_{sr} \cos(\theta_m + \frac{2\pi}{3}) I_{ra} + M_{sr} \cos(\theta_m + \frac{4\pi}{3}) I_{rb} + M_{sr} \cos \theta_m I_{rc}
\end{aligned}
\tag{2.7}$$

In a similar way, the rotor flux linkages can be expressed as follows:

$$\begin{aligned}
\varphi_{ra} &= M_{sr} \cos(-\theta_m) I_{sA} + M_{sr} \cos(-\theta_m + \frac{2\pi}{3}) I_{sB} + M_{sr} \cos(-\theta_m + \frac{4\pi}{3}) I_{sC} + L_r I_{ra} + M_r I_{rb} + M_r I_{rc} \\
\varphi_{rb} &= M_{sr} \cos(-\theta_m + \frac{4\pi}{3}) I_{sA} + M_{sr} \cos(-\theta_m) I_{sB} + M_{sr} \cos(-\theta_m + \frac{2\pi}{3}) I_{sC} + L_r I_{rb} + M_r I_{ra} + M_r I_{rc} \\
\varphi_{rc} &= M_{sr} \cos(-\theta_m + \frac{2\pi}{3}) I_{sA} + M_{sr} \cos(-\theta_m + \frac{4\pi}{3}) I_{sB} + M_{sr} \cos(-\theta_m) I_{sC} + L_r I_{rc} + M_r I_{rb} + M_r I_{ra}
\end{aligned}
\tag{2.8}$$

Taking into account for all the previous equations from 2.7 to 2.8, and using the matrix notation for compactness of all the expressions, the following expression is thus obtained

$$\begin{bmatrix} V_{sA} \\ V_{sB} \\ V_{sC} \\ V_{ra} \\ V_{rb} \\ V_{rc} \end{bmatrix} = \begin{bmatrix} R_s + pL_s & pM_s & pM_s & pM_{sr} \cos \theta_m & pM_{sr} \cos \theta_{mi} & pM_{sr} \cos \theta_{mi} \\ pM_s & R_s + pL_s & pM_s & pM_{sr} \cos \theta_{mi} & pM_{sr} \cos \theta_m & pM_{sr} \cos \theta_{mi} \\ pM_s & pM_s & R_s + pL_s & pM_{sr} \cos \theta_{mi} & pM_{sr} \cos \theta_{mi} & pM_{sr} \cos \theta_m \\ pM_{sr} \cos \theta_m & pM_{sr} \cos \theta_{mi} & pM_{sr} \cos \theta_{mi} & R_r + pL_r & pM_s & pM_s \\ pM_{sr} \cos \theta_{mi} & pM_{sr} \cos \theta_m & pM_{sr} \cos \theta_{mi} & pM_s & R_r + pL_r & pM_s \\ pM_{sr} \cos \theta_m & pM_{sr} \cos \theta_{mi} & pM_{sr} \cos \theta_m & pM_s & pM_s & R_r + pL_r \end{bmatrix} \begin{bmatrix} i_{sA} \\ i_{sB} \\ i_{sC} \\ i_{ra} \\ i_{rb} \\ i_{rc} \end{bmatrix}
\tag{2.9}$$

2.1.3 Application of Park's transformation

In order to reduce the expressions of the induction motor equation voltages given in equation 2.1, 2.2, 2.3, 2.4, 2.5 and 2.6 and obtain constant coefficients in the differential equations, the Park's transform needs to be applied. Physically, it can be understood as transforming the three windings of the induction motor to just two windings, which is shown in figure 2.2 [22].

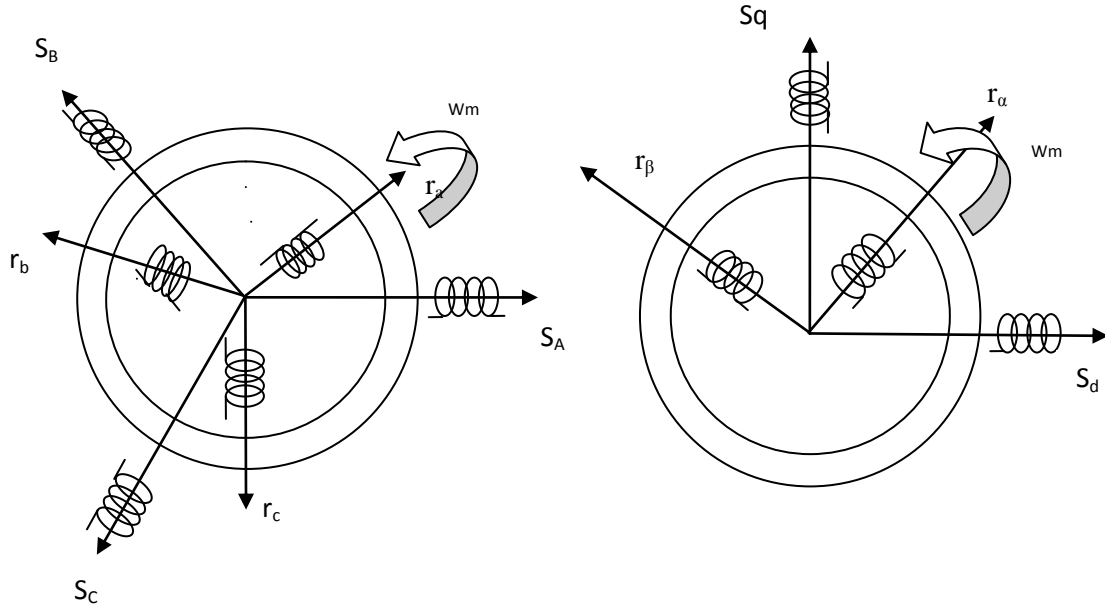


Figure 2.2. Scheme of equivalent axis transformation

In the symmetrical three-phase machine, the direct and the quadrature-axis stator magnitudes are considered to be fictitious. The equivalent for these direct (D) and quadrature (Q) magnitudes with the magnitudes per phase are given as:

$$\begin{bmatrix} V_{sD} \\ V_{sQ} \\ V_{s0} \end{bmatrix} = c \begin{bmatrix} \cos \theta & \cos(\theta - 2\pi/3) & \cos(\theta + 2\pi/3) \\ -\sin \theta & -\sin(\theta - 2\pi/3) & -\sin(\theta + 2\pi/3) \\ 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \end{bmatrix} \begin{bmatrix} V_{sA} \\ V_{sB} \\ V_{sC} \end{bmatrix} \quad \text{----- (2.10)}$$

$$\begin{bmatrix} V_{sA} \\ V_{sB} \\ V_{sC} \end{bmatrix} = c \begin{bmatrix} -\cos \theta & -\sin \theta & 1/\sqrt{2} \\ \cos(\theta - 2\pi/3) & -\sin(\theta - 2\pi/3) & 1/\sqrt{2} \\ \cos(\theta + 2\pi/3) & -\sin(\theta + 2\pi/3) & 1/\sqrt{2} \end{bmatrix} \begin{bmatrix} V_{sD} \\ V_{sQ} \\ V_{s0} \end{bmatrix} \quad \text{----- (2.11)}$$

where "c" is a constant that can take either the values 2/3 or 1 for the non-power invariant form or the value 2 / 3 is selected for the power-invariant form. These previous equations can be applied for any other magnitudes such as currents and fluxes. Now the expression 2.9 can be further simplified into a much smaller expression in 2.12 by means of applying the above mentioned Park's transform.

$$\begin{bmatrix} V_{sD} \\ V_{sQ} \\ V_{r\alpha} \\ V_{r\beta} \end{bmatrix} = \begin{bmatrix} R_s + pL_s & -L_s p\theta_s & pL_m & -L_m(P\omega_m + p\theta_r) \\ L_s p\theta_s & R_s + pL_s & L_m(P\omega_m + p\theta_r) & pL_m \\ pL_m & -L_m(p\theta_s - P\omega_m) & R_r + pL_r & -L_r p\theta_r \\ L_m(p\theta_s - P\omega_m) & pL_m & L_r p\theta_r & R_r + pL_r \end{bmatrix} \begin{bmatrix} i_{sD} \\ i_{sQ} \\ i_{r\alpha} \\ i_{r\beta} \end{bmatrix} \quad \text{-----}(2.12)$$

where,

$$L_s = L_s - L_m, L_r = L_r - L_m \text{ and } L_m = \frac{3}{2} M_{sr} \quad \text{-----} (2.13)$$

2.1.4 Voltage matrix equations

If the matrix expression 2.12 is simplified, new matrixes are obtained as shown in equations 2.14, 2.15 and 2.16.

2.1.4.1 – Fixed to the stator

It means that $w_s = 0$ and consequently $w_r = -w_m$

$$\begin{bmatrix} V_{sD} \\ V_{sQ} \\ V_{rd} \\ V_{rq} \end{bmatrix} = \begin{bmatrix} R_s + pL_s & 0 & pL_m & 0 \\ 0 & R_s + pL_s & 0 & pL_m \\ pL_m & P\omega_m L_m & R_r + pL_r & P\omega_m L_r \\ L_m(p\theta_s - P\omega_m) & pL_m & -P\omega_m L_r & R_r + pL_r \end{bmatrix} \begin{bmatrix} i_{sD} \\ i_{sQ} \\ i_{rd} \\ i_{rq} \end{bmatrix} \quad \text{-----} (2.14)$$

2.1.4.2 – Fixed to the rotor

It means that $w_r = 0$ consequently $w_s = w_m$

$$\begin{bmatrix} V_{sD} \\ V_{sQ} \\ V_{rd} \\ V_{rq} \end{bmatrix} = \begin{bmatrix} R_s + pL_s & -L_s p\omega_m & pL_m & -L_m P\omega_m \\ L_s P\omega_m & R_s + pL_s & L_m P\omega_m & pL_m \\ pL_m & 0 & R_r + pL_r & 0 \\ 0 & pL_m & 0 & R_r + pL_r \end{bmatrix} \begin{bmatrix} i_{sD} \\ i_{sQ} \\ i_{rd} \\ i_{rq} \end{bmatrix} \quad \text{-----} (2.15)$$

2.1.4.3 – Fixed to the synchronism

$w_r = s w_s$

$$\begin{bmatrix} V_{sD} \\ V_{sQ} \\ V_{rd} \\ V_{rq} \end{bmatrix} = \begin{bmatrix} R_s + pL_s & -L_s \omega_s & pL_m & -L_m \omega_s \\ L_s \omega_s & R_s + pL_s & L_m p \omega_s & pL_m \\ pL_m & -L_m s \omega_s & R_r + pL_r & -L_r s \omega_s \\ L_m s \omega_s & pL_m & L_r s \omega_s & R_r + pL_r \end{bmatrix} \begin{bmatrix} i_{sD} \\ i_{sQ} \\ i_{rd} \\ i_{rq} \end{bmatrix} \quad \text{----- (2.16)}$$

After finding out the voltage matrix, these vectors are used for the Space Phasor Notations for current phasors which allows the transformation of the natural instantaneous values of the three phase system into a complex plane located in the cross section of the motor. In this plane, the space phasor rotates with an angular speed equal to the angular frequency of the supply of the three phase system. The torque expression is further derived by means of energy consideration.

CHAPTER 3

DIRECT TORQUE CONTROL OF INDUCTION MOTORS

3.1 Introduction

3.1.1 DC drives analogy

In a dc machine, neglecting the armature reaction effect and field saturation, the developed torque is expressed as

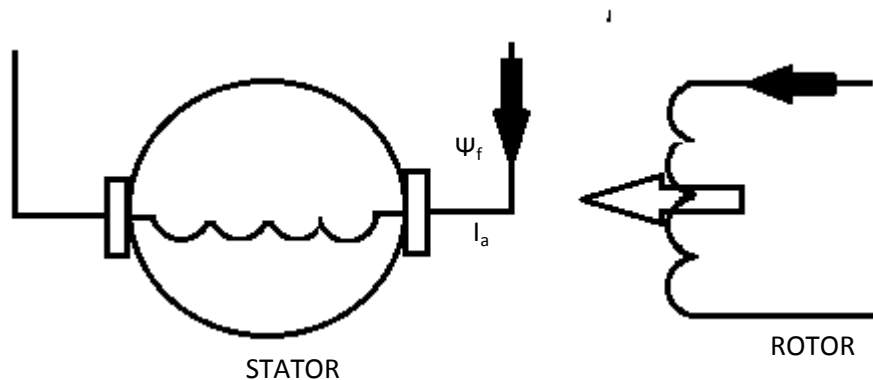


Figure 3.1 Basic model of a separately excited DC motor

$$T_e = K_a \cdot I_a \cdot I_f \quad \text{----- (3.1)}$$

Where I_a = armature current and I_f = field current

The construction of a DC machine is such that the field flux ψ_f produced by the field current I_f is perpendicular to the armature flux ψ_a , which is produced by the armature current I_a . These space vectors, being stationary in space, are orthogonal to each other or decoupled in nature as shown in figure 3.1. This means that when torque can be controlled by controlling the current I_a , the field flux ψ_f is not effected and thus we get the fast transient response and high torque/ampere ratio at the rated ψ_f . Because of this decoupling, when the field current I_f is controlled, it affects only the field flux ψ_f , but not the armature flux ψ_a . The DC machine like performance is being observed in Direct Torque Controlled (DTC) drives. In DTC drives, the de-coupling of the torque and flux components can be achieved by using hysteresis comparators which compares the actual and estimated values of the electromagnetic torque and stator flux. The DTC drive

consists of a controller, torque calculator and flux calculator, along with a Voltage Source Inverter (VSI).

3.1.2 Principle of direct torque control of induction motor

Direct torque control was developed by Takahashi and Depenbrock as an alternative to field-oriented control [22][24]. In a direct torque controlled (DTC) induction motor drive which gets supply from a voltage source inverter, it is possible to control the stator flux linkage ψ_s (or the rotor flux ψ_r or the magnetizing flux ψ_m) and the electromagnetic torque directly with the selection of an optimum inverter voltage vector. The selection of different voltage vectors of the voltage source inverter are made in such a way that their flux and torque error restricts within their respective flux and torque hysteresis bands and to obtain the fastest torque response and higher efficiency during every operation. DTC enables both quick torque response in the transient operation and reduction of the harmonic losses and acoustic noise. As it has been introduced, the electromagnetic torque in the three phase induction machines can be expressed as follows [22]:

$$t_e = \frac{3}{2} P \psi_s \times i_s \quad \text{----- (3.2)}$$

Where ψ_s is the stator flux, i_s is the stator current (both fixed to the stationary reference frame fixed to the stator) and P the number of pole pairs. The previous equation can be modified and given as follows:

$$t_e = \frac{3}{2} P \psi_s \times i_s \sin(\alpha_s - \rho_s) \quad \text{----- (3.3)}$$

Where ρ_s is the stator flux angle and α_s is the stator current angle both referred to the horizontal axis of the stationary frame which is fixed to the stator. If the modulus of the stator flux is kept constant and the stator flux angle ρ_s is changed very quickly, then the electromagnetic torque can be directly controlled. The same conclusion can be obtained using another expression for the electromagnetic torque which can be written as:

$$t_e = \frac{3}{2} P \frac{L_m}{L}$$

$$t_e = \frac{3}{2} P \frac{L_m}{L_s L_r - L_m^2} \psi_r \psi_s \sin(\rho_s - \rho_r) \quad \text{----- (3.4)}$$

Because of the rotor time constant is larger than the stator one, the rotor flux changes slowly compared to the stator flux, in fact, the rotor flux can be assumed to be almost constant. (The fact that the rotor flux can be assumed constant is true as long as the response time of the control is much faster than the rotor time constant). As long as the modulus of the stator flux is kept constant, then the required electromagnetic torque can be rapidly changed and controlled by means of changing the angle difference $(\rho_s - \rho_r)$. [22] [25].

3.2 DTC Controller

The one way to impose the required stator flux is by means of finding out the most suitable voltage source inverter state. If the ohmic drops are neglected for simplicity, then the stator voltage can be related directly with the stator flux in accordance with the following equations

$$\frac{d\psi_s}{dt} = V_s \quad \text{----- (3.5)}$$

or

$$\nabla\psi_s = V_s \Delta t \quad \text{----- (3.6)}$$

Decoupled control of the stator flux modulus and torque can be achieved by acting on the radial and tangential components respectively of the stator flux-linkage space vector in its locus. These two components are directly proportional ($R_s=0$) to the components of the same voltage space vector in the same directions. So selection of proper voltage vector is an important phenomenon in direct torque control of induction motor. This can be achieved by using voltage source inverter.

3.2.1 Voltage Source Inverter

There are many topologies for the voltage source inverter used in DTC control of induction motors that give high number of possible output voltage vectors [23] [24] but the most common one is the six step inverter. A six step voltage inverter provides the variable frequency AC voltage input to the induction motor in DTC method. The DC supply to the inverter is provided either by a DC source like a battery, or a rectifier supplied from a three phase (or single phase) AC source. Figure 3.2 shows a six step voltage source inverter. The inductor L is inserted to limit shot through fault current. A large electrolytic capacitor C is inserted to stiffen the DC link voltage. The switching

devices in the voltage source inverter bridge must be capable of being turned off and on. Insulated gate bipolar transistors (IGBT) are used because they have this ability in addition; they offer high switching speed with enough power rating. Each IGBT has an inverse parallel-connected diode. This diode provide alternate path for the motor current after the IGBT, is turned off [23].

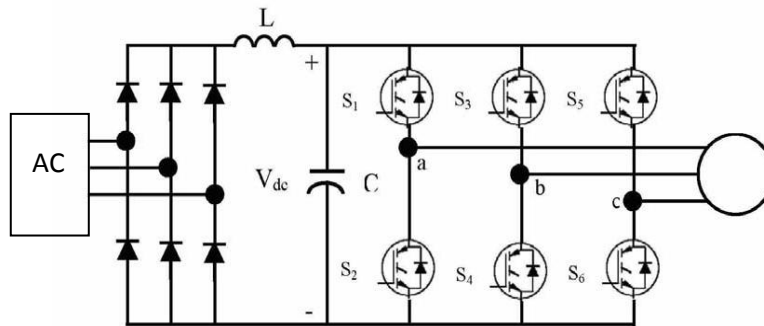


Figure 3.2 Voltage Source Inverter

Each leg of the inverter has two switches one connected to the high state (+) of the DC link and the other to the low state (-), only either one of the two can be on at any instant. When the high state gate signal is on the phase is assigned the binary number 1, and assigned the binary number 0 when the low state gate signal is on. Considering the combinations of status as above of three phases a, b and c the inverter has eight switching modes ($V_a V_b V_c = 000-111$) two are zero voltage vectors V_0 (000) and V_7 (111) where the motor terminals gets short circuited and the others are non zero voltage vectors V_1 to V_6 , which provides the motor the required voltage.

The dq model for the voltage source inverter in the stationary reference frame is obtained by applying the dq transformation equation for the inverter switching modes. The six nonzero voltages space vectors will have the orientation shown in figure 3.3, which also shows the possible dynamic locus of the stator flux and its different variation depending on the VSI states. The possible global locus can be divided into six different sectors feed through the discontinuous line. Each vector lies in the centre of a sector of 60° width named S1 to S6 according to the voltage vector it contains.

From Equation 3.5 it can be seen that the inverter voltage directly reflect the stator flux, the required stator flux locus will be obtained by choosing the appropriate inverter switching state. Thus the stator flux linkage move in space in the direction of the stator

voltage space vector at a speed that is proportional to the magnitude of the stator voltage space vector. By selecting step by step the appropriate stator voltage vector, it is then possible to change the stator flux in the required way. If an increase of the torque is required then the torque is controlled by applying voltage vectors that advance the flux linkage space vector in the direction of rotation. If a decrease in torque is required then zero switching vector is applied, the zero vector that minimize inverter switching is selected. In summary if the stator flux vector lies in the k-th sector and the motor is running anticlockwise

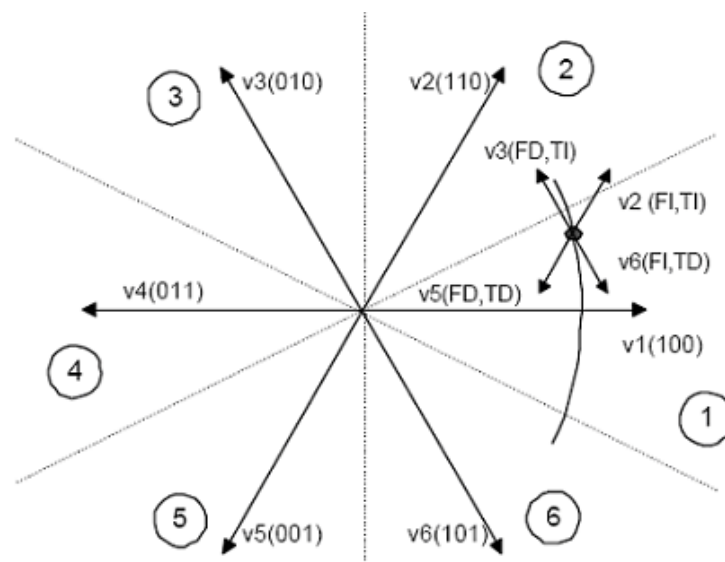


Figure 3.3 Stator flux vector locus and different possible switching voltage vectors, FD: Flux Decrease, FI: Flux Increase, TD: Torque Decrease, TI: Torque Increase

then torque can be increased by applying stator voltage vectors V_{k+1} or V_{k+2} , and decreased by applying a zero voltage vector V_0 or V_7 . Decoupled control of the torque and stator flux can be achieved by acting on the radial and tangential components of the stator voltage space vector in the same directions, and thus can be controlled by the appropriate inverter switching. In general, when the stator flux linkage vector lies in the k-th sector its magnitude can be increased by using switching vectors V_{k-1} (for clockwise rotation) or V_{k+1} (for anticlockwise rotation), and can be decreased by applying voltage vectors V_{k-2} (for clockwise rotation) or V_{k+2} (for anticlockwise rotation). With the help of the sector division as depicted in figure 3.3, the general table 3.1 can be formulated. It can be seen from table 3.1, that the states V_k and V_{k+3} , are not considered in the torque

because they can both increase (first 30 degree) or decrease (second 30 degree) the torque at the same sector depending on the stator flux position.

VOLTAGE VECTOR	INCREASE	DECREASE
Stator Flux	V_k, V_{k+1}, V_{k-1}	$V_{k+2}, V_{k-2}, V_{k+3}$
Torque	V_{k+1}, V_{k+2}	V_{k-1}, V_{k-2}

Table3.1. General Selection Table for Direct Torque Control, "k" being the sector number.

This can be tabulated as the look-up Table 3.2 Finally, the DTC conventional look up table can be formed as follows:

Flux Error D_{ψ}	Torque Error $d\tau$	S1	S2	S3	S4	S5	S6
1	1	V_2	V_3	V_4	V_5	V_6	V_1
	0	V_0	V_7	V_0	V_7	V_0	V_7
	-1	V_6	V_1	V_2	V_3	V_4	V_5
-1	1	V_3	V_4	V_5	V_6	V_1	V_2
	0	V_0	V_7	V_0	V_7	V_0	V_7
	-1	V_5	V_6	V_1	V_2	V_3	V_4

Table3.2 Conventional DTC look up table

3.3 DTC Schematic

In figure 1.2 a schematic diagram of Direct Torque Control is shown. As it can be seen, there are two different loops corresponding to the magnitudes of the stator flux and torque. The reference values for the flux stator modulus and the torque are compared with the actual values, and the resulting error values are fed into the two level and three-level hysteresis blocks respectively. The outputs of the stator flux error and torque error hysteresis blocks, together with the position of the stator flux are used as inputs of the

look up table (see table 3.2). The inputs to the look up table are given in terms of +1,0,-1 depend on whether the torque and flux errors within or beyond hysteresis bands and the sector number in which the flux sector is present at that particular instant. In accordance with the figure 1.2, the modulus of stator flux and torque errors are required to be restricted within their respective hysteresis bands. From the schematic of DTC it is cleared that, in order to have proper selection of voltage sector from lookup table, the DTC scheme requires the flux and torque estimations.

3.3.1 Methods for Estimation of Stator Flux in DTC

Accurate flux estimation in Direct Torque controlled induction motor drives is important to ensure proper drive operation and for its stability during the operation. Most of the flux estimation techniques proposed is based on voltage model, current model, or the combination of both [26]. The estimation based on current model normally applied at low frequency, and it requires the knowledge of the stator current and rotor mechanical speed or position. In some industrial applications, the use of incremental encoder to get the speed or position of the rotor is undesirable since it reduces the robustness and reliability of the drive. It has been widely known that even though the current model has managed to eliminate the sensitivity to the stator resistance variation. The use of rotor parameters in the estimation introduced error at high rotor speed because of the rotor parameter variations. So in this resent DTC control scheme the flux and torque are estimated by using voltage model described by equations 3.5- 3.6, which does not need a position sensor and the only motor parameter used is the stator resistance.

$$\psi_{sD} = \frac{L_x}{L_m} + \psi_{rd} \frac{L_m}{L_r} \quad \text{----- (3.7)}$$

$$\psi_{sQ} = \int (V_{sQ} - R_S I_{sQ}) dt = i_{sQ} \frac{L_x}{L_m} + \psi_{rq} \frac{L_m}{L_r} \quad \text{----- (3.8)}$$

where $L_x = L_s L_r - L_m^2$ and i_{sD} , i_{sQ} are calculated from equations and subsequently the torque can be estimated by

$$t_e = \frac{3}{2} P (\psi_{sD} i_{sQ} - \psi_{sQ} i_{sD}) \quad \text{----- (3.9)}$$

By using these estimated values from voltage model we proceed to the lookup table to select the proper voltage vector the voltage model gives accurate estimation at high speeds however, at low speed, some problems arise. In practical implementation, even a

small DC off-set present in the back emf due to noise or measurement error inherently present in the current sensor, can cause the integrator to saturate [22].

3.3.2 Direct flux control

In stationary reference frame the stator flux equation can be written as:

$$\Psi_s = \int V_s - i_s dt \quad \text{----- (3.10)}$$

If the stator voltage drop is neglected applied voltage vector and the equation will be reduced to

$$\Delta\Psi_s = V_s \Delta \quad \text{----- (3.11)}$$

which means, by applying stator voltage vector V_s for a time increment Δt , Ψ_s can be changed incrementally. The command value of the stator flux vector Ψ_s^* follows a circular trajectory, the plane of stator flux is divided into six sectors as shown in figure 3.3. Each sector has a different set of voltage vector to increase or decrease the stator flux. The command flux vector rotates in anticlockwise direction in a circular path and the actual stator flux vector Ψ_s tracks the command flux in a zigzag path but constrained to the hysteresis band which is shown in figure 3.4.

In general the active forward voltage vectors ($V_{s,k+1}$ and $V_{s,k+2}$) are applied to increase or decrease the stator flux respectively when the stator flux lies in sector k . The radial voltage vectors ($V_{s,k}$ and $V_{s,k+3}$) which quickly affect the flux are generally avoided. The active reverse voltage vectors ($V_{s,k-1}$ and $V_{s,k-2}$) are used to increase or decrease the stator flux in reverse direction.

The stator flux vector change due to stator voltage vector is quick whereas change rotor flux is sluggish because of its large time constant T_r . That is why Ψ_s movement is jerky and Ψ_r moves uniformly at frequency ω_e as it is more filtered. However the average speed of both remains the same in steady state condition.

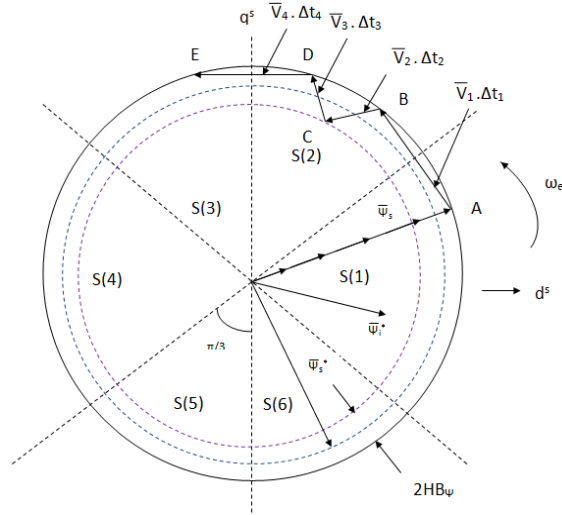


Figure.3.4 Circular trajectory of stator flux

3.3.3 Torque Hysteresis Controller

The Torque hysteresis is controller which is a three level controller which means the torque control loop has three levels of digital outputs. The torque error ΔT_e is given to the torque hysteresis controller and the output is torque error status (dT_e) which can have three values -1, 0 or 1. The width of the hysteresis band is $2\Delta T_e$. Torque error status is given to the switching table for optimum voltage vector selection for the inverter. Figure 3.5 shows the hysteresis torque controller.

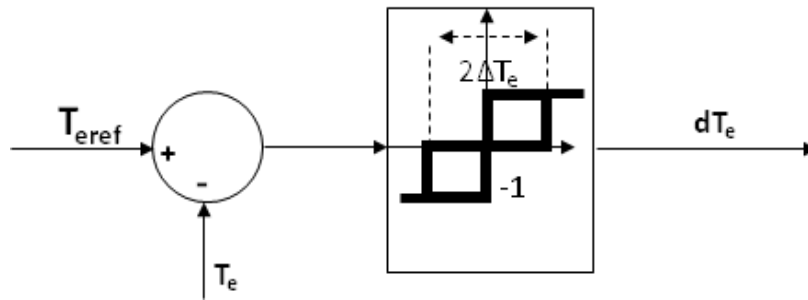


Figure 3.5 Torque hysteresis controller

Torque error is given by

$$\Delta T_e = T_{e \text{ ref}} - T_e \quad \text{----- (3.12)}$$

$|dT_e| = 1$ if $|T_e| < |T_{e \text{ ref}}| - |\Delta T_e|$: Torque to be increased

$|dT_e| = -1$ if $|T_e| > |T_{e \text{ ref}}| + |\Delta T_e|$: Torque to be decreased

$|dT_e| = 0$ if $|T_{e \text{ ref}}| - |\Delta T_e| \leq |T_e| \leq |T_{e \text{ ref}}| + |\Delta T_e|$: Torque to remain unchanged.

The objective is to limit the torque error to the hysteresis band limits for direct torque and flux control.

CHAPTER 4

SPACE VECTOR PULSE WIDTH MODULATION FED INDUCTION MOTOR DRIVES

4.1 Introduction

The use of PWM drive is advantageous in many ways, for example it obtains its dc input through uncontrolled rectification of commercial AC mains and has good power factor, good efficiency, relatively free from regulation problems, it has the ability to operate the motor with nearly sinusoidal current waveform. The conventional PWM techniques are suitable for open loop control, for the implementation of a closed loop controlled AC drive Space vector PWM (SVPWM) technique is applied. In this technique, the switching patterns for the bridge inverter are generated from the knowledge of stator voltage space phasor. A reference voltage vector is generated to generate a field synchronous with the rotating voltage vector by utilizing the different switching states of a three phase bridge inverter

4.2 Theory of Space vector pulse width modulation

When three phase supply is given to the stator of the induction machine, a three phase rotating magnetic field is produced. Due to this field flux, a three phase rotating voltage vectors are generated which lags the flux by 90° . This field can also be visualised by a logical combination of the inverter switching state which is the basic concept of SVPWM.

4.2.1 Realization of voltage space phasor

The three phase bridge inverter has eight possible switching states, six active and two zero states. The six switches (IGBT/Thyristors) have a well-defined state of ON or OFF for each configurations. At a particular instant, only one switch in each of the three legs is ON. Corresponding to each ON/OFF state of the inverter, there is one voltage space vector. For example for state zero it is V_0 , for state 1 it is V_1 and so on. These switching state vectors have equal magnitude but 60° apart from each other. These vectors can be written in generalized form as follows:

$$V_k = V_{dc} e^{j(k-1/3)\pi} \quad k = 1, 2, \dots, 6 \quad \text{----- (4.1)}$$

$$= 0 \quad k = 0, 7$$

Where k = inverter state number.

V_{dc} = dc link voltage of the inverter

The inverter state vectors can be drawn as shown in figure 4.1

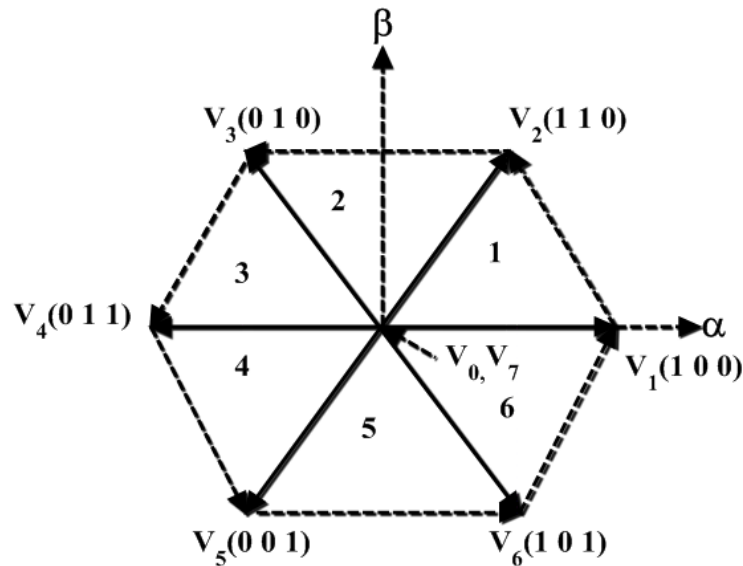


Figure 4.1 Inverter switching state vectors

The space bounded by two inverter space vectors is called a sector. So the plane is divided into six sectors each spanning 60° . In a balanced three phase system the voltage vectors are 120° apart in space and are represented by rotating vectors, whose projections on the fixed three phase axes are, sinusoidal in nature. So they can be represented as three sinusoidal references obtained by a voltage reference space vector V_{ref}^* or V_s^* . The reference vector is assumed to be rotating in counter-clockwise direction with respect to d-axis (α -axis) as shown in figure 4.2 through six sectors [27].

The reference space vector can be synthesized by a combination of eight state vectors and is constant in magnitude at switching instant t_s in case the switching frequency much higher than the output frequency. In a time average sense the reference vector at that instant can be approximated by two active voltage states of the inverter. For only certain amount of time these states are valid.

$$V_s^* = V_k t_k + V_{k+1} t_{k+1} \quad k = 0, 1, 2, \dots, 7 \quad \text{----- (4.2)}$$

In SVPWM, it is assumed that the s space phasor of stator voltage V_s^* , is moving in α - β plane with constant angular velocity describing approximately a circular path. The basis of SVPWM scheme is to sample the V_s^* at sufficiently high rate, in between the sampling instants the vector is assumed to be constant in magnitude as shown in figure 4.2 [28][29]

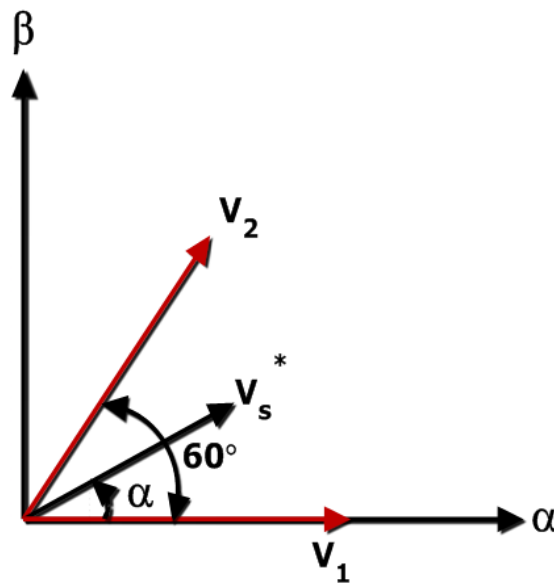


Figure 4.2 Reference vector in sector 1

4.2.2 Pulse pattern generation

The PWM pattern generation means the generation of gating pulses for the six switches (IGBT/Thyristors) of the inverter, for correct intervals so that appropriate switching state vectors are active for the required time intervals as the reference space vector moves over a full cycle.

In order to obtain minimum switching frequency, it is desired that only one phase of the inverter changes state from $+\frac{V_{dc}}{2}$ to $-\frac{V_{dc}}{2}$ while changing the switching vectors. So the arrangement of the switching sequence should be such that the transition from one state to the next state is performed by switching only one inverter phase. This is done by sequential switching the inverter legs starting from one zero state and ending at another zero state. Figure 4.3 shows the optimum inverter phase to dc center tap voltages. It can be noted that the switching frequency of the inverter is half of the sampling frequency

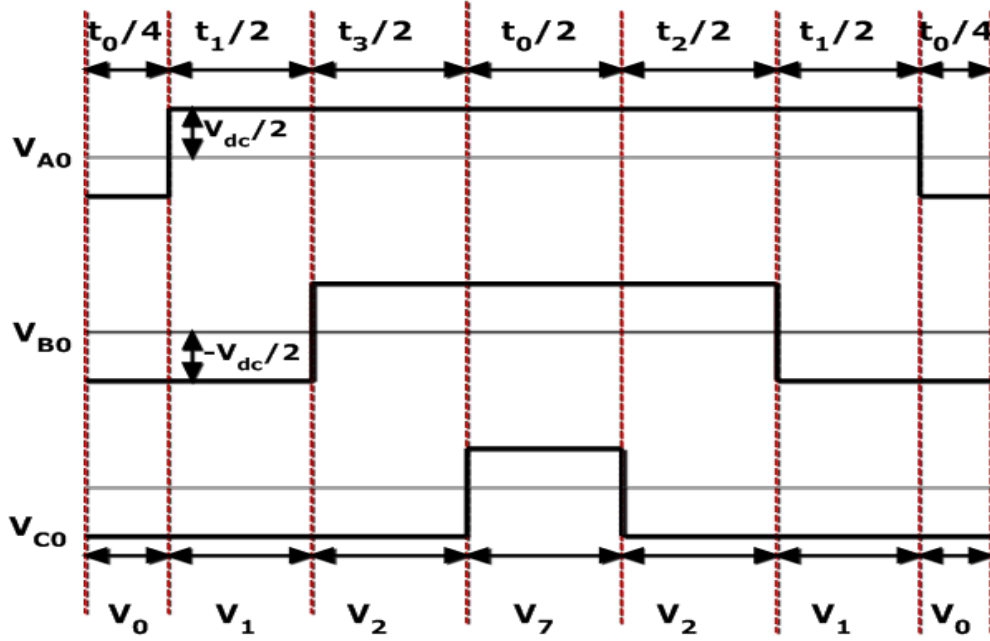


Figure 4.3 Leg voltages and space vector disposition in sector 1

The mean values of the phase to center tap voltages (V_{Ao} , V_{Bo} , V_{Co}) can be evaluated, averaging over one sampling period t_s as follows:

$$V_{Ao} = \frac{V_{dc}}{2t_s} \left(-\frac{t_0}{2} + t_1 + t_2 + \frac{t_0}{2} \right) \quad \text{----- (4.3)}$$

$$V_{Bo} = \frac{V_{dc}}{2t_s} \left(-\frac{t_0}{2} - t_1 + t_2 + \frac{t_0}{2} \right) \quad \text{----- (4.4)}$$

$$V_{Co} = \frac{V_{dc}}{2t_s} \left(-\frac{t_0}{2} - t_1 - t_2 + \frac{t_0}{2} \right) \quad \text{----- (4.5)}$$

Substituting the values of t_1 , t_2 in above three equations we get,

$$V_{Ao} = a \frac{V_{dc}}{\sqrt{3}} \sin \left(\gamma + \frac{\pi}{3} \right) \quad \text{----- (4.6)}$$

$$V_{Bo} = a \frac{V_{dc}}{\sqrt{3}} \sin \left(\gamma - \frac{\pi}{6} \right) \quad \text{----- (4.7)}$$

$$V_{Co} = -V_{Ao} \quad \text{----- (4.8)}$$

CHAPTER 5

MODEL SIMULATION AND RESULTS

5.1 Improvement by Sector Change

The conventional DTC utilizes the sector 1 from -30 deg to + 30 deg and the switching table for the conventional DTC is given in the look up table (Table 3.2) and the corresponding simulation model is given in figure 5.1.

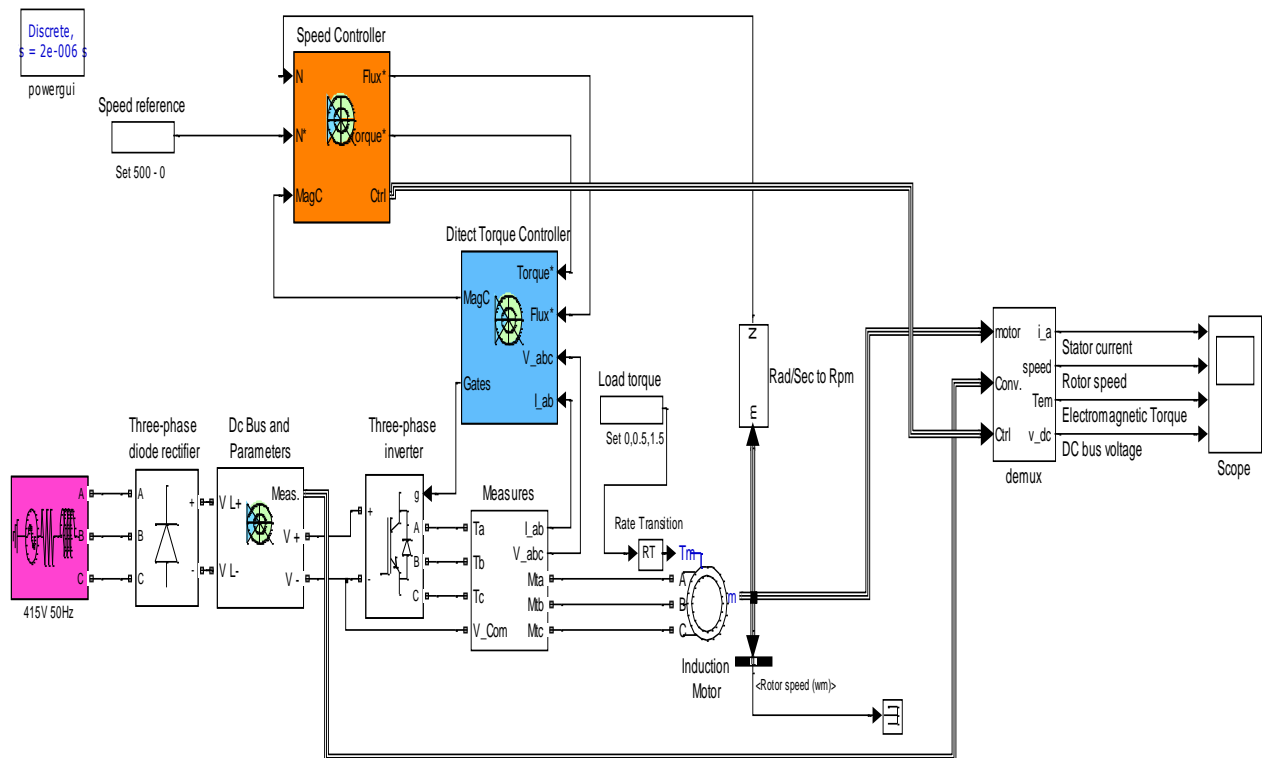


Figure 5.1 Simulink model for Conventional DTC

Now considering the Sector1 from 0 deg to 60 deg instead of -30 deg to + 30 deg and the corresponding look up table developed for all the six sectors are given in table 5.1

Flux Error $\frac{d\psi}{dt}$	Torque Error $d\tau$	S1	S2	S3	S4	S5	S6
1	1	V_2	V_3	V_4	V_5	V_6	V_1
	0	V_7	V_0	V_7	V_0	V_7	V_0
	-1	V_1	V_2	V_3	V_4	V_5	V_6
-1	1	V_4	V_5	V_6	V_1	V_2	V_3
	0	V_0	V_7	V_0	V_7	V_0	V_7
	-1	V_5	V_6	V_1	V_2	V_3	V_4

Table 5.1 Look Up table for Modified DTC

After obtaining the look up table the change in the voltage vectors for flux error and torque error is given in figure 5.2 and corresponding Simulink model for modified DTC is shown in figure 5.3

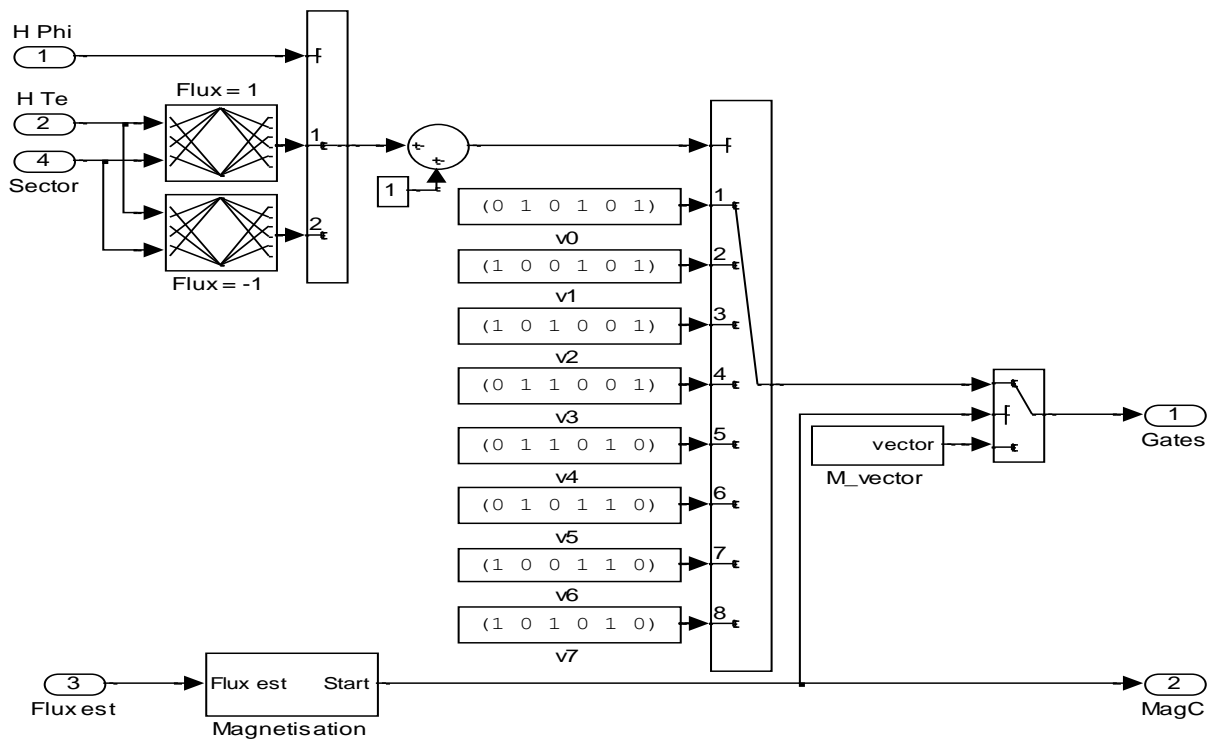


Figure 5.2 Implementation of Flux and Torque Error through Voltage Vectors

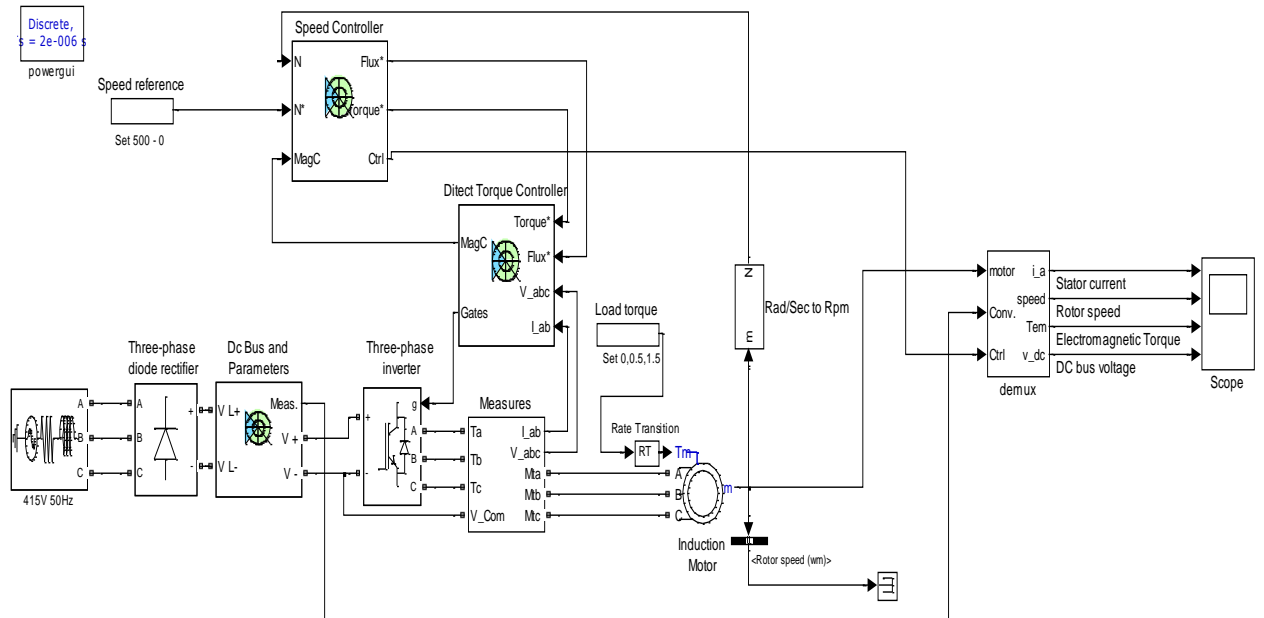


Figure 5.3 Simulink Model for Modified DTC

5.2 Improvement through 12 Sector SVPWM DTC

Further, the model was developed for Space Vector Pulse Width Modulation (SVPWM) with 12 sectors starting from 0 deg to 30 deg, 30 deg to 60 deg, 60 deg to 90 deg, 90 deg to 120 deg, 120 deg to 150 deg, 150 deg to 180 deg, 180 deg to -150deg, -150 deg to -120 deg, -120 deg to -90 deg, -90 deg to -60 deg, -60 deg to -30 deg, -30 deg to 0 deg[28]. The corresponding voltage vectors are shown in figure 5.4

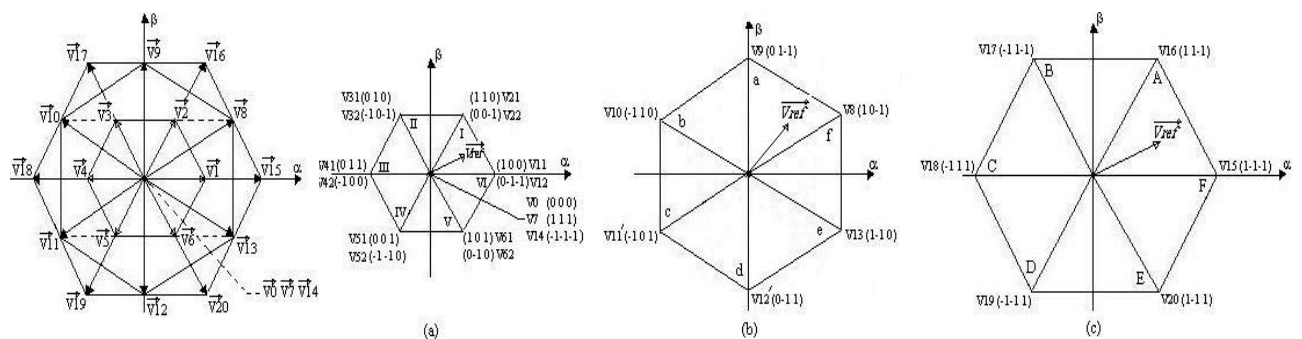


Figure 5.4 Three-level voltage inverter vectors' hexagons. (a). Small hexagon. (b). Middle hexagon. (c). Big hexagon.

The switching table for 24 active voltage vectors and 3 zero voltage vectors are calculated and the 12 sectors are implemented through the Simulink model given in figure 5.5 and the Simulink model for the SVPWM is given in figure 5.6.

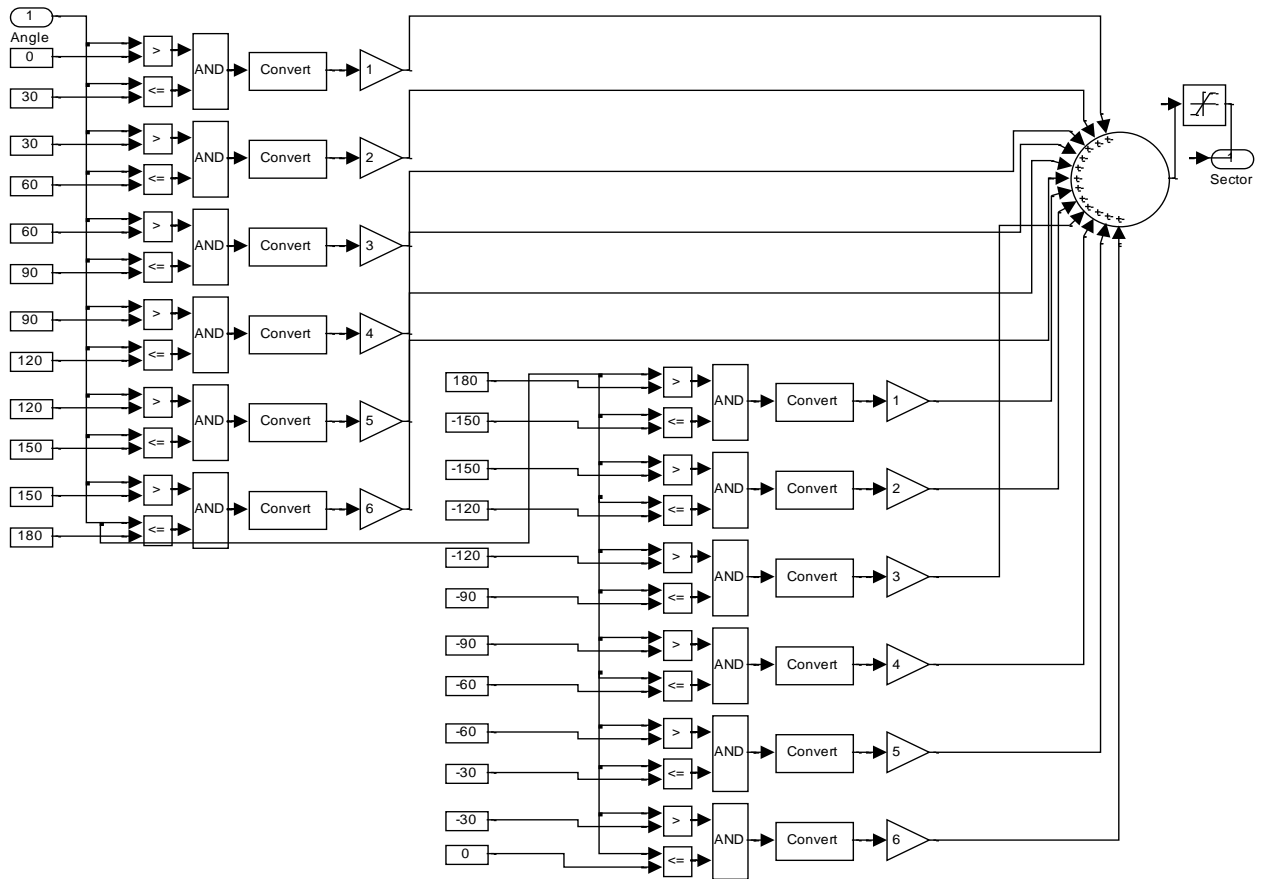


Figure 5.5 Implementation of 12 Sectors in the SVPWM DTC Simulink Model

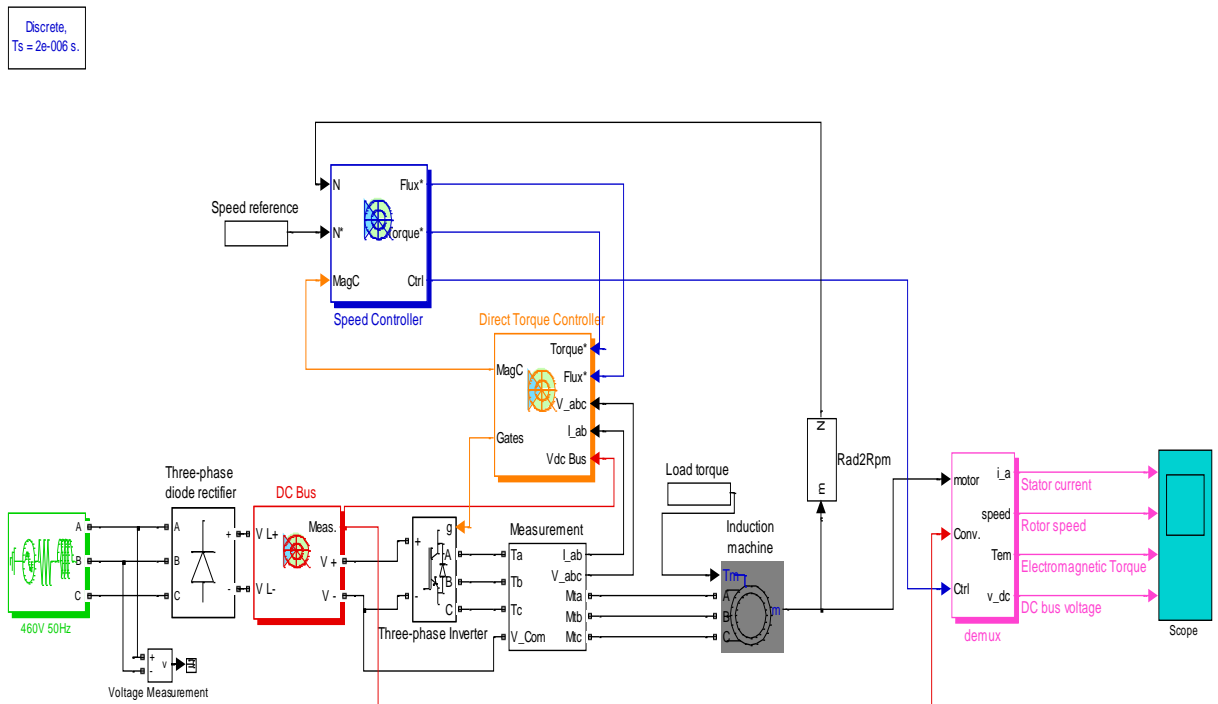


Figure 5.6 The Simulink Model for 12 Sectors SVPWM based DTC

5.3 Induction Motor Parameters

A squirrel cage induction motor is considered in stationary reference frame and has the following parameters.

Power rating (P)	:	3 Kw
Power Line Voltage (V_{L-L})	:	460 Volts
Frequency	:	50 Hz
Stator Resistance (R_s)	:	0.73933 Ohm
Stator Inductance (L_s)	:	0.01507 Henry
Rotor Resistance (R_r)	:	0.46227 Ohm
Rotor Inductance (L_r)	:	0.01507 Henry
Mutual Inductance (L_m)	:	0.52077 Henry
Inertia (J)	:	3.1 Kg-m ²
Friction Factor	:	0.08
No. of Poles	:	4

5.4 Results for Conventional DTC

The Stator Current (i_a), Torque (τ) and Speed (ω) are plotted against the time with the reference speed of 500 rpm at time ($t = 0$) and 0 rpm at time ($t = 1$). The load torque is at 0, 792 Nm, -792 Nm for time (t) 0, 0.5 and 1.5 Sec respectively. The simulation is carried out for 2 Sec and the results are obtained in figure 5.7, figure 5.8 and figure 5.9.

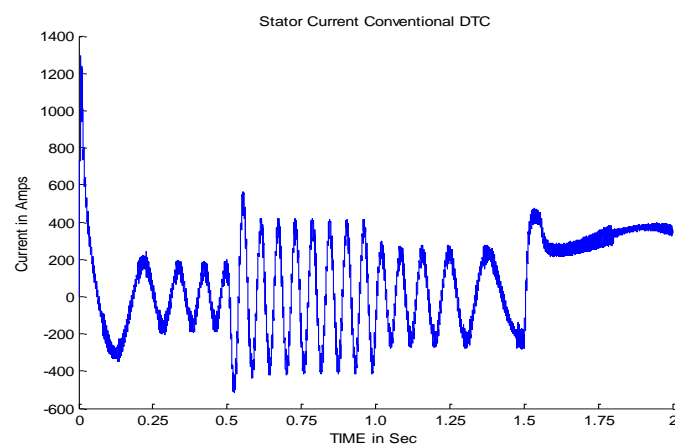


Figure 5.7 Stator Current Characteristics of Conventional DTC

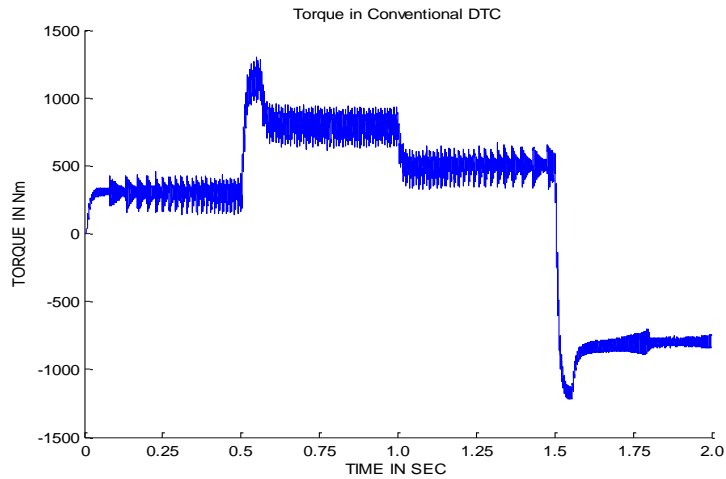


Figure 5.8 Torque Characteristics of Conventional DTC

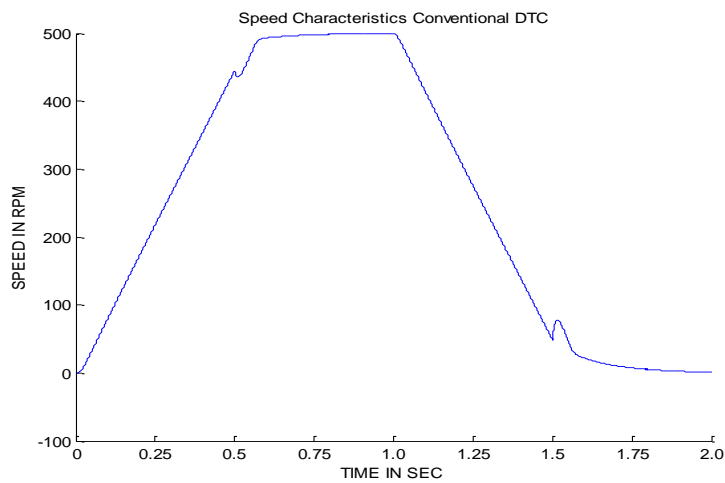


Figure 5.9 Rotor Speed Characteristics of Conventional DTC

5.5 Results for Modified DTC

The Stator Current (i_a), Torque (τ) and Speed (ω) are plotted against the time with the reference speed of 500 rpm at time ($t = 0$) and 0 rpm at time ($t = 1$). The load torque is at 0, 792 Nm, -792 Nm for time (t) 0, 0.5 and 1.5 Sec respectively. The simulation is

carried out for 2 Sec and the results are obtained in figure 5.10, figure 5.11 and figure 5.12.

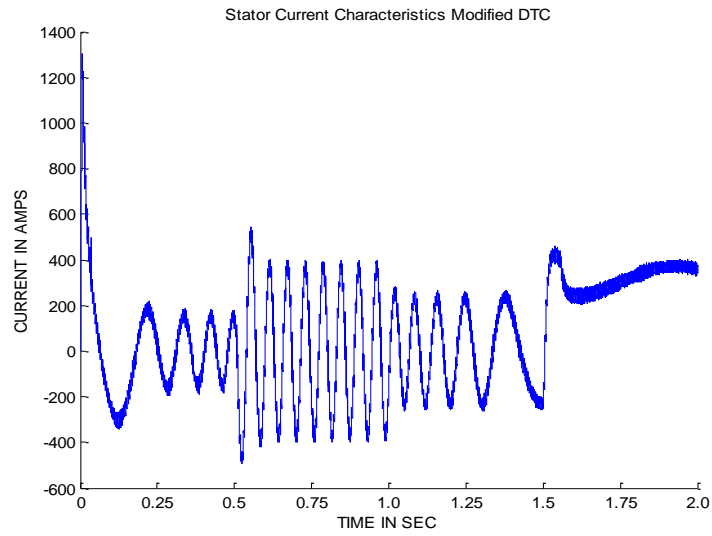


Figure 5.10 Stator Current Characteristics of Modified DTC

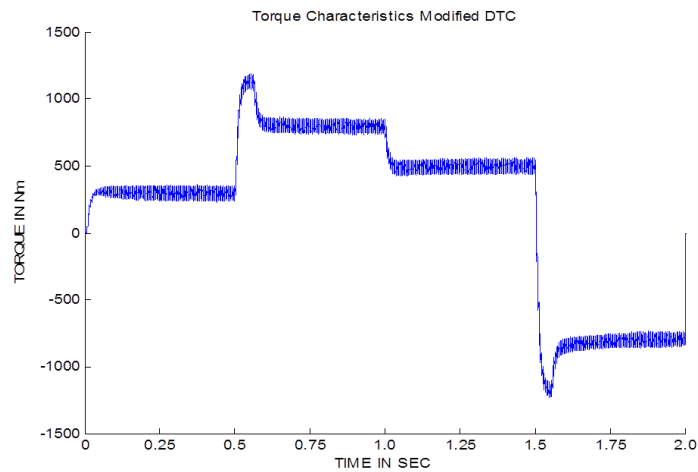


Figure 5.11 Torque Characteristics of Modified DTC

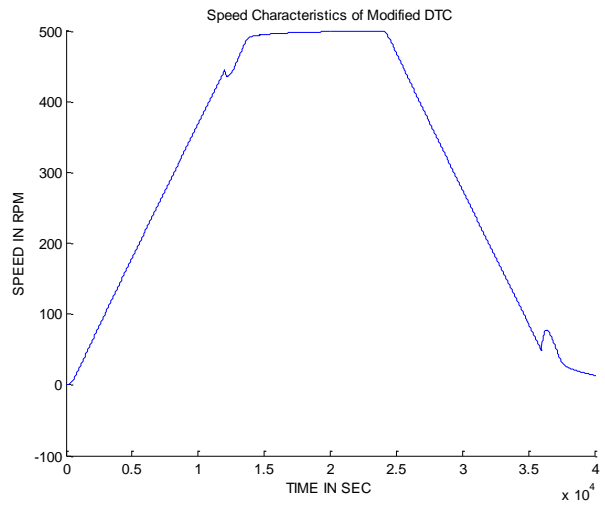


Figure 5.12 Rotor Speed Characteristics of Conventional DTC

5.6 Results for 12 Sectors SVPWM DTC

The stator current (i_a), torque (τ) and speed (ω) are plotted against the time with the reference speed of 500 rpm at time ($t = 0$) and 0 rpm at time ($t = 1$). The load torque is at 0, 792 Nm, -792 Nm for time (t) 0, 0.5 and 1.5 Sec respectively. The simulation is carried out for 2 Sec and the results are obtained in figure 5.13, figure 5.14 and figure 5.15.

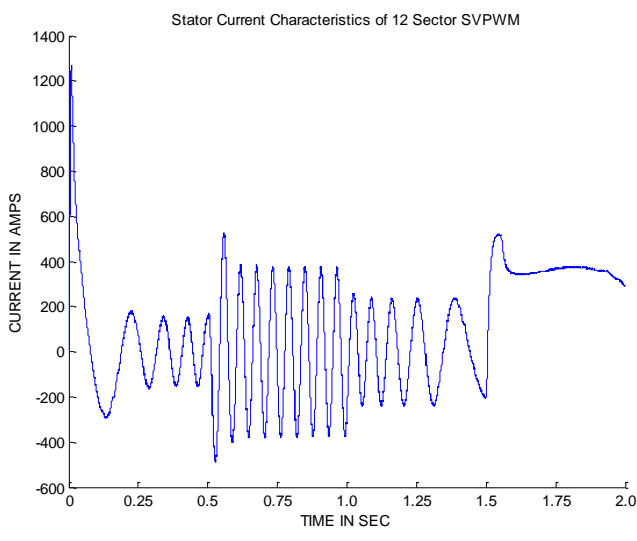


Figure 5.13 Stator Current Characteristics of 12 Sectors SVPWM DTC

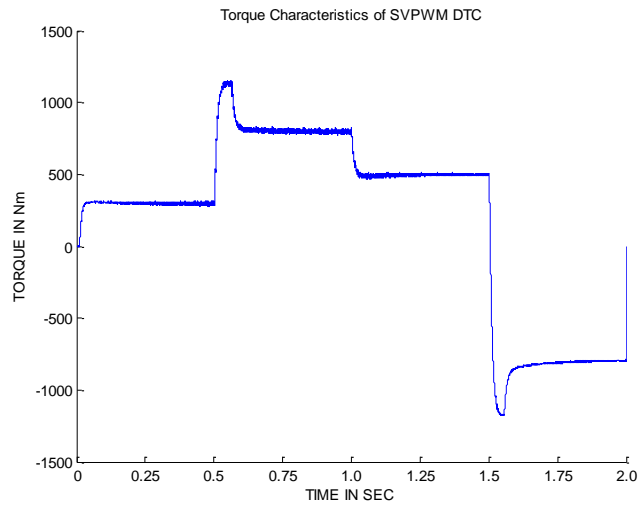


Figure 5.14 Torque Characteristics of 12 Sectors SVPWM DTC

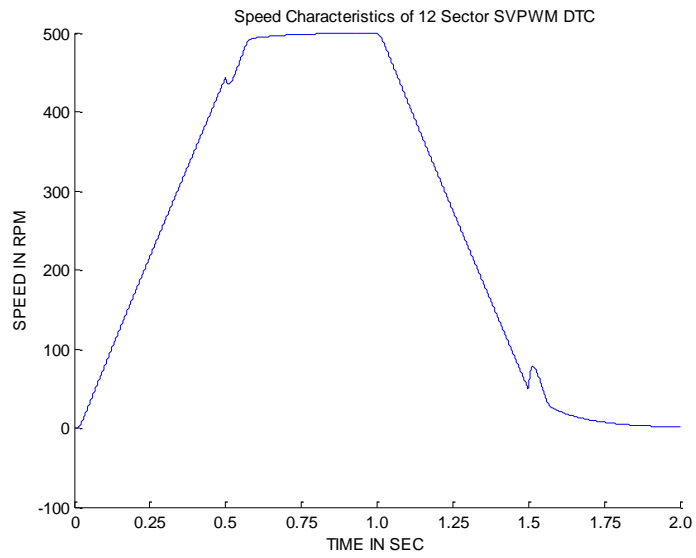


Figure 5.15 Rotor Speed Characteristics of 12 Sectors SVPWM DTC

CHAPTER 6

COMPARISON AND CONCLUSIONS

6.1 Comparison

The comparison is made between the above three different techniques of DTC for improvement in its performance. The comparison is done for stator current and torque characteristics between Conventional DTC, Modified DTC and 12 Sectors SVPWM DTC and the results are shown in figure 6.1 and figure 6.2 respectively.

No significant change is observed in the Speed characteristics among the three methods. However significant improvement in Stator Current and Torque ripples were observed in both the methods of improvement as mentioned in 5.1 and 5.2 .

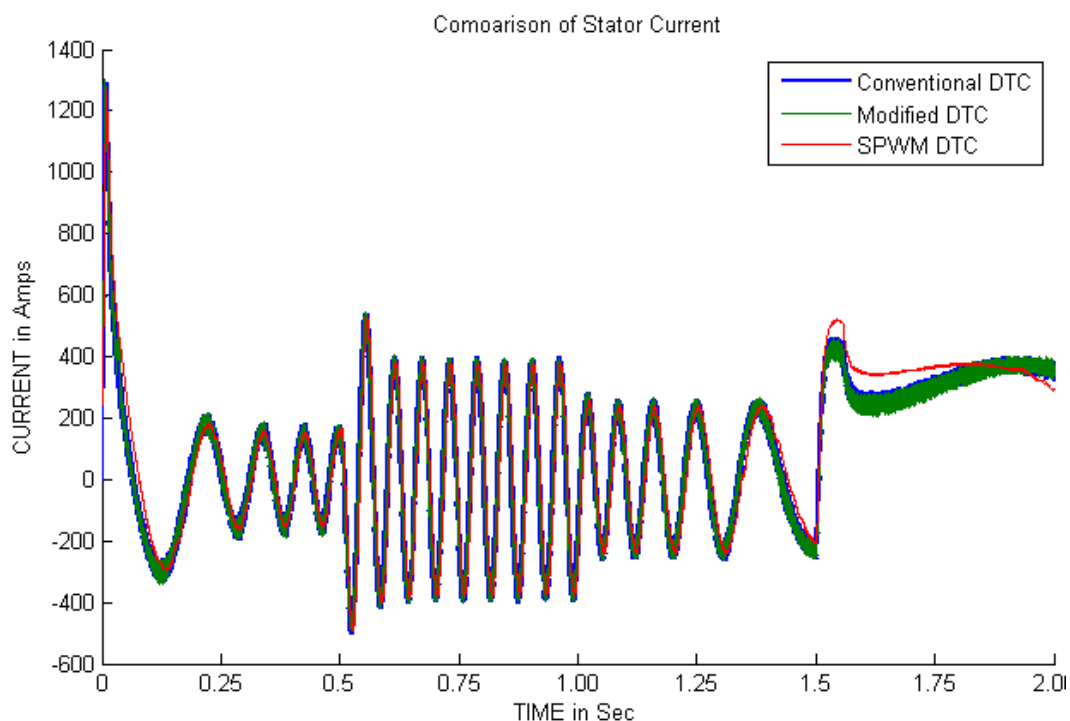


Figure 6.1 Comparison of Stator Current Characteristics

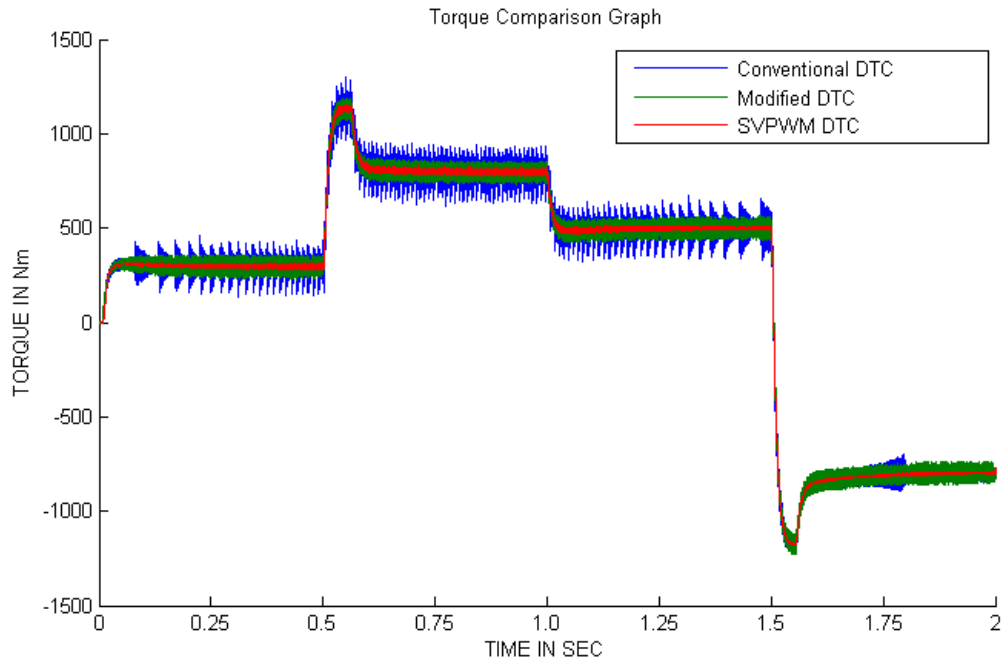


Figure 6.2 Comparison of Torque Characteristics

6.2 Conclusion

From the above comparison it may be concluded that there is considerable improvement in ripples in the torque as well as stator current of Modified and SVPWM DTC Techniques and SVPWM DTC is considered to be the best method in terms of ripple reduction in stator current and torque as compared to the conventional DTC.

6.3 Future Scope

The 12 Sector division is being simulated in 2 level inverter, it may be employed for three level inverter and also for different types of multi level inverter in the future scope of work.

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