

Libyan Academy



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Department of Electrical and Computer Engineering**

**Simulation of Dynamic and Transient Performance of
Permanent Magnet Motor**

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بِسْمِ اللَّهِ الرَّحْمَنِ الرَّحِيمِ

إِنَّمَا أَمْرُهُ إِذَا أَرَادَ شَيْئًا أَنْ يَقُولَ لَهُ كُنْ فَيَكُونُ ﴿٨١﴾ فَسُبْحَانَ الَّذِي يَبْدَأُ الْمَلَكُوتَ

كُلِّ شَيْءٍ وَإِلَيْهِ تُرْجَعُونَ ﴿٨٢﴾

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CONTENTS

Acknowledgment.....	III
Abstract.....	IV
List of Tables	VI
List of Figures.....	VII
Symbols and abbreviations.....	XI

CHAPTER 1

1. INTRODUCTION.....	2
1.1. Motivation.....	2
1.2. Objectives of the Thesis.....	2
1.3. Outline of the Present Work	3
1.4. Previous Work	9

CHAPTER 2

2. DESCRIPTION OF THE DRIVE SYSTEM OF PM MOTOR.....	12
2.1. Permanent Magnet Motor Drive System.....	12
2.2. Position Sensor.....	12
2.3. Current Controlled Inverter.....	16
2.4. Permanent Magnet Motor.....	22
2.5. PM Motor Speed Control.....	36
2.6. Data of the Motor under Study.....	42

CHAPTER 3

3. MODELLING, SIMULATION AND CONTROL OF BLDC MOTOR.....	45
3.1. Dynamics of the BLDC Motor.....	45
3.2. BLDC Motor Performance in Steady-State.....	54
3.3. BLDC Motor Speed Control.....	62

CHAPTER 4

4. MODELLING, SIMULATION AND CONTROL OF PMSM.....	66
4.1. Dynamics of the Synchronous Motor.....	66
4.2. Steady-state Model of the synchronous Motor.....	72

4.3. Permanent Magnet Synchronous Motor Control.....	79
--	----

CHAPTER 5

5. SIMULATION RESULTS OF PM MOTOR.....	95
---	-----------

5.1. Simulation Results of Permanent Magnet BLDC motor.....	95
---	----

5.2. Simulation Results of PMLDC motor with PI Controller.....	99
--	----

5.3. Simulation Results of PMSM.....	100
--------------------------------------	-----

5.4. Simulation Results of PMSM with PI Controller.....	104
---	-----

CHAPTER 6

6. CONCLUSION.....	106
---------------------------	------------

 REFERENCES.....	109
---	-----

 APPENDIX A: M-FILES FOR BLDC MOTOR OPERATION.....	112
---	-----

 APPENDIX B: M-FILES FOR SYNCHRONOUS MOTOR OPERATION.....	113
---	-----

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ABSTRACT

The performance of a three-phase permanent magnet (PM) motor operating as a synchronous motor (PMSM) and brushless DC (BLDC) motor is discussed. The PM motor, when operating as a synchronous machine, is supplied with constant frequency, meaning constant rotor speed operation. When operating as a BLDC motor, the supply frequency changes according to the actual rotor speed. It means that the speed signal is fed back to the controller, which generates the appropriate frequency to supply the stator winding. This means that the BLDC motor cannot operate without a position sensor.

The objective of the project was to analyze and compare the performance of the motor operating as a BLDC and synchronous motor. To do this, mathematical models for both operation modes (that allowed us to analyze the performance in dynamic conditions) were proposed. To analyze the motors performance in steady-state conditions, simpler motor models were proposed and calculations carried out. The results obtained from the two types of models were compared.

The motor was supplied from a 3-phase inverter (controlled by PI controller) which operated in MATLAB/SIMULINK environment . This allowed us to control the motor speed in both types of models (BLDC and PMSM), and the results obtained from the two types of models were compared.

The important concluded results were that the (BLDC) motor is faster than (PMSM) in reaching steady state. Regarding speed and efficiency the (PMSM) was the best if one's application requires maintaining a constant speed on their motor. When the motor works with a control circuit (PI controller), the synchronous motor's behavior is better than that of BLDC motor behavior.

الخلاصة

درسنا طريقة عمل المحرك ثلاثي الطور ذي المغناطيس الثابت PM الذي يعمل كمحرك تزامني (PMSM) ومحرك تيار مستمر بدون فرش (BLDC). عندما يعمل المحرك كمحرك تزامني فإنه يغذى بمصدر ثابت التردد، بمعنى أنه يعمل بسرعة ثابتة، عندما يعمل المحرك كمحرك تيار مستمر (BLDC) فإن تردد المصدر يتغير تبعاً لسرعة الدواران الحقيقية، هذا يعني إن إشارة السرعة تغذى ارتجاعياً للمتحكم (controller) الذي يولد التردد المناسب للملف الثابت. هذا يعني أن محرك التيار المستمر لا يمكن أن يعمل بدون متحسس الموضع.

الغرض من هذا المشروع هو تحليل و مقارنة عمل المحرك عندما يعمل كمحرك تيار مستمر (BLDC) وكمحرك تزامني (PMSM). وللقيام بذلك استخدمنا النموذج الرياضي الذي يمثل كل محرك (هذا يمكننا من تحليل عملها أو تصرفها في حالة الحركة).

لدراسة عمل المحركات في حالة الثبات فإننا استخدمنا النموذج البسيط لإجراء الحسابات لكلا المحركين وقارنا بين النتائج المتحصل عليها.

يغذى المحرك بمحول ثلاثي الطور (يتحكم فيه باستخدام المتحكم PI) يعمل في بيئة MATLAB/SIMULINK، هذا يمكننا من التحكم في سرعة المحرك في كلا النموذجين (BLDC and PMSM) و مقارنة النتيجة المتحصل عليها في كلا المحركين.

من أهم النتائج المستخلصة هي أن المحرك (BLDC) أسرع في الوصول إلى حالة الاستقرار من (PMSM)، فيما يتعلق بالسرعة و الكفاءة فإن (PMSM) هو الأفضل خاصة في التطبيقات التي تحتاج ثبوت في السرعة، كذلك مع وجود المتحكم (PI) فإن (PMSM) أفضل استجابة من (BLDC).

LIST OF TABLES

Table 2.1 Device Power and Switching Capabilities	17
Table 2.2. Switching sequence.....	33
Table 2.3 A summary of the Differences between BLDC Motor and PMSM.....	36
Table 2.4 Recommended PID value setting.....	41
Table 2.5 Data of the motor under study.....	42
Table 3.1 Electromechanical parameters of BLDC motor in steady-state.....	53
Table 4.1 Electromechanical parameters of synchronous motor performed at rated conditions.....	71
Table 4.2 Parameters of the equivalent circuit.....	77

LIST OF FIGURES

Fig.2.1 Drive System Schematic.....	12
Fig.2.2 Optical Encoder.....	13
Fig.2.3 Quadrature Encoder Channels.....	14
Fig.2.4 Absolute Encoder.....	14
Fig.2.5 Resolver.....	15
Fig.2.6 Excitation and Output Signal of the Resolver.....	16
Fig.2.7 Voltage Source Inverter Connected to a Motor.....	17
Fig.2.8 Inverter with IGBTs and Antiparallel Diodes.....	18
Fig.2.9 IGBT Symbol and Equivalent Circuit.....	18
Fig.2.10.a PWM current controller	20
Fig.2.10.b pulse wave, showing the definitions of y_{min} , y_{max} and D.....	21
Fig.2.11 Hysteresis controller.....	22
Fig.2.12 summary of classification of Permanent Magnet Motors has been shown...	23
Fig.2.13 Surface Permanent Magnet Motor.....	24
Fig.2.14 Interior Permanent Magnet Motor.....	25
Fig.2.15 Motor with 3-phase concentrated windings.....	25
Fig.2.16 Diagram showing 3-phase windings with: (a) overlapping spread coils, (b) concentrated coils.....	26
Fig.2.17 Circuit diagrams of 3-phase inverters: (a) uni-polar inverter, (b) bipolar inverter.....	26
Fig.2.18 Scheme of 3-phase PM, salient-pole motor with buried magnets in the rotor.....	27
Fig.2.19 Scheme of 3-phase PM motor with the surface mounted magnets on the cylindrical rotor.....	28
Fig.2.20 Supply circuit scheme of permanent magnet synchronous motor (PMSM).	28
Fig.2.21 Supply circuit scheme of brushless dc (BLDC) PM motor.....	28
Fig.2.22 ideal sinusoidal vs. trapezoidal back EMF waveforms, normalized to RMS=1.....	30
Fig.2.23 . Ideal back emf 's, phase currents, and position sensor signals.....	32
Fig.2.24. BLDC motor cross section and phase energizing sequence.....	33
Fig.2.25. Simplified bldc drive scheme.....	33

Fig.2.26. PM Synchronous Motor[30].....	34
Fig.2.27. Basic block diagram of PID controller.....	37
Fig.3.1 Equivalent circuit of 3-phase PM BLDC motor.....	45
Fig. 3.2 Schematic representation of equation 3.1.....	46
Fig. 3.3 Illustration of the rotor position angle θ_e	47
Fig.3.4 Scheme of the mechanical system.....	48
Fig.3.5.a Block Diagram for BLDC motor operation.....	50
Fig.3.5.b Block diagram behind subsystem shown in Fig.3.5.a.....	51
Fig.3.5.c Block diagram behind subsystem shown in Fig.3.5.b.....	52
Fig.3.6 Waveforms of electromechanical quantities obtained from: the start-up process of the BLDC motor, and steady-state.....	54
Fig.3.7 Diagram of DC commutator motor, which visually explains its operation...55	
Fig.3.8 Scheme of a DC motor with 3 coils or phases in the armature: (a) coils commutated by mechanical commutator, (b) coils commutated by electronic commutator (inverter).....	56
Fig.3.9 Scheme of DC motor with 3-phase star connected winding: (a) with mechanical commutator (winding placed on rotor), (b) with inverter (winding is on stator and magnetic poles rotate).....	57
Fig.3.10 Rotor positions at two subsequent instants: (a) at time t_1 , (b) at time t_2 and (c) mutual positions of the armature flux with respect to the field flux position at the two time instants.....	58
Fig.3.11 Brush DC motor armature circuit diagram.....	58
Fig.3.12 Brush motor's equivalent circuit model with armature inductance.....	59
Fig.3.13 Simulation model of BLDC motor with controller.....	62
Fig.3.14. Structure of PI controller.....	63
Fig.3.15. The MATLAB simulation of PI controller.....	64
Fig.4.1 Explanation of synchronous motor operation.....	67
Fig.4.2.a ABC equivalent circuit diagram.....	67
Fig.4.2.b Motor side of Fig. 4.2.a.....	67
Fig.4.3.a Block diagram of synchronous motor operation.....	69
Fig.4.3.b Block diagram behind subsystem shown in Fig.4.3 a.....	70
Fig.4.4. Phasor diagram showing angles, voltage and current.....	71
Fig.4.5 Synchronous machine equivalent circuit: (a) armature reaction reactance X_{ar} , armature leakage reactance X_{al} , (b) synchronous reactance X_s , (c) armature	

impedance Z_s	73
Fig.4.6. Phasor diagram drawn for $R_a = 0$	75
Fig.4.7 Torque-power angle characteristic.....	77
Fig.4.8. Permanent magnet synchronous motor drive system.....	79
Fig.4.9. Motor axis.....	80
Fig.4.10. Scheme of ABC and dq system.....	81
Fig.4. 11. Equivalent circuit of a PM synchronous motor.....	83
Fig.4.12. MATLAB Simulation of d-q model of PMSM.....	84
Fig.4.13. Three phase voltage source PWM inverter circuit.....	85
Fig.4.14.a. SIMULINK model of voltage source PWM inverter circuit.....	86
Fig.4.14.b. Phase A voltage input and output of PWM inverter.....	87
Fig.4.15. MATLAB /SIMULINK of PI controller.....	87
Fig.4.16. subsystem of (a) Inverse (Park & Clarke) transform (b) (Park & Clarke) transform.....	88
Fig.4.17. Inverse Park Transform.....	88
Fig.4.18. MATLAB /SIMULINK of Inverse Park Transform.....	89
Fig.4.19. Inverse Clark Transform.....	89
Fig.4.20. MATLAB /SIMULINK of Inverse Clark Transform.....	90
Fig.4.21. Clark Transform.....	90
Fig.4.22. illustrate MATLAB /SIMULINK of Clark Transform.....	91
Fig.4.23. Park Transform.....	91
Fig.4.24. MATLAB /SIMULINK of Park Transform.....	92
Fig.4.25. MATLAB /SIMULINK of PMSM drive system.....	92
Fig.5.1. Phase A voltage [v], and current [A].....	95
Fig.5.2. BLDC motor, speed response.....	96
Fig.5.3. BLDC motor Electromagnetic Torque with Load Torque.....	96
Fig.5.4. BLDC motor Phase A current response.....	97
Fig.5.5. BLDC motor source current [A].....	97
Fig.5.6. Speed and current-torque characteristics determined in steady-state, from the dynamic simulation.....	98
Fig.5.7. Characteristics obtained from both the BLDC motor model after it reaches steady-state and the brush DC motor model.....	98
Fig 5.8. BLDC Motor speed with PID controller.....	99
Fig.5.9. phase A current of motor with PID controller.....	99

Fig.5.10. 3-Phase voltage source of PMSM.....	100
Fig.5.11. phase A current and voltage of PMSM.....	100
Fig.5.12. Motor speed response of PMSM.....	101
Fig.5.13. Electromagnetic Torque response with Load Torque of PMSM.....	101
Fig.5.14. Input Power of PMSM.....	102
Fig.5.15. Output Power response of PMSM.....	102
Fig.5.16. Efficiency waveform of PMSM.....	103
Fig.5.17. Characteristics of steady-state and dynamic models for the synchronous motor.....	103
Fig.5.18a. Speed response of PMSM with PID controller.....	104
Fig.5.18b. Zoom in Speed Response of PMSM with PID controller.....	104

SYMBOLS & ABBREVIATIONS

A	phase- A
B	phase -B
C	phase -C
PM	Permanent Magnet Motor
BLDC	Brushless DC Motor
PMSM	Permanent Magnet Synchronous Motor
PWM	Pulse Width Modulation
rpm	Revolution Per minute
emf	Electromotive Force
mmf	Magnet Moving Force
PI	Proportional Integral
PID	Proportional Integral Derivative
d	Direct Polar Axis
q	Quadrature or Interpolar Axis
BJT	Bipolar Junction Transistor
MOSFET	Metal Oxide Semiconductor Field Effect Transistors
IGBT	Insulated Gate Bipolar Transistors
GTO	Gate Turn Off Thyristors
CSI	Current Source Inverters
VSI	Voltage Source Inverters

1. INTRODUCTION

Permanent magnet (PM) motors are widely used in low and mid power applications such as computer peripheral equipments, robotics, adjustable speed drives and electric vehicles.

The increase in the market of PM motor drives has demanded the need of simulation tools capable of use motor drive simulations. Simulations have helped the process of developing new systems including motor drives, by reducing cost and time.

Simulation tools have the capabilities of performing dynamic simulations of motor drives in a visual environment so as to make easy the development of new systems.

In this work, the simulation of a PM motor drive system was developed using MATLAB Simulink. The simulation circuit included all realistic components of the drive system. This enables the calculation of currents and voltages in different parts of the inverter and motor under transient and steady-state conditions. The losses in different parts can be calculated to ease the design of the inverter. Implementation has been done in Simulink. A comparative study of two mode of PM motors (brushless dc motor and synchronous motor) has been made in terms of transient and dynamic performance of PM motor. Simulation results are given for two the modes, brushless dc motor (BLDC) and permanent magnet synchronous motor (PMSM).

1.1 Motivation.

Modeling and simulation is usually used in designing PM drives compared to building system prototypes because of the cost. Having selected all components, the simulation process can start to calculate steady state and dynamic performance and losses that would have been obtained if the drive were actually constructed. This practice reduces time, cost of building prototypes and ensures that requirements are achieved.

1.2 Objectives of the Thesis

The objectives of this project are;

- To determine the performance of the permanent magnet (PM) motor while operating as a (BLDC) motor and (PMSM) .
- To study the difference in operation between the two mentioned above.

The tasks to be accomplished are as follows;

- A Literature study on:
 - Permanent magnet AC motors supplied from DC/AC converters.
 - Control of the motor using the PID controller.
- Modeling of the PM 3-phase synchronous and PM 3-phase brushless DC motors in dynamic and steady-state conditions.
- Determination of the synchronous and brushless DC motor's performance in steady-state and dynamic conditions with the application of the MATLAB package.
- Control of motors (PMSM and BLDC) using the PID controller.
- Comparison of simulation results and conclusions.

1.3 Outline of the Present Work

The thesis is divided into 6 chapters.

Chapter 2 This chapter deals with the brief description of the different components such as position sensors, inverters , current controllers of the drive system and concentrate on permanent magnet motors. A review of permanent magnet materials and classification of permanent magnet motors, so PM motor speed control is also given.

Chapter 3 deals with the detailed dynamic modeling of BLDC motor, simulation of motor dynamics , operation of the BLDC motor in steady state , explain the similarity and difference between BLDC motor and Brush PM DC motor, BLDC Motor control.

Chapter 4 is presents the dynamic modeling and simulation of PM synchronous motor, focuses on the abc model and the second system known as dq model, discussed the synchronous motor in steady state model, so the Permanent Magnet Synchronous Motor Control with dq model is present.

Chapter 5 illustrate the results obtained from the simulation of permanent magnet motor models (BLDC motor and PMSM), and offers a brief description for all figures.

Chapter 6 offers a conclusion, including a general description of what was done in the project and a comparison of the motors' performance in the dynamic and steady-state models before and after controlling.

1.4 Previous Work

PM motor drives have been a topic of interest for the last twenty years. Different authors have carried out modeling and simulation of such drives.

In 1986 Sebastian, T., Slemon, G. R. and Rahman, M. A. [1] reviewed permanent magnet synchronous motor advancements and presented equivalent electric circuit models for such motors and compared computed parameters with measured parameters. Experimental results on laboratory motors were also given.

The main advantage is the improvement of torque and flux ripple characteristics at any speed region; this provides an opportunity for motor operation under minimum switching loss and noise. So this produces the required torque with minimum torque ripples. And the experimental results on laboratory are identical to the computed results.

In 1986 Jahns, T.M., Kliman, G.B. and Neumann, T.W. [2] discussed that interior permanent magnet (IPM) synchronous motors possessed special features for adjustable speed operation which distinguished them from other classes of ac machines. They were robust high power density machines capable of operating at high motor and inverter efficiencies over wide speed ranges, including considerable range of constant power operation. The magnet cost was minimized by the low magnet weight requirements of the IPM design. The impact of the buried magnet configuration on the motor's electromagnetic characteristics was discussed. The rotor magnetic saliency preferentially increased the quadrature-axis inductance and introduced a reluctance torque term into the IPM motor's torque equation. The electrical excitation requirements for the IPM synchronous motor were also discussed. The control of the sinusoidal phase currents in magnitude and phase angle with respect to the rotor orientation provided a means for achieving smooth responsive torque control. A basic feed forward algorithm for executing this type of current vector torque control was discussed, including the implications of current regulator saturation at high speeds. The key results were illustrated using a combination of simulation and prototype IPM drive measurements.

A speed controller has been designed successfully for closed loop operation of the PMSM drive system so that the motor runs at the commanded or reference speed. The simulated system has a fast response with zero steady state error , with smooth response of electromagnetic torque, and there is no overshoot .

In 1988 Pillay and Krishnan, R. [3], presented PM motor drives and classified them into two types such as permanent magnet synchronous motor drives (PMSM) and brushless dc motor (BLDC) drives. The PMSM has a sinusoidal back emf and requires sinusoidal stator currents to produce constant torque while the BLDC motor has a trapezoidal back emf and requires rectangular stator currents to produce constant torque. The PMSM is very similar

to the wound rotor synchronous machine except that the PMSM that is used for servo applications tends not to have any damper windings and excitation is provided by a permanent magnet instead of a field winding. Hence the d, q model of the PMSM can be derived from the well known model of the synchronous machine with the equations of the damper windings and field current dynamics removed. Equations of the PMSM are derived in rotor reference frame and the equivalent circuit is presented without dampers. The damper windings are not considered because the motor is designed to operate in a drive system with field-oriented control. Because of the non sinusoidal variation of the mutual inductances between the stator and rotor in the BLDC motor, it is also shown in this paper that no particular advantage exists in transforming the abc equations of the BLCD motor to the d, q frame.

The concluded results was, the ripple torque of the BLDC motor is higher than that of the PMSM. The ripple torque in the PMSM is due to only to the ripple in the currents. These ripple torques are of high frequency and are easily damped out by the rotor. In addition to these ripples, the BLDC motor has commutation ripple that depends on the speed of the machine. This makes the BLDC motor less suitable for high performance position applications. The PMSM is capable of higher speed range than a BLDC of the same parameters.

As an extension of his previous work, Pillay, P. and Krishnan, R. in 1989 [4] presented the permanent magnet synchronous motor (PMSM) which was one of several types of permanent magnet AC motor drives available in the drives industry. The motor had a sinusoidal flux distribution. The application of vector control as well as complete modeling, simulation, and analysis of the drive system were given. State space models of the motor and speed controller and real time models of the inverter switches and vector controller were included. The machine model was derived for the PMSM from the wound rotor synchronous motor. All the equations were derived in rotor reference frame and the equivalent circuit was presented without dampers. The damper windings were not considered because the motor was designed to operate in a drive system with field-oriented control. Performance differences due to the use of pulse width modulation (PWM) and hysteresis current controllers were examined. Particular attention was paid to the motor torque pulsations and speed response and experimental verification of the drive performance were given.

The torque pulsations may be different due to the use of different controllers, the average value, which determined the overall speed response, is the same. The motor torque

pulsation are related linearly to the size of the hysteresis bands. And from the results obtained the overshoot, rise time, steady state error, torque ripple are eliminated.

Morimoto, S., Tong, Y., Takeda, Y. and Hirasaka, T. in 1994 [5], in their paper aimed to improve efficiency in permanent magnet (PM) synchronous motor drives. The controllable electrical loss which consisted of the copper loss and the iron loss could be minimized by the optimal control of the armature current vector. The control algorithm of current vector minimizing the electrical loss was proposed and the optimal current vector could be decided according to the operating speed and the load conditions. The proposed control algorithm was applied to the experimental PM motor drive system, in which one digital signal processor was employed to execute the control algorithms, and several drive tests were carried out. The operating characteristics controlled by the loss minimization control algorithm were examined in detail by computer simulations and experimental results.

Simulation and experimental results show that the proposed control scheme guarantees stable and robust response of the PMSM drive, under a wide range of operating conditions. Subsequently, it can be utilized in high performance motion control applications.

The paper in 1997 by Wijenayake, A.H. and Schmidt, P.B. [6], described the development of a two-axis circuit model for permanent magnet synchronous motor (PMSM) by taking machine magnetic parameter variations and core loss into account. The circuit model was applied to both surface mounted magnet and interior permanent magnet rotor configurations. A method for on-line parameter identification scheme based on no-load parameters and saturation level, to improve the model, was discussed in detail. Test schemes to measure the equivalent circuit parameters, and to calculate saturation constants which govern the parameter variations were also presented.

The proposed field oriented vector controlled PMSM drive can handle the effects of step change in reference speed and parameter variations. The overall system performances are quite good in terms of dynamic, transient and steady-state responses.

In 1997 Jang-Mok, K. and Seung-Ki, S. [7], proposed a novel flux weakening scheme for an Interior Permanent Magnet Synchronous Motor (IPMSM). It was implemented based on the output of the synchronous PI current regulator reference voltage to PWM inverter. The on-set of flux weakening and the level of the flux were adjusted inherently by the outer voltage regulation loop to prevent the saturation of the current regulator. Attractive features of this flux weakening scheme included no dependency on the machine

parameters, the guarantee of current regulation at any operating condition, and smooth and fast transition into and out of the flux weakening mode.

With this type of control method, wide range of speed control is possible for different loading conditions. From the simulation results, it is evident that during step change in speed command, the transient response in current, speed and torque in terms of rise and settling time are high compared to the transient response during the step change in torque command.

Bose, B. K., in 2001 [8], presented different types of synchronous motors and compared them to induction motors. The modeling of PM motor was derived from the model of salient pole synchronous motor. All the equations were derived in synchronously rotating reference frame and was presented in the matrix form. The equivalent circuit was presented with damper windings and the permanent magnet was represented as a constant current source.

Bowen, C., Jihua, Z. and Zhang, R. in 2001 [9], addressed the modeling and simulation of permanent magnet synchronous motor supplied from a six step continuous inverter based on state space method. The motor model was derived in the stationary reference frame and then in the rotor reference frame using Park transformation.

The obtained results from stationary reference frame model are closer to that in the rotor reference frame model as rise time, overshoot, steady state error, input and output power, but controller used with rotor reference frame model.

In 2002 Mademlis, C. and Margaris, N. [10], presented an efficiency optimization method for vector-controlled interior permanent-magnet synchronous motor drive. Based on theoretical analysis, a loss minimization condition that determines the optimal q-axis component of the armature current was derived. Selected experimental results were presented to validate the effectiveness of the proposed control method.

In 2004, Jian-Xin, X., Panda, S. K., Ya-Jun, P., Tong Heng, L. and Lam, B. H. [11] applied a modular control approach to a permanent-magnet synchronous motor (PMSM) speed control. Based on the functioning of the individual module, the modular approach enabled the powerfully intelligent and robust control modules to easily replace any existing module which did not perform well, meanwhile retaining other existing modules which were still effective. Property analysis was first conducted for the existing function modules in a conventional PMSM control system: proportional integral (PI) speed control module,

reference current-generating module, and PI current control module. Next, it was shown that the conventional PMSM controller was not able to reject the torque pulsation which was the main hurdle when PMSM was used as a high-performance servo. By virtue of the internal model, to nullify the torque pulsation it was imperative to incorporate an internal model in the feed-through path. This was achieved by replacing the reference current-generating module with an iterative learning control (ILC) module. The ILC module records the cyclic torque and reference current signals over one entire cycle, and then uses those signals to update the reference current for the next cycle. As a consequence, the torque pulsation could be reduced significantly. In order to estimate the torque ripples which might exceed certain bandwidth of a torque transducer, a novel torque estimation module using a gain-shaped sliding-mode observer was further developed to facilitate the implementation of torque learning control. The proposed control system was evaluated through real-time implementation and experimental results validated the effectiveness.

A speed controller has been designed successfully for closed loop operation of the PMSM drive system so that the motor runs at the commanded or reference speed. The simulated system has a fast response with practically zero steady state error thus validating the design method of the speed controller.

Araujo, R.E., Leite, A.V. and Freitas, D.S. in 1997 [12], mentioned the different simulation tools available and the benefits that were obtained by accelerating the process for the development of visual design concepts. Among various software packages for simulation of electronic circuits, like SPICE and SABER, EMTP, EUROSTAG, or for specialized simulations tools for power electronics system like SIMPLORER, POSTMAC, SIMSEN, ANSIM, and PSCAD, they had chosen MATLAB/Simulink. MATLAB/Simulink had user-friendly environment, visual design, Real-Time Workshop and libraries of models for the various components of a power electronic system.

The transient and steady state values of current, speed and torque curves are analyzed by using MATLAB/Simulink. The three phase currents show less distortion and torque curves have very little ripples. Thus PWM technique has the advantage of less overshoot, lower torque pulsation and quick response.

Ong, C in 1998 [13], explained the need for powerful computation tools to solve complex models of motor drives. Among the different simulation tools available for dynamic simulation he had chosen MATLAB/SIMULINK® as the platform for his book because of

the short learning curve required to start using it, its wide distribution, and its general purpose nature.

Macbahi, H. Ba-razzouk, A. Xu, J. Cheriti, A. and Rajagopalan, V. in 2000 [14], mentioned that a great number of universities and researchers used the MATLAB/SIMULINK software in the field of electrical machines because of its advantages such as user friendly environment, visual oriented programming concept, non-linear standard blocks and a large number of toolboxes for special applications.

In 1997 Reece, J.H., Bray, C.W., Van Tol, J.J. and Lim, P.K. [15], discussed three possible computer simulation tools such as PSpice, HARMFLO and the Electromagnetic Transients Program (EMTP) in their project on power systems containing adjustable speed drives. They selected EMTP as the primary simulation tool because of its broad range of capabilities, which were well matched to their problem.

French, C.D., Finch, J.W. and Acarnley, P.P. in 1998 [16], had found that in recent years the increase in desktop computing power has lead to an increase in the sophistication of both design and simulation tools available to the design engineer. One such tool becoming more wide spread amongst academia and industry was Mathwork's Simulink / Matlab package. This paper described how Simulink could be used as an integrated development environment for simulation and real time control of electric motor drive systems. This was carried out with the aid of motor models together with simulation and real time control circuits. It was demonstrated how such a set-up could be used as a cost effective control system rapid prototyping scheme.

Onoda, S. and Emadi, A. in 2004 [17], had developed a modeling tool to study automotive systems using the power electronics simulator (PSIM) software. PSIM was originally made for simulating power electronic converters and motor drives. This user friendly simulation package was able to simulate electric/electronic circuits.

Venkaterama, G. [18]; had developed a simulation for permanent magnet motors using Matlab/simulink. The motor was a 5 hp PM synchronous line start type. Its model included the damper windings required to start the motor and the mathematical model was derived in rotor reference frame. The simulation was presented with the plots of rotor currents, stator currents, speed and torque.

The paper deal with modeling of five phase PMSM. The simulation results show that the performance of the motor during free acceleration at zero load is in the line with the

expected results. The model can be used for study of closed loop drives based on the above motor and its design of motor control theory.

Simulink PM Synchronous Motor Drive demo circuit (2005) [19] used the AC6 block of Simi power Systems library. It modeled a permanent magnet synchronous motor drive with a braking chopper. The PM synchronous motor was fed by a PWM voltage source inverter, which was built using a Universal Bridge Block. The speed control loop used a PI regulator to produce the flux and torque references for the vector control block. The vector control block computed the three reference motor line currents corresponding to the flux and torque references and then fed the motor with these currents using a three-phase current regulator. Motor current, speed, and torque signals were available at the output of the block.

The demo circuit (2005) in Simulink for Permanent magnet synchronous motor fed by PWM inverter [20] had a three-phase motor rated 1.1 kW, 220 V, 3000 rpm. The PWM inverter was built entirely with standard Simulink blocks. Its output went through Controlled Voltage Source blocks before being applied to the PMSM block's stator windings. Two control loops were used. The inner loop regulated the motor's stator currents. The outer loop controlled the motor's speed. Line to line voltages, three phase currents, speed and torque were available at the output of the scope blocks.

The results obtained from the demo circuit in simulink show the speed response development and torque ripple improvement.

2. Description of the drive system of PM motor.

This chapter deals with the brief description of the different components of the drive system of PM motor such as position sensors, inverters, current controllers of the drive system, so a review of permanent magnet materials and classification of permanent magnet motors is also given.

2.1 Permanent magnet motor drive system

The motor drive consists of four main components, the PM motor, inverter, control unit and the position sensor. The components are connected as shown in Figure 2.1.

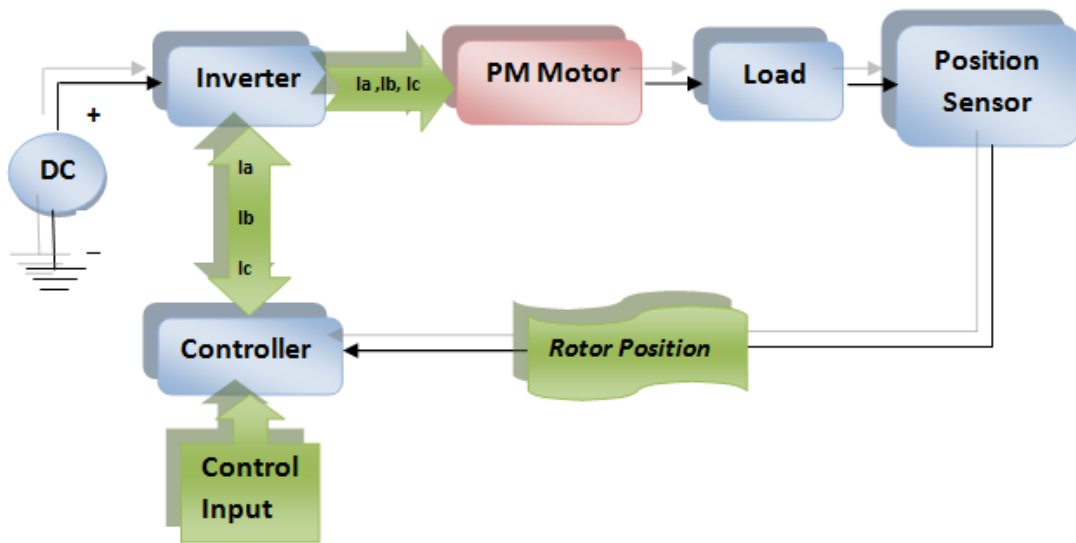


Fig 2.1 Drive System Schematic

Descriptions of the different components are as follows:

2.2 Position sensor

Operation of permanent magnet synchronous motors requires position sensors in the rotor shaft when operated without damper winding. The need of knowing the rotor position requires the development of devices for position measurement. There are four main devices for the measurement of position, the potentiometer, linear variable differential transformer, optical encoder and resolvers. The ones most commonly used for motors are encoders and revolvers. Depending on the application and performance desired by the motor a position sensor with the required accuracy can be selected.

2.2.1 Optical encoders

The most popular type of encoder is the optical encoder as shown in Figure 2.2, which consists of a rotating disk, a light source, and a photo detector (light sensor). The disk, is mounted on the rotating shaft, has coded patterns of blinded and transparent sectors. As the disk rotates, these patterns stop the light emitted onto the photo detector, generating a digital pulse or output signal[35].

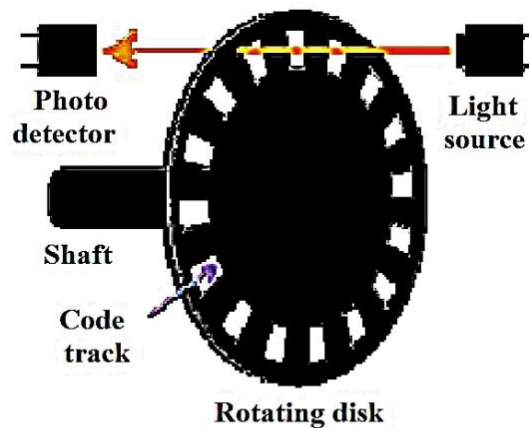


Fig 2.2 Optical Encoder [35]

Optical encoders offer the advantages of digital interface. There are two types of optical encoders, incremental encoder and absolute encoder.

2.2.1.1 Incremental encoders

Incremental encoders have good accuracy and are simple to implement but they suffer from lack of information when the motor is at rest position and in order for exact position the motor must be stop at the starting point.

The most common type of incremental encoder uses two output channels (A and B) to sense position. Using two code tracks with sectors positioned 90° degrees out of phase, the two output channels of the quadrature encoder indicate both position and direction of rotation as shown in Figure 2.3. If A leads B, for example, the disk is rotating in a clockwise direction.

If B leads A, then the disk is rotating in a counter-clockwise direction. By monitoring both, the number of pulses and the relative phase of signals A and B, it's possible to track position and direction of rotation. Some quadrature encoders also include a third output channel, called a zero or index or reference signal, which supplies a single pulse per revolution. This single pulse is used for exact determination of a reference position. The

precision of the encoder is fix by its code disk but it can be increased by detecting the Up and Down transitions on both the A and B channels[36].

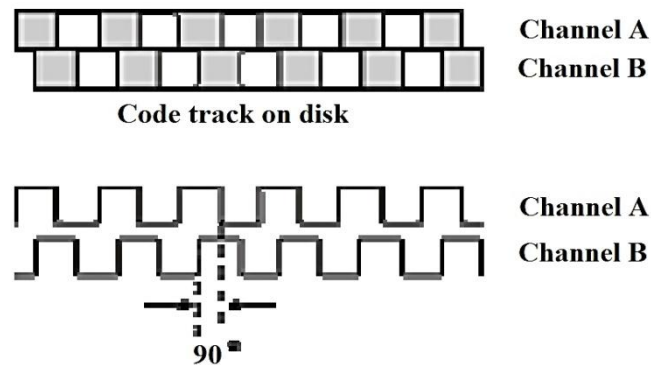


Fig 2.3 Quadrature Encoder Channels [36]

2.2.1.2 Absolute encoders

The absolute encoder, as shown in Figure 2.4 captures the exact position of the rotor with a accuracy directly related to the number of bits of the encoder. It can rotate indefinitely and even if the motor stops the position can be measured or obtained.

It provides a “whole word” output with a unique code pattern representing each position. This code is derived from independent tracks on the encoder disc (one for each “bit” of resolution) corresponding to individual photo detectors. The output from these detectors is HI (light) or LO (dark) depending on the code disc pattern for that particular position. Absolute encoders are used in applications where a device is stationary for long periods of time or moves at a slow rate, such as flood gate control, telescopes, cranes, valves. They are also recommended in systems that must keep hold of position information through a power outage[26].

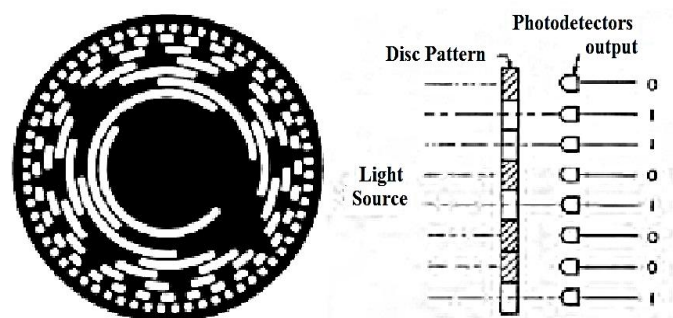


Fig 2.4 Absolute Encoder [26]

2.2.2 Position resolver

Position resolver as shown in Figure 2.5, also called rotary transformers, works on the transformer principle. The primary winding is placed on the rotor and depending upon the rotor shaft angle the induced voltage at the two secondary windings of the transformer shifted by 90° would be different. The position can be calculated using the two voltages.

The resolver is basically a rotary transformer with one rotating reference winding (V_{ref}) and two stator windings. The reference winding is fixed on the rotor, and therefore, it rotates jointly with the shaft passing the output windings, as shown in Figure 2.5. Two stator windings are placed in quadrature (shifted by 90°) with one another and generate the sine and cosine voltages (V_{sin} , V_{cos}) respectively. Both windings will be further referred to as output windings. In effect of the excitement applied on the reference winding V_{ref} and along with the angular movement of the motor shaft θ , the respective voltages are generated by resolver output windings V_{sin} , V_{cos} [27].

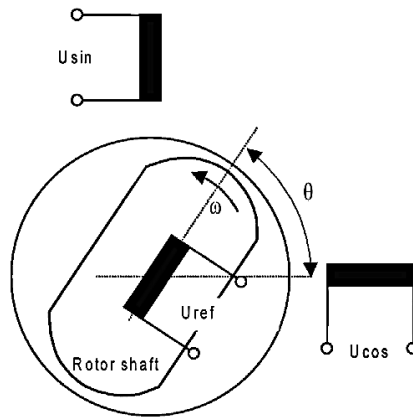


Fig 2.5 Resolver [27]

The frequency of the generated voltages is identical to the reference voltage and their amplitudes vary according to the sine and cosine of the shaft angle θ . Considering that one of the output windings is aligned with the reference winding, then it is generated full voltage on that output winding and zero voltage on the other output winding and vice versa. The rotor angle θ can be extracted from these voltages shown in Figure.2.6.

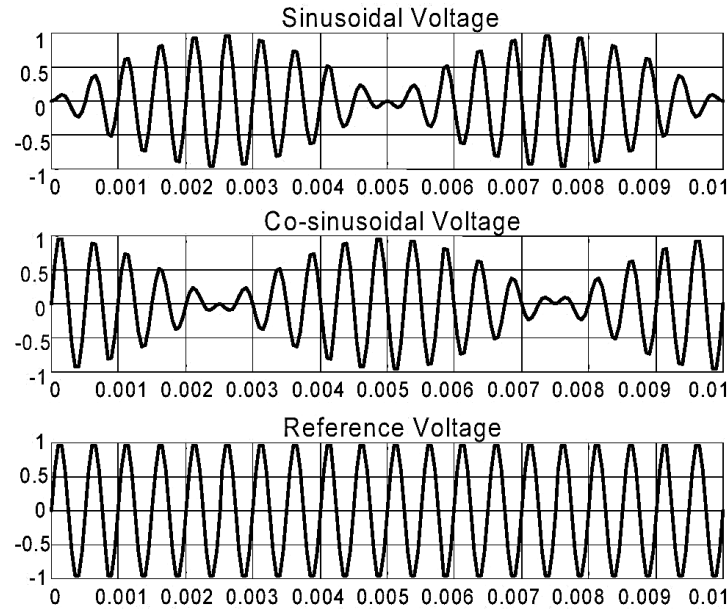


Fig 2.6 Excitation and Output Signal of the Resolver [37]

The shaft angle can be determined by an Inverse Tangent [8] function of the quotient of the sampled resolver output voltages V_{sin} , V_{cos} . This determination can be expressed, in terms of resolver output voltages, as follows:

$$\theta = a \tan \left(\frac{V_{sin}}{V_{cos}} \right) \quad (2.1)$$

2.3 Current controlled inverter

The motor is fed from a voltage source inverter with current control. The control is performed by regulating the flow of current through the stator of the motor. Current controllers are used to generate gate signals for the inverter. Proper selection of the inverter devices and selection of the control technique will ensure the efficiency of the drive.

2.3.1 Inverter

Voltage Source Inverters are devices that convert a DC voltage to AC voltage of variable frequency and magnitude. They are very commonly used in adjustable speed drives and are characterized by a well defined switched voltage wave form in the terminals [23]. Figure 2.7 shows a voltage source inverter. The AC voltage frequency can be variable or constant depending on the application.

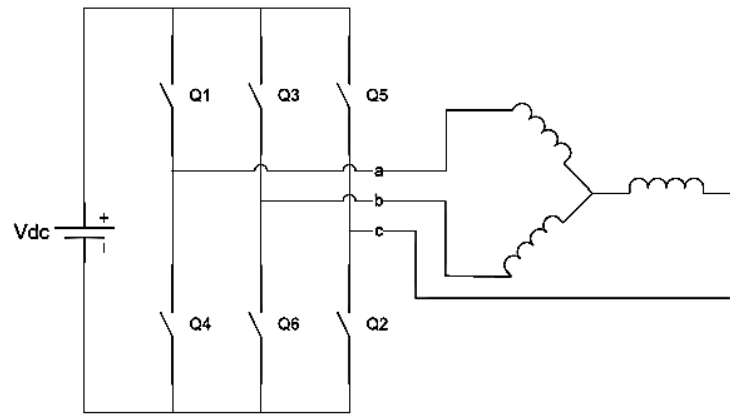


Fig 2.7 Voltage Source Inverter Connected to a Motor [23]

Three phase inverters consist of six power switches connected as shown in Figure 2.7 to a DC voltage source. The inverter switches must be carefully selected based on the requirements of operation, ratings and the application. There are several devices available today and these are thyristors, bipolar junction transistors (BJTs), MOS field effect transistors (MOSFETs), insulated gate bipolar transistors (IGBTs) and gate turn off thyristors (GTOs). The devices list with their respective power switching capabilities are shown in table 2.1 MOSFETs and IGBTs are favored by industry because of the MOS gating permits high power gain and control advantages. While MOSFET is considered a universal power device for low power and low voltage applications, IGBT has wide acceptance for motor drives and other application in the low and medium power range.

Table 2.1 Device Power and Switching Capabilities

Device	Power Capability	Switching Speed
BJT	Medium	Medium
GTO	High	Low
IGBT	Medium	Medium
MOSFET	Low	High
THYRITOR	High	Low

The power devices when used in motor drives applications require an inductive motor current path provided by antiparallel diodes when the switch is turned off. Inverters with antiparallel diodes are shown in Figure 2.8.

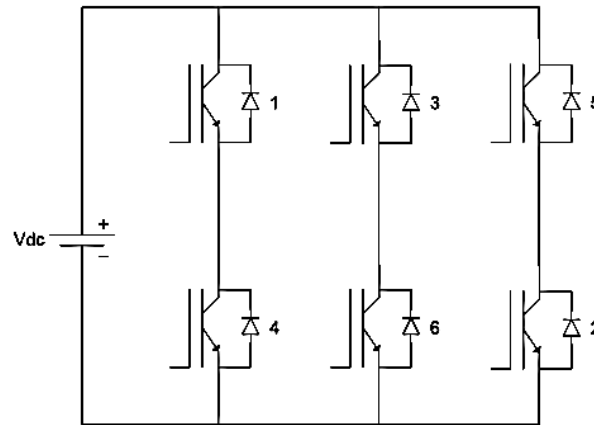


Fig 2.8 Inverter with IGBTs and Antiparallel Diodes

2.3.2.IGBTs

IGBTs provide high input impedance and are used for high voltage applications. The high input impedance allows the device to switch with a small quantity of energy and for high voltage applications the device must have large blocking voltage ratings. The device behavior is described by parameters like voltage drop or on-resistance, turn on time and turn off time. The symbolic representation and the equivalent circuit of an IGBT are shown in Figure 2.9.

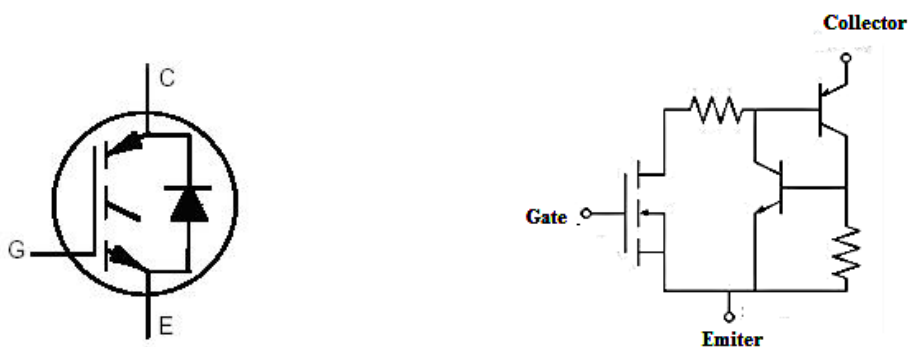


Fig 2.9 IGBT Symbol and Equivalent Circuit

2.3.3 Current control

The power converter in a high-performance motor drive used in motion control fundamentally functions as a power amplifier, reproducing the low power level control signals generated in the field orientation controller at power levels appropriate for the driven machine. High-performance drives employ control strategies which develop command signals for the AC machine currents. The basic reason for the selection of current as the controlled variable is the same as for the DC machine; the stator dynamics (effects of stator resistance, stator inductance, and induced EMF) are eliminated. Thus, to the extent that the current regulator functions as an ideal current supply, the order of the system under control is reduced and the complexity of the controller can be much simplified. Current regulators for AC drives are complex because an AC current regulator must control both the amplitude and phase of the stator current. The AC drive current regulator forms the inner loop of the overall motion controller. As such, it must have the widest bandwidth in the system and must, be necessary, have zero or nearly zero steady-state error. Both current source inverters (CSI) and voltage source inverters (VSI) can be operated in controlled current modes. The current source inverter is a "natural" current supply and can readily be adapted to controlled current operation. The voltage source inverter requires more complexity in the current regulator but offers much higher bandwidth and elimination of current harmonics as compared to the CSI and is almost exclusively used for motion control applications.

Current controllers can be classified into two groups, hysteresis and PWM current controllers. Both types are discussed below.

2.3.3.1 PWM current controller

PWM current controllers are widely used. The switching frequency is usually kept constant. They are based in the principle of comparing a triangular carrier wave of desired switching frequency and is compared with error of the controlled signal. The error signal comes from the sum of the reference signal generated in the controller and the negative of the actual motor current. The comparison will result in a voltage control signal that goes to the gates of the voltage source inverter to generate the desired output. Its control will respond according to the error. If the error command is greater than the triangle waveform, the inverter leg is held switched to the positive polarity (upper switch on). When the error command is less than the triangle waveform, the inverter leg is switched to the negative

polarity (lower switch on). This will generate a PWM signal like in Figure 2.10.a. The inverter leg is forced to switch at the frequency of the triangle wave and produces an output voltage proportional to the current error command. The nature of the controlled output current consists of a reproduction of the reference current with high-frequency PWM ripple superimposed[23].

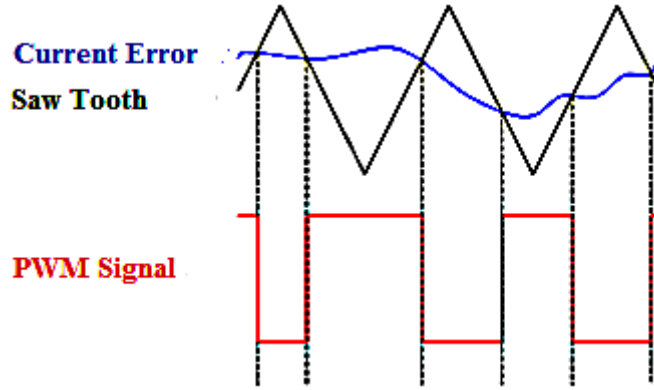


Figure 2.10.a. PWM current controller[23]

Pulse-width modulation uses a rectangular pulse wave whose pulse width is modulated resulting in the variation of the average value of the waveform. If we consider a pulse waveform $f(t)$, with period T , low value y_{min} , a high value y_{max} and a duty cycle D (see Figure 2.10.b), the average value of the waveform is given by:

$$\tilde{y} = \frac{1}{T} \int_0^T f(t) dt \quad (2.2)$$

Where: $f(t)$ is a pulse wave, its value is y_{max} for $0 < t < D * T$, and y_{min} for $D * T < t < T$. The above expression then becomes:

$$\tilde{y} = \frac{1}{T} \left(\int_0^{DT} y_{max} dt + \int_{DT}^T y_{min} dt \right) \quad (2.3)$$

$$\tilde{y} = \frac{D * T * y_{max} + T(1-D)y_{min}}{T} \quad (2.4)$$

$$\tilde{y} = D * y_{max} + (1 - D)y_{min} \quad (2.5)$$

This latter expression can be fairly simplified in many cases where $y_{min} = 0$ as $\tilde{y} = D * y_{max}$. From this, it is obvious that the average value of the signal ($\tilde{y}$) is directly dependent on the duty cycle D .

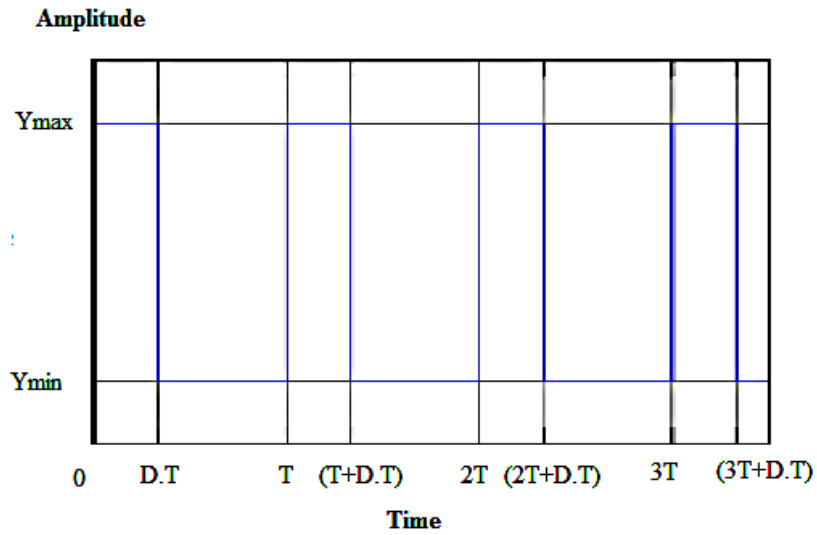


Figure 2.10.b. pulse wave, showing the definitions of y_{min} , y_{max} and D .

2.3.3.2 Hysteresis current controller

Hysteresis current controller can also be implemented to control the inverter currents. The controller will generate the reference currents with the inverter within a range which is fixed by the width of the band gap. In this controller the desired current of a given phase is summed with the negative of the measured current. The error is fed to a comparator having a hysteresis band. When the error crosses the lower limit of the hysteresis band, the upper switch of the inverter leg is turned on. But when the current attempts to become less than the upper reference band, the bottom switch is turned on. Figure 2.11 shows the hysteresis band with the actual current and the resulting gate signals. This controller does not have a specific switching frequency and changes continuously but it is related with the band width [4] [23].

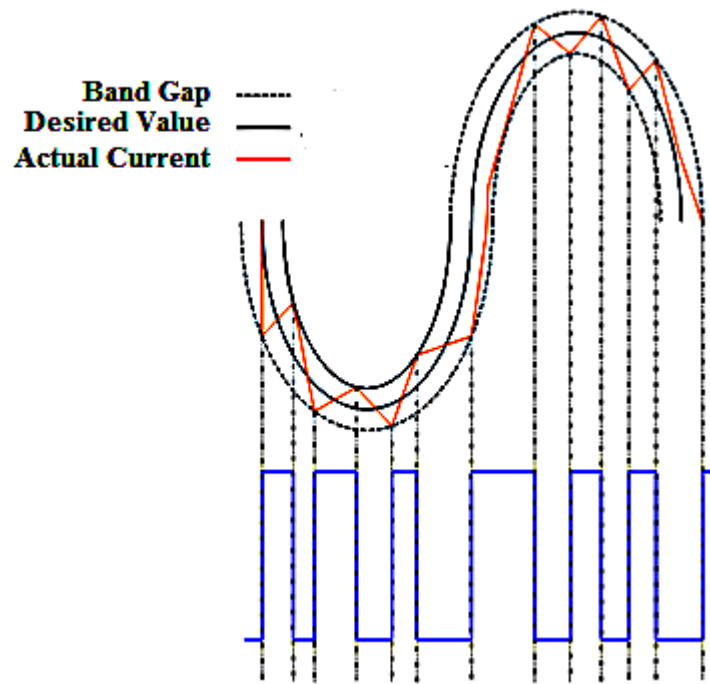


Fig 2.11 Hysteresis controller

2.4 Permanent magnet motor

A permanent magnet motor is a motor that uses permanent magnets to produce the air gap magnetic field rather than using electromagnets. These motors have important advantages, attracting the interest of researchers and industry for use in many applications.

2.4.1 Classification of permanent magnet motors

In Figure 2.12 summary of classification of Permanent Magnet Motors has been shown.

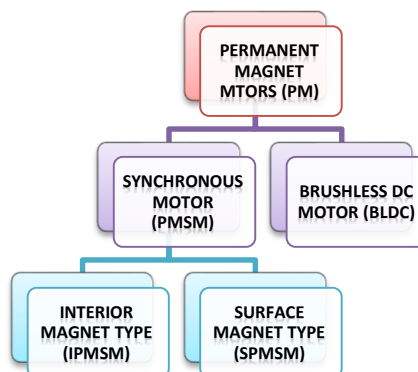


Fig 2.12 summary of classification of Permanent Magnet Motors has been shown

2.4.1.1 Direction of field flux

PM motors are generally classified by the direction of the field flux. The first field flux classification is radial field motor, meaning that the flux is along the radius of the motor. The second is axial field motor, meaning that the flux is perpendicular to the radius of the motor. Radial field flux is most commonly used in motors and axial field flux have become a topic of interest for study and used in a few applications.

2.4.1.2 Flux density distribution

PM motors are classified on the basis of the flux density distribution and the shape of current excitation. They are PM synchronous motor (PMSM) and PM brushless dc motors (BLDC).

The PMSM has a sinusoidal-shaped back EMF and is designed to develop sinusoidal back EMF waveforms.

They have the following:

1. Sinusoidal distribution of magnet flux in the air gap
2. Sinusoidal current waveforms
3. Sinusoidal distribution of stator conductors.

BLDC motor has a trapezoidal-shaped back EMF and is designed to develop trapezoidal back EMF waveforms.

They have the following:

1. Rectangular distribution of magnet flux in the air gap
2. Rectangular current waveform
3. Concentrated stator windings.

2.4.1.3 Permanent magnet radial field motors

In PM motors, the magnets can be placed in two different ways on the rotor. Depending on the placement they are called either as surface permanent magnet motor or interior permanent magnet motor.

Surface mounted PM motors have a surface mounted permanent magnet rotor. Each of the PM motor is mounted on the surface of the rotor, making it easy to build, and specially skewed poles are easily magnetized on this surface mounted type to minimize cogging torque.

This configuration is used for low speed applications because of the limitation that the magnets will break up during high-speed operations. These motors are considered to have small saliency, thus having practically equal inductances in both axes [22]. The permeability of the permanent magnet is almost that of the air, thus the magnetic material becoming an extension of the air gap. For a surface permanent magnet motor $L_d = L_q$ [8].

The rotor has an iron core that may be solid or may be made of punched laminations for simplicity in manufacturing [23]. Thin permanent magnets are mounted on the surface of this central part using a glue. Alternating magnets of the opposite magnetization direction produce radially directed flux density across the air gap. This flux density then reacts with currents in windings placed in slots on the inner surface of the stator to produce torque. Figure 2.13 shows the placement of the magnet.

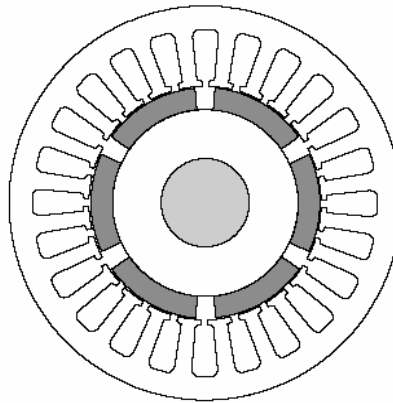


Fig 2.13 Surface Permanent Magnet Motor [21]

Interior PM motors have interior mounted PM rotor as shown in Figure 2.14. Each permanent magnet is mounted inside the rotor. It is not as common as the surface mounted type but it is a good nominee for high-speed operation. There is inductance variation for this type of rotor because the permanent magnet part is equivalent to air in the magnetic circuit calculation. These motors are considered to have saliency with q axis inductance greater than the d axis inductance ($L_q > L_d$) [8].

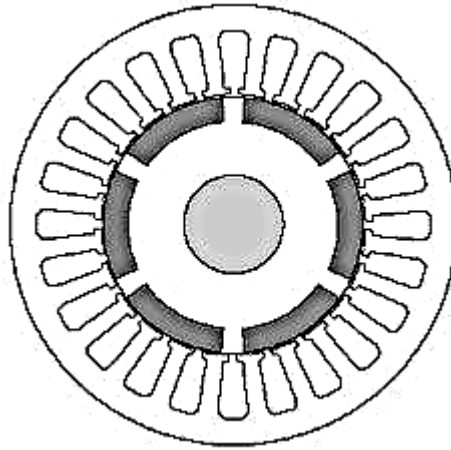


Fig 2.14 Interior Permanent Magnet Motor [21]

2.4.2 Operation techniques of PMSM and BLDC motors

Any PM single-phase or multi-phase motor can operate as a synchronous or BLDC motor. The particular mode of operation depends on the supply and control circuit. This is generally true if the stator has conventional windings placed in slots and distributed symmetrically around the stator periphery.

The BLDC motor, due to its different type of supply, may also have other types of windings with concentrated coils. Such a motor is shown in Figure.2.15. The BLDC motor with concentrated windings evolved from the stepper motor structure. In fact, the stepper motor operating at high speed does not differ much from the BLDC motor operation.

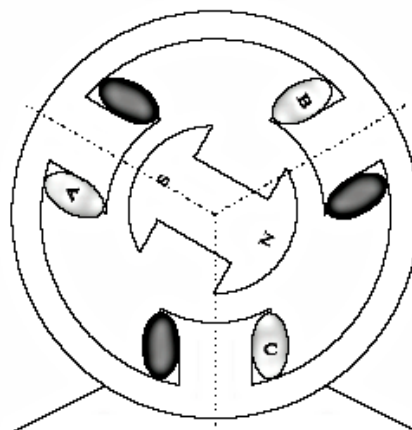


Fig. 2.15 Motor with 3-phase concentrated windings.

The winding diagrams with spread and concentrated coils are shown in Figure. 2.16. The 3-phase windings with distributed coils are supplied from a bi-polar inverter Figure 2.17b and those with concentrated coils are usually supplied from a uni-polar converter Figure 2.17a. The BLDC motor has to be equipped with a position sensor which informs the controller what the position of the rotor magnetic pole is, with respect to the particular stator phase winding. This is done in order to switch the motor ON and OFF. The position sensors that are applied are usually optical and Hall's [24.]. The encoder will act as a position sensor, in the experiment carried out in this project. The BLDC motor can also operate by using sensor-less control. In this case, the position of the rotor is known from the value of the back EMF with respect to the particular phase [25].

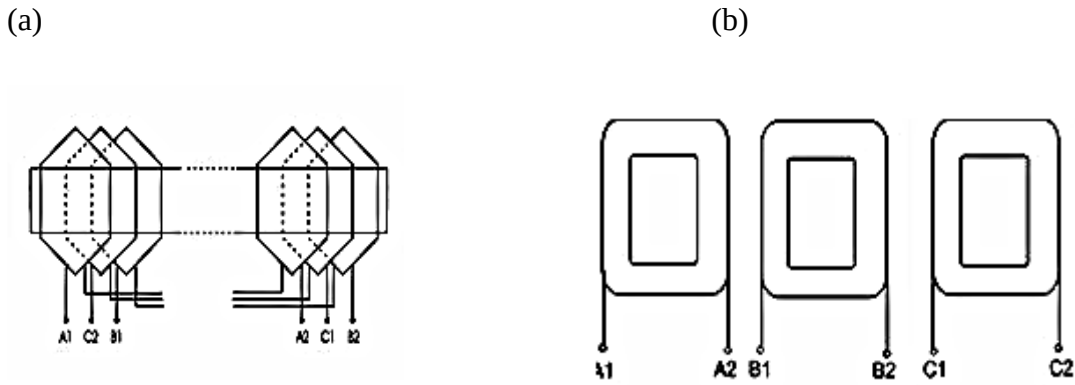


Fig. 2.16 Diagram showing 3-phase windings with: (a) overlapping spread coils, (b) concentrated coils[25].

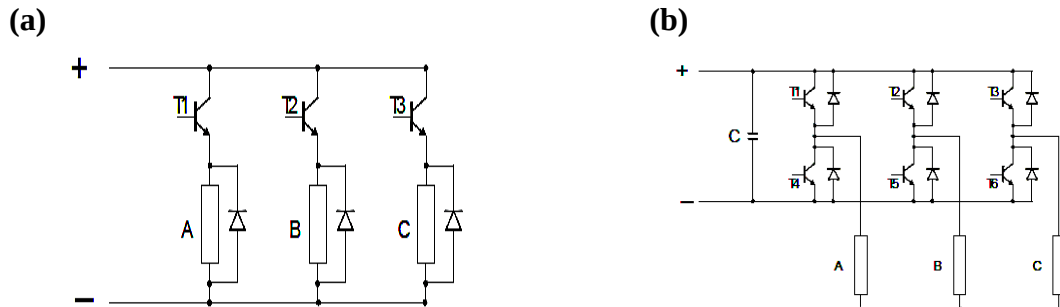


Fig. 2.17 Circuit diagrams of 3-phase inverters: (a) uni-polar inverter, (b) bipolar inverter[25].

In each case (PMSM and BLDC) of permanent magnet modes, the motor consists of a wound stator and a rotor. The stator may have a single-phase or multi-phase winding which is sometimes called the armature winding Figure 2.18. The rotor just has permanent magnets.

There are two types of rotors.

- Salient-pole rotor as shown in Figure 2.18 – mostly for low-speed machines
- Cylindrical rotor as shown in Figure 2.19 – usually for high-speed machines.

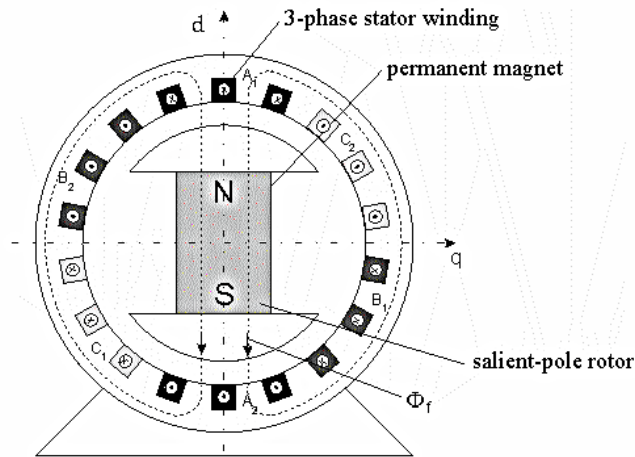


Fig. 2.18 Scheme of 3-phase PM, salient-pole motor with buried magnets in the rotor.

In both cases (PMSM and BDCM), the PM can be attached to the rotor surface Figure 2.19 or it can be buried. Figure 2.18 Since there is no winding in the rotor (that would be supplied through the brushes), the AC PM machines are also called brushless PM machines. The rotor of the AC PM motor rotates synchronously with the magnetic field generated by the stator winding. At the motor start-up, the rotor has zero speed. Due to its high inertia, it cannot instantaneously shoot to its synchronous speed. However, this can happen if we supply the stator winding with frequency, rising gradually from 0 to its rated value. To do this, the motor has to be supplied from a variable frequency inverter.

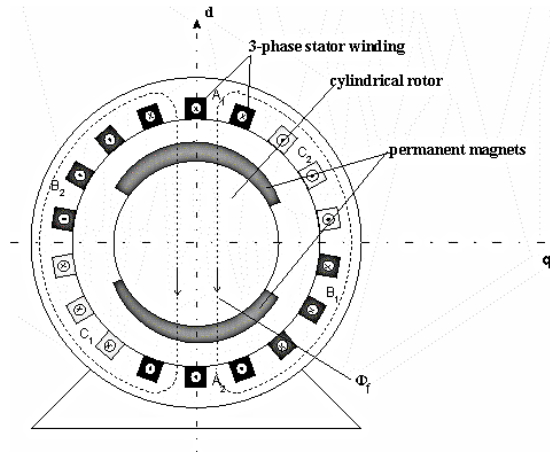


Fig. 2.19 Scheme of 3-phase PM motor with the surface mounted magnets on the cylindrical rotor.

The frequency can be controlled by imposing the desired reference frequency (see Figure 2.20), or the motor itself can set a suitable frequency value required for its actual speed. In the second case, the motor must be equipped with a speed or position sensor Figure.2.21. The AC PM motor with the frequency controlled as described in the previous mode above operates as a synchronous motor. On the other hand, if operating with “self controlled frequency” as explained in the latter mode above, then this operation is that of a brushless dc (BLDC) motor.

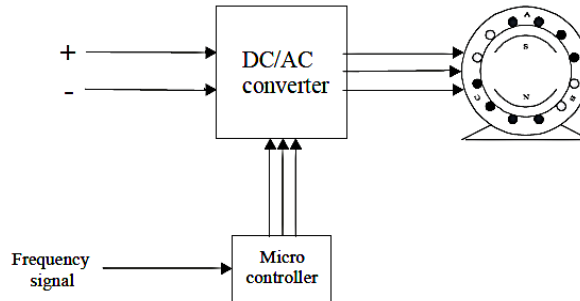


Fig. 2.20 Supply circuit scheme of permanent magnet synchronous motor (PMSM) [24].

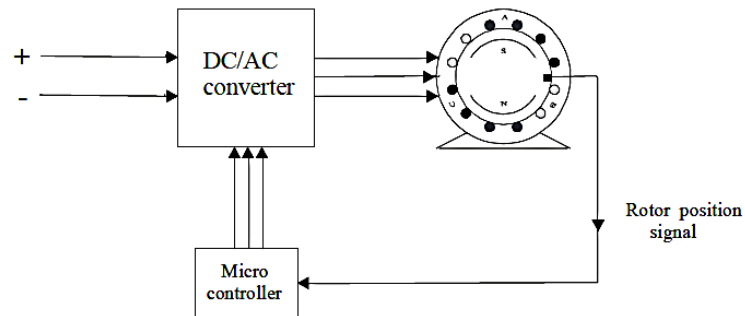


Fig. 2.21 Supply circuit scheme of brushless dc (BLDC) PM motor [24].

brushless dc PM motor is not different in construction from brushless synchronous PM motor. The difference between these two motors lies in the control of supply circuit. This influences the motor performance of these two motors.

One of the major difference in their operation is that brushless dc PM motor operates with the speed that varies with the load, while the speed of brushless synchronous PM motor is constant at the variable load. Another difference is that the brushless dc PM motor develops the starting torque while the synchronous motor does not. Thus, its electromechanical characteristics will differ importantly.

Modern research [26]-[27] has indicated that the permanent magnet motor drives, which include the permanent magnet synchronous motor (PMSM) and the brushless dc motor (BLDC) could become serious competitors to the induction motor (IM) for servo applications. The PMSM has a sinusoidal back EMF and requires sinusoidal stator currents to produce constant torque while the BLDC motor has a trapezoidal back EMF and requires rectangular stator currents to produce constant torque. The PMSM is very similar to the standard wound rotor synchronous machine except that the PMSM has no damper windings and excitation is provided by a permanent magnet instead of a field winding.

2.4.2.1 Trapezoidal -vs - sinusoidal back EMF

To start solve the problem, brushless motors themselves can be divided into two different types based on the shape of their back EMF, sinusoidal and trapezoidal. The classification is based on the shape of the back EMF waveform of the motor, which is the voltage it produces at its terminals as a function of rotor position with no load. The amplitude of the back EMF is proportional to the angular velocity of the motor, but its shape will not change with speed. This relationship is completely related by the following equations [28]:

$$\lambda_r = \lambda_r(\theta) \quad (2.2)$$

$$E = \frac{d\lambda_r}{dt} \quad (2.3)$$

Where:

λ_r : rotor flux linkage , E : back electromotive force

Rotor flux linkage, λ_r , is a function of rotor angular position. Back EMF, E , is the rate of change of rotor flux linkage in the winding. Therefore, the amplitude of the back EMF waveform is a function of angular velocity and the shape is a function of angular position. Some factors effects the shape are: magnet geometry, magnetization, stator core geometry, and winding distribution. These are all properties of the motor itself, and do not depend on the drive.

The two extreme shapes that are considered in this section are sinusoidal and trapezoidal. Figure.2.22 shows the ideal sinusoidal and trapezoidal back EMF waveforms.. In order to keep the comparison “fair,” the amplitudes of the two back EMF waveforms are normalized such that they both have an RMS value of 1.

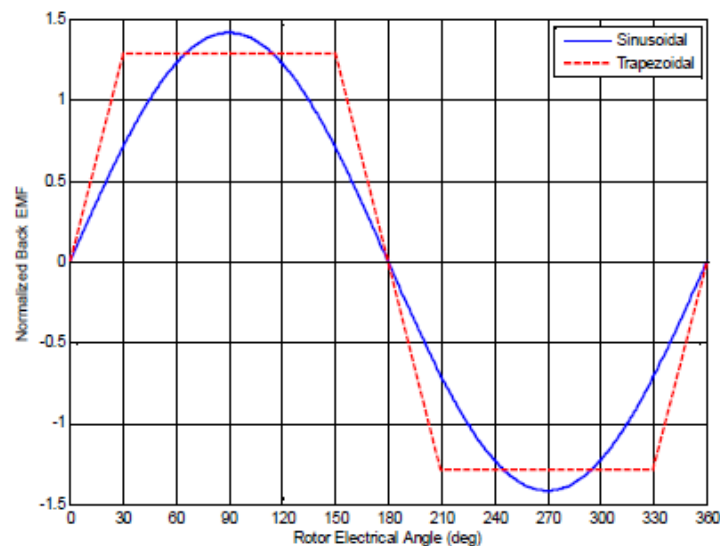


Fig. 2.22 Ideal sinusoidal vs. trapezoidal back EMF waveforms, normalized to RMS=1[28]

Some physical conditions that would lead to a trapezoidal back EMF are:

- Concentrated windings.
- No stator or magnet skew.
- Discrete magnet poles with uniform magnetization.

These conditions all lead to sharp transitions in the flux linkage, which give the trapezoidal back EMF waveform its different shape. While it does affect the shape of the back EMF, stator core saturation is not the reason for the flat top of the back EMF. (Saturation include small rate of change of flux, so it would affect the shape near the back EMF zero crossing, not at the peaks.)

Some physical conditions that would lead to a sinusoidal back EMF are as follows:

- Overlapped or sinusoidal-distributed windings.
- A coreless stator (with windings only, no steel laminations).
- Stator and/or magnet skew.
- Sinusoidal magnetization.

These conditions all round off the flux linkage so that it more closely approximates an ideal sinusoid. Practically, a motor with these characteristics can be more difficult to make, and thus more costly. In general, low-cost motors tend to have a more of the characteristics that lead to a trapezoidal back EMF [28].

2.4.3.2 BLDC motor specifics and operation

A BLDC motor is a rotating electric machine where the stator is a classic 3-phase wound stator, like that of an induction motor, and the rotor has surface-mounted permanent magnets. When the wound stator is energized by a 3-phase alternating current, it creates a rotating magnetic flux that causes the rotor to rotate synchronously with it.

In a BLDC motor, the position of the rotor (and hence its permanent magnetic field) is sensed with respect to the stator coils (phases), and the supply current is switched electronically (commutated) to the appropriate phases. BLDC motor control systems often include either internal or external position sensors to sense the actual rotor position. Alternatively, the rotor position can be detected without sensors by measuring the back-EMF in each stator winding.

The BLDC motor is driven by trapezoidal currents coupled with the given rotor position. The three phase BLDC motor is operated in a two-phases-on method. i.e. the two phases that produce the highest torque are energized while the third phase is off. Which two phases are energized depends on the rotor position.

The signals from the position sensors produce a three digit number that changes every 60° (electrical degrees) as shown in Figure 2.23 . (H1, H2, H3). The figure also shows ideal current and back-EMF waveforms. Figure 2.24. Shows a cross section of a three-phase star-connected motor along with its phase energizing sequence. Each interval starts with the rotor and stator field lines 120° apart and ends when they are 60° apart. Maximum torque is reached when the field lines are perpendicular. Current commutation is done by a six-step inverter as shown in a simplified form in Figure 2.25. The switches are shown as

bipolar junction transistors but MOSFET switches are more common. Table 2.2 shows the switches sequence. The current direction and the position sensor signals.[29]

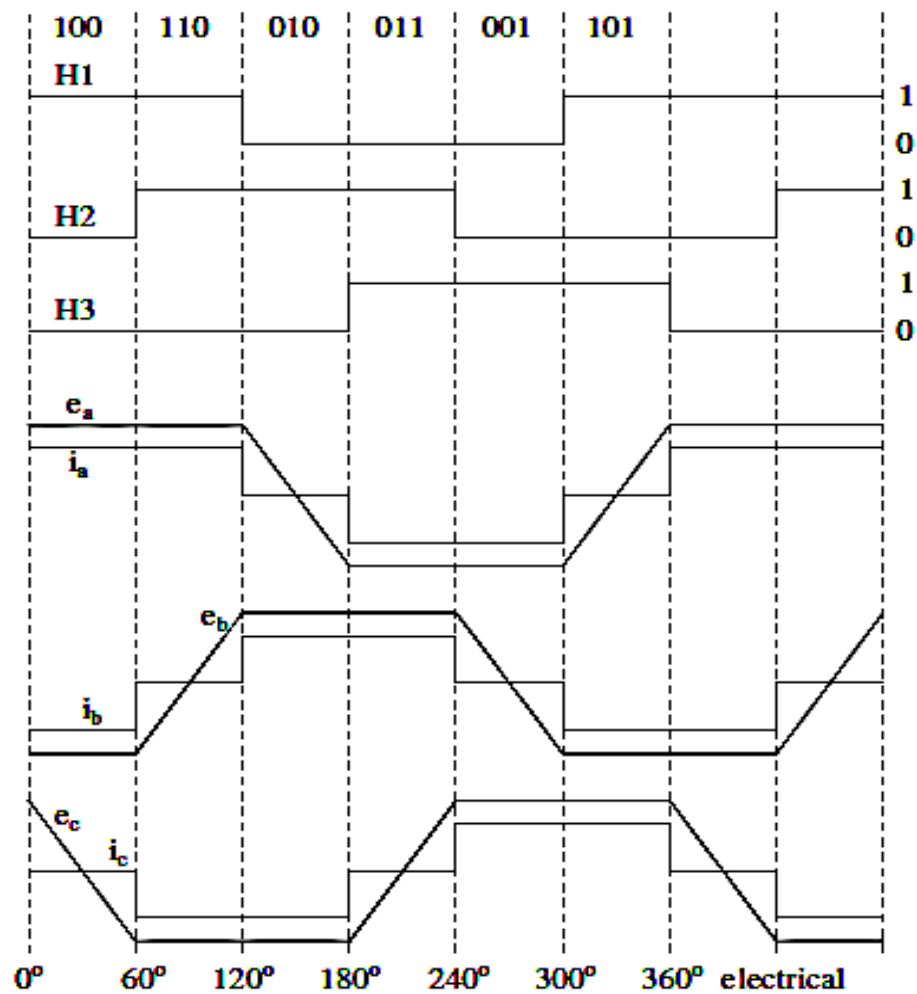


Fig. 2.23 . Ideal back emf 's, phase currents, and position sensor signals[29]

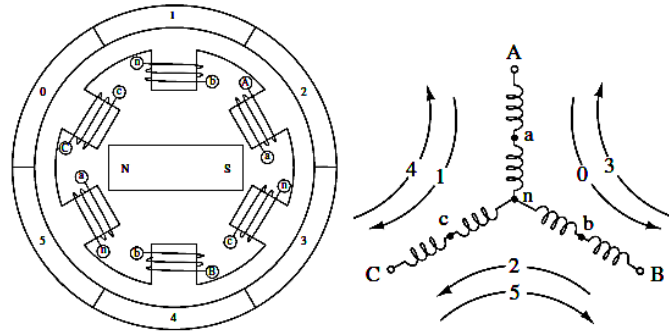


Fig. 2.24. BLDC motor cross section and phase energizing sequence[29]

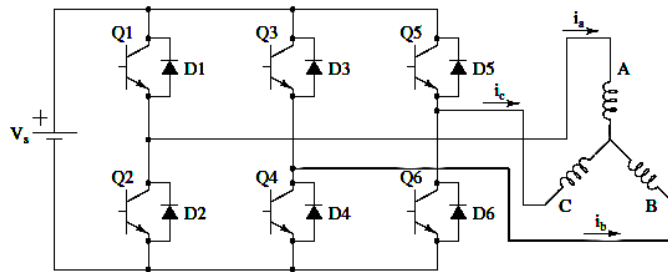


Fig.2.25. Simplified bldc drive scheme[29]

Table 2.2. Switching sequence

Switching interval	Seq. number	Pos. sensors			Switch closed		Phase current		
		H1	H2	H3			A	B	C
$0^\circ - 60^\circ$	0	1	0	0	Q1	Q4	+	-	off
$60^\circ - 120^\circ$	1	1	1	0	Q1	Q6	+	off	-
$120^\circ - 180^\circ$	2	0	1	0	Q3	Q6	off	+	-
$180^\circ - 240^\circ$	3	0	1	1	Q3	Q2	-	+	off
$240^\circ - 300^\circ$	4	0	0	1	Q5	Q2	-	off	+
$300^\circ - 360^\circ$	5	0	0	1	Q5	Q4	off	-	+

2.4.2.3 PMSM specifics and operation

Similar to the BLDC motor, the rotor of a PMSM consists of permanent magnets. The stator of a PMSM has its 3-phase windings distributed sinusoidal, as opposite to the trapezoidal salient-pole distribution found in a BLDC motor. A PMSM operates in the same way as a BLDC motor, when the wound stator is energized by a 3-phase alternating current, it creates a rotating magnetic flux that causes the rotor to rotate synchronously with it.

The PMSM is driven by sinusoidal currents coupled with the given rotor position. Figure 2.26 illustrates a theoretical cross-sectional view of a 3-phase, 2-pole IPM synchronous motor along with two reference frames. To illustrate the inductance difference ($L_q > L_d$), rotor is drawn with saliency although actual rotor structure is more likely a cylinder. The stator reference axis for the a-phase is chosen to the direction of maximum magnet moving force (mmf) when a positive a-phase current is supplied at its maximum level. Reference axis for b- and c- stator frame are chosen 120° and 240° (electrical angle) ahead of the a-axis, respectively. Following the convention of choosing the rotor reference frame, the direction of permanent magnet flux is chosen as the d-axis, while the q-axis is 90 degrees ahead of the d-axis. The angle of the rotor q-axis with respect to the stator a-axis is defined as θ . Note that as the machine turns, the d-q reference frame is rotating at a speed of $\omega = \frac{d\theta}{dt}$, while the stator a,b,c- axes are fixed in space. We will find out later that the choice of this rotating frame greatly simplifies the dynamic equations of the model.

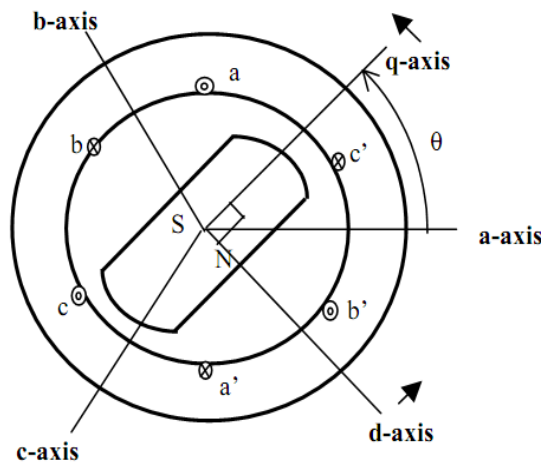


Fig.2.26. PM Synchronous Motor[30]

2.4.2.4 Differences between BLDC motor and PMSM

The expression “BLDC” is a marketing driven label to advance the idea that a BLDC motor with the appropriate drive is a suitable drop-in replacement for a brushed DC motor and its drive.

Technically, a BLDC motor is still an AC motor; that is, AC inputs are required in order for the motor to move.

BLDC motors usually are trapezoidal commutated (also known as six-step commutation for a 3-phase motor), using Hall effect position sensors for simplicity. The motors are typically trapezoidal wound; when driven, the back-EMF profile is also approximately trapezoidal. The generated torque is almost constant, but it has significant torque ripple which occurs at each step of the six-step commutation sequence.

In difference, PMSM employs sinusoidal commutation typically working from a high resolution position feedback sensor such as a resolver, and tends to be better controlled with a more complicated sinusoidal commutation algorithm. PMSM has sinusoidal distributed windings, which then create a sinusoidal back-EMF when the rotor rotate. The drive signals to the stator windings are controlled to generate current waveforms that are sinusoidal. The resulting torque is smoother with less ripple than a similar BLDC motor, although the peak torque available from a PMSM is lower.

Both BLDC motor and PMSM are quite similar, with the shape of the optimal input stator currents being the main difference. A BLDC motor can be driven by either trapezoidal or sinusoidal currents, although trapezoidal currents provide a better match for the stator windings. The sinusoidal excitation of the stator windings in a PMSM generates a much smoother resultant torque than that of a BLDC motor.

A summary of the differences between both types of motor is given in Table 2.3

Table 2.3. A summary of the Differences between BLDC Motor and PMSM

Parameter	3-Phase BLDC Motor	3-Phase PMSM
Windings distribution	Trapezoidal	Sinusoidal
Energizes phase	Two phase on at any time	Three phases on at any time
Stator current	Square/trapezoidal wave	Sinusoidal wave
Back-EMF	Trapezoidal wave	Sinusoidal wave
Electromagnetic torque	Almost constant	Constant
Torque strength	Stronger	Weaker

2.5. PM motor speed control

Many applications, such as robotics and factory automation, require exact control of speed and position. Speed Control Systems allow one to easily set and adjust the speed of a motor. The control system consists of a speed feedback system, a motor, an inverter, a controller and a speed setting device. A properly designed feedback controller makes the system not sensitive to disturbance and changes of the parameters.

The purpose of a motor speed controller is to take a signal representing the demanded speed, and to drive a motor at that speed. Closed Loop speed control systems have fast response, but become expensive due to the need of feedback components such as speed sensors.

It is interesting to note that most of the industrial controllers in use today use PID or modified PID control schemes. Analog PID controllers are mostly hydraulic ,pneumatic, electric, and electronic types or their combinations. now, many of these are transformed into digital forms through the use of microprocessors.

Because most PID controllers are adjusted on site, many different types of tuning rules have been proposed. But in this thesis automatic tuning method have been used.

2.5.1.PID controller

A proportional – integral – derivative controller (PID controller) is a control loop feedback mechanism (controller) widely used in industrial control system [40]. Figure 2.27. show basic block diagram of PID controller. The PID is the most commonly used feedback controller. A PID controller calculates an error value as the difference between a measured process variable and a desired set point. The controller attempts to minimize the error by adjusting the process control inputs. The PID controller calculation (algorithm) involves three separate constant parameters. And is accordingly sometimes called three-term control, the proportional, the integral and derivative values, denoted as P,I, and D respectively. These values can be interpreted in terms of time ; P depends on the present error, I on the accumulation of past errors, and D is a prediction of future errors, based on current rate of change. The weighted sum of these three action is used to adjust the process. By tuning the three parameter in the PID controller algorithm, the controller can provide control action designed for specific process requirements. The response of the controller can be described in terms of the sensitivity of the controller to an error, the degree to which the controller overshoots the set point and the degree of system oscillation. Note that the use of the PID algorithm for control does not guarantee optimal control of the system or system stability. PID control is very useful method used in feedback control system. The error generated after the comparison between the measured signal and the target signal is proportionally multiplied (proportional), integrated (integral) and differentiated (derivative) and the outputs of the three operations are linearly summed to generate the signal applied to the actuator.

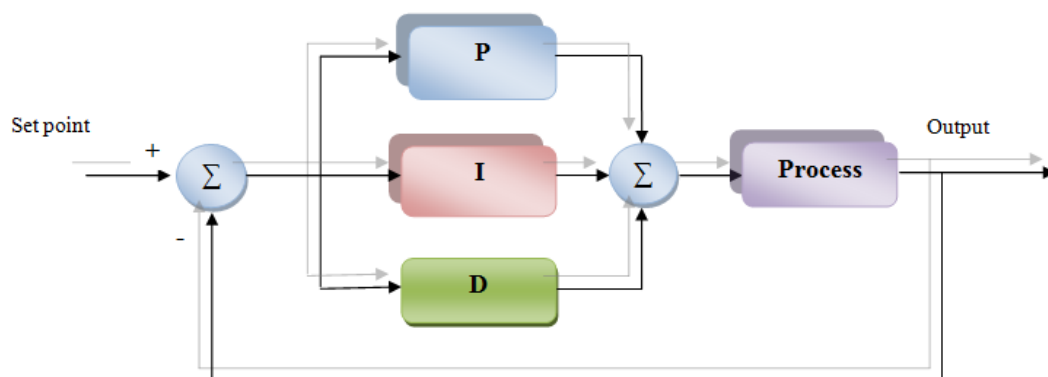


Fig.2.27. Basic block diagram of PID controller

Some application may require using only one or two action to provide the appropriate system control. This is achieved by setting the other parameter to zero. A PID controller will be called a PI, PD, P or I controller in the absence of the respective control action. PI controllers fairly common, since derivative action is sensitive to measurement noise, where the absence of an integral term may avoid the system from reaching its target value due to the control action.

2.5.1.1 PID controller theory

The PID control scheme is named after its three correction terms, whose sum constitutes the manipulated variable (MV). The proportional, integral, and derivative terms are summed to calculate the output of the PID controller [40]. Defining $u(t)$ as the controller output the final form of the PID algorithm is :

$$u(t) = K_p e(t) + K_i \int_0^t e(t) dt + K_d \frac{d}{dt} e(t) \quad (2.4)$$

Where:

K_p : Proportional gain, a tuning parameter

K_i : Integral gain, a tuning parameter

K_d : Derivative gain, a tuning parameter

e : error = SP – PV (SP: Set point , PV: Process Variable)

t : Time or instantaneous time (the present)

2.5.1.2. Proportional term (P)

The proportional term makes a change to the output that is proportional to the current error value. The proportional response can be adjusted by multiplying the error by a constant K_p called the proportional gain. The proportional term is given by :

$$P_{out} = K_p e(t) \quad (2.5)$$

A high proportional gain results in large change in the output for a given change in the error. If the proportional gain is too high, the system can become unstable. In contrast, a small gain results in small output response to a large input error, and a less responsive or less sensitive controller. If proportional gain is too low, the control action may be too small when responding to system disturbances. Tuning theory and industrial practice indicate that the proportional term should give the largeness of the output change.

2.5.1.3. Integral term (I)

The contribution from the integral term is proportional to both the magnitude of the error and the duration of the error. The integral in a PID controller is the sum of the instantaneous error over time and gives the accumulated offset that should have been corrected previous. The integral term is given by:

$$I_{out} = K_i \int_0^t e(t) dt \quad (2.6)$$

The integral term accelerates the movement of the process towards set point and eliminates the remaining steady state error that occurs with a pure proportional controller. However, since the integral term responds to accumulated errors from the past, it can cause the present value to overshoot the set point value.

2.5.1.4. Derivative term (D)

The derivative of the process error is calculated by determining the slope of the error over time and multiplying this rate of change by the derivative gain, K_d . The magnitude of the contribution of the derivative term to the overall control action is termed the derivative gain, K_d . The derivative term is given by:

$$D_{out} = K_d \frac{d}{dt} e(t) \quad (2.7)$$

The derivative term slows the rate of change of the controller output. Derivative control is used to reduce the magnitude of the overshoot produced by the integral component and improve the combined controller process stability. However, the derivative term slows transient response of the controller. Also, differentiation of a signal amplifies noise and thus this term in the controller is highly sensitive to noise in the error term, and can cause a process to become unstable if the noise and the derivative gain are sufficiently large. Hence an approximation to a differentiation with a limited bandwidth is more commonly used. Such a circuit is known as a phase lead compensator.

2.5.2. Loop tuning

Tuning a control loop is the adjustment of its control parameter (gain/proportional band, integral gain/reset, derivative gain/rate) to optimum values for the desired control response [40]. Stability (bounded oscillation) is a basic requirement, but further than, the different systems have different behavior, different applications have different requirements, and requirements may disagreement with one another. Some processes have a degree of nonlinearity and so parameters that work well at full load conditions don't work when the process is starting up from no load. This can be corrected by gain scheduling (using different parameters in different operating regions). PID controllers often provide acceptable control using default tunings, but performance can generally be improved by careful tuning, and performance may be unacceptable with poor tuning. PID tuning a difficult problem, although there are only three parameters and in principle is simple to describe, because it must satisfy complex criteria with the limitations of PID control. There are accordingly various methods for loop tuning, and more difficult techniques are the subject of patents; this section describes some traditional manual methods for loop tuning. If the PID controllers (the gains of the proportional, integral and derivative terms) are chosen incorrectly, the controlled process input can be unstable, i.e. its output diverges, with or without oscillation, and is limited only by saturation or mechanical breakage. Instability is caused by overload gain, particularly in the presence of important lag. Generally, stability of response is required and the process must not oscillate for any combination of process conditions and set points, though sometimes marginal stability (bounded oscillation) is acceptable or desired.

The optimum behavior on a process change or set point change varies depending on the application. Two basic requirements are regulation (disturbance rejection – staying at a given set point) and command tracking (implementing set point changes) ,these refer to how well the controlled variable tracks the desired value. Specific criteria for command tracking include rise time and setting time. Some process must not allow an overshoot of the process variable away from the set point if, for example, this would be unsafe. Other processes must minimize the energy expended in reaching a new point. There are several methods for tuning a PID loop. The most effective methods generally include the development of some form of process model, then choosing P, I, and D based on the dynamic model parameters. Manual tuning methods can be relatively ineffective, particularly if the loops have response times on the order of minutes or longer. The choice of method will depend largely on whether or not the loop can be taken "offline" for tuning, and the response time of the system. If the system can be taken offline, the best tuning

method often include 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.

2.5.3 Manual tuning

If the system must remain online, one tuning method is to first set K_i and K_d values to zero. Increase the K_p until the output of the loop oscillates, then the K_p should be set to approximately half of that value for a "quarter amplitude" type. Then increase K_i until any offset (overshoot, steady-state error, stability) is correct in sufficient time for the process. However, too much K_i will cause instability. Finally, increase K_d , if required, until the loop is acceptably quick to reach its reference after a load disturbance. However, too much will cause too much response and overshoot. A fast PID loop tuning usually overshoots slightly to reach the set point more quickly; however, some systems cannot accept overshoot, in which case an over damped closed loop system is required, which will require a K_p setting significantly less than half that of the K_p setting causing oscillation.

Effects of increasing the parameters independently:

Table 2.4 Recommended PID value setting

Parameter	Rise Time	Overshoot	Settling Time	Steady state error	Stability
K_p	Decrease	Increase	Small Change	Decrease	Degrade
K_i	Decrease	Increase	Increase	Decrease significantly	Degrade
K_d	Small decrease	Small decrease	Small decrease	No effect in theory	Improve if K_d is small

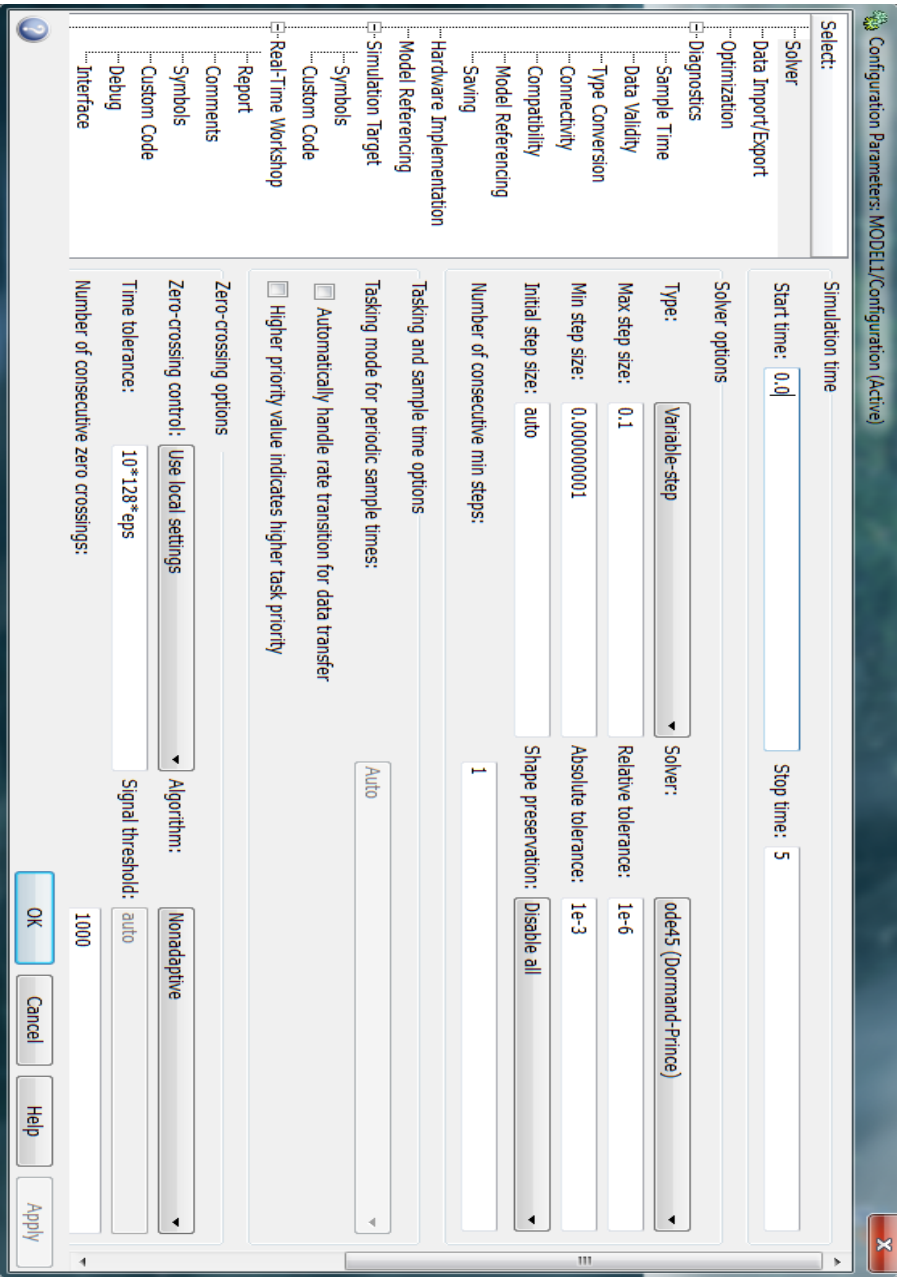
2.6 Data of the motor under study.

The motor used in this research, i.e. whose data is being used and whose performance is being analyzed, is the 3-phase PM brushless DC motor. Table 2.5 is the data shown on the name plate. As seen in Figure 2.16, the motor has three connectors for the three phases, A, B and C.

Table 2.5 Data of the motor under study.

Motor Type	Permanent Magnet motor
Rated stator voltage (Vs rat)	380 V
Rated Frequency (f rat)	60Hz
Rated Speed (w, rat)	1500 rpm
Armature resistance (Ra)	3.59 Ω
Per phase inductance (L Ph)	0.0435 H
Moment of inertia (J)	0.015
Rated Torque (T)	6.45 N.m
Number of pole pairs (P)	3
Direct-axis inductance (Ld)	0.036 H
Quadrature-axis inductance (Lq)	0.051 H
Permanent magnet flux (λ_m)	0.545 Vs
constant, K_E	0.148 V.rad/s

The simulation conditions are as follows : the integration method used was the Runge-Kutta-5 methods, the maximum and minimum step size used are 0.1sec and 0.000000001 sec, respectively with a relative tolerance of $1e^{-6}$ relative and absolute tolerance of $1e^{-3}$ as bellow.



3. Modelling, simulation and control of BLDC motor.

This chapter deals with the detailed dynamics of BLDC motor, performance of the BLDC motor in steady state, similarity and difference between BLDC motor and Brush PM DC motor, and BLDC motor speed control.

3.1 Dynamics of the BLDC motor

In this section we study the dynamic model of BLDC motor, and simulate the motor dynamics

3.1.1 Dynamic model of BLDC motor

It is assumed that the BLDC motor is connected to the output of the inverter, while the inverter input terminals are connected to a constant supply voltage, as was shown in Figure 2.17. The equivalent circuit model that refers to this circuit diagram is shown in Figure.3.1. Another assumption is that there are no power losses in the inverter and the 3-phase motor winding is connected in star.

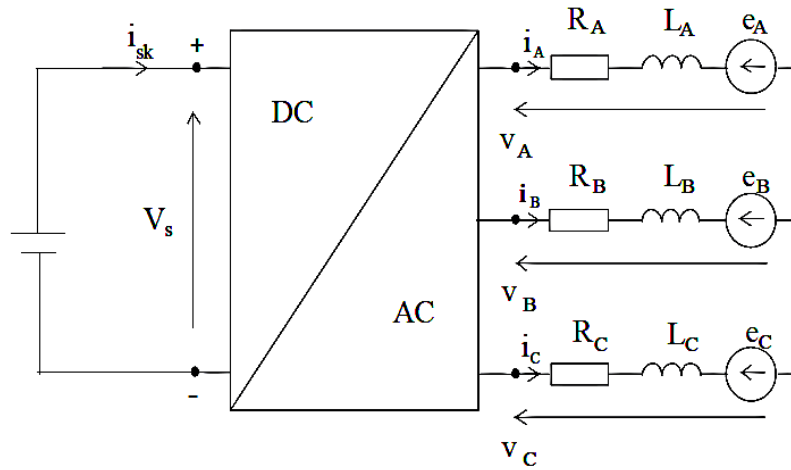


Fig. 3.1 Equivalent circuit of 3-phase PM BLDC motor.[31]

The equivalent circuit shown in Figure.3.1 can be represented by the circuit diagram in Figure. 3.2. The equations that represent this model are as follows.

$$V_A = V_N + V_{SA}$$

$$V_B = V_N + V_{SB}$$

(3.1)

$$V_C = V_N + V_{SC}$$

Where :

V_{SA}, V_{SB}, V_{SC} - are the inverter output voltages that supply the 3 – phase winding.

V_A, V_B, V_C - are the voltages across the motor armature winding.

V_N - voltage at the neutral point.

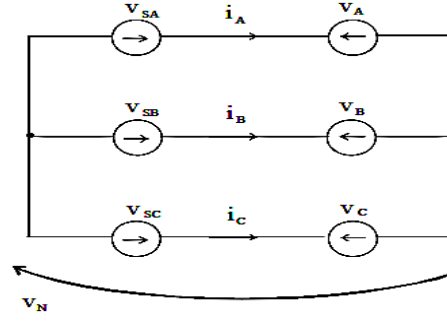


Fig. 3.2 Schematic representation of equation 3.1.

For a symmetrical winding and balanced system, the voltage equation across the motor winding is as follows:

$$\begin{bmatrix} V_A \\ V_B \\ V_C \end{bmatrix} = \begin{bmatrix} R_A & 0 & 0 \\ 0 & R_B & 0 \\ 0 & 0 & R_C \end{bmatrix} \begin{bmatrix} i_A \\ i_B \\ i_C \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} L_A & L_{AB} & L_{AC} \\ L_{BA} & L_B & L_{BC} \\ L_{CA} & L_{CB} & L_C \end{bmatrix} \begin{bmatrix} i_A \\ i_B \\ i_C \end{bmatrix} + \begin{bmatrix} e_A \\ e_B \\ e_C \end{bmatrix} \quad (3.2)$$

or in the shortened description representing the vector:

$$V_a = R_a \cdot I_a + \frac{d}{dt} (L_a I_a) + E_a \quad (3.3)$$

Since $R_A = R_B = R_C = R_a$, the resistance takes the following vector form:

$$R_a = \begin{bmatrix} R_a & 0 & 0 \\ 0 & R_a & 0 \\ 0 & 0 & R_a \end{bmatrix} \quad (3.4)$$

As for the inductances, since the self and mutual inductances are constant for surface mounted permanent magnets on the cylindrical rotor, and the winding is symmetrical:

$$L_A = L_B = L_C = L, \text{ and } L_{AB} = L_{BC} = L_{CA} = L_{BA} = L_{AC} = L_{CB} = M \quad (3.5)$$

Hence the inductance takes the form:

$$L_a = \begin{bmatrix} L & M & M \\ M & L & M \\ M & M & L \end{bmatrix} \quad (3.6)$$

For a Y-connected stator winding,

$$i_A + i_B + i_C = 0 \quad (3.7)$$

Therefore, the voltage takes the following form:

$$\begin{bmatrix} V_A \\ V_B \\ V_C \end{bmatrix} = \begin{bmatrix} R_A & 0 & 0 \\ 0 & R_B & 0 \\ 0 & 0 & R_C \end{bmatrix} \begin{bmatrix} i_A \\ i_B \\ i_C \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} L_s & 0 & 0 \\ 0 & L_s & 0 \\ 0 & 0 & L_s \end{bmatrix} \begin{bmatrix} i_A \\ i_B \\ i_C \end{bmatrix} + \begin{bmatrix} e_A \\ e_B \\ e_C \end{bmatrix} \quad (3.8)$$

where the synchronous inductance, $L_s = L - M$

The angle between a particular phase and the rotor, at any given time, is called θ_e . Figure.

3.3 illustrates the position of this angle, with respect to phase **A** for example.

Since phase **A** is chosen as the reference (see Figure. 3.3), the electromotive forces written in the form of a matrix, E_a take the form:

$$E_a = \frac{K_E}{P} \begin{bmatrix} \sin \theta_e \\ \sin \left(\theta_e - \frac{2}{3}\pi \right) \\ \sin \left(\theta_e - \frac{4}{3}\pi \right) \end{bmatrix} \frac{d\theta_e}{dt} \quad (3.9)$$

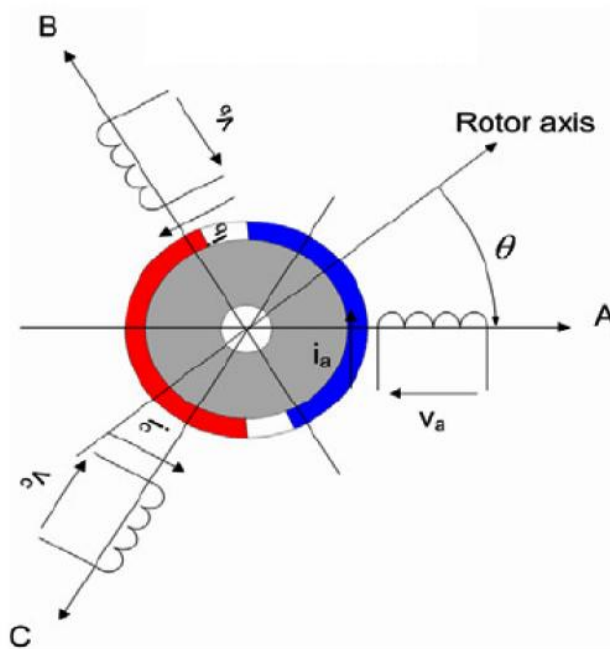


Fig. 3.3 Illustration of the rotor position angle θ_e .

To link the input voltages and currents of the inverter with those of the output, the power equality equation, $P_{in} = P_{out}$ is assumed at both sides.

From this, the inverter input current is:

$$i_{sk} = \frac{1}{V_s} (i_A \cdot V_{sA} + i_B \cdot V_{sB} + i_C \cdot V_{sC}) \quad (3.10)$$

where V_{sA} , V_{sB} and V_{sC} are the phase voltages that supply the motor.

The mechanical system shown schematically in Figure. 3.4 is defined by the following equations:

$$T_{em} = J_{eq} \frac{d\omega_m}{dt} + B \cdot \omega_m + T_L \quad (3.11)$$

where is $J_{eq} = J_M + J_L$ the equivalent moment of inertia, and J_M , J_L are the moments of inertia of the motor and load respectively, B -friction coefficient and T_L -load torque.

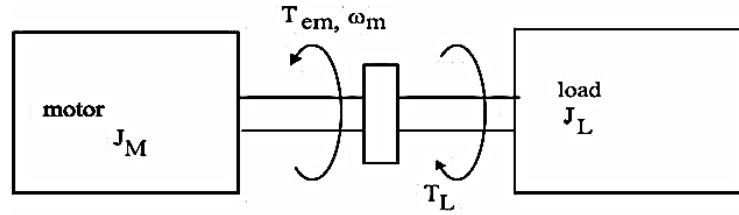


Fig. 3.4 Scheme of the mechanical system.

The electromagnetic torque for this 3-phase motor is dependent on the current (i), speed (ω_m) and electromotive force (e). The equation is:

$$T_{em} = \left(\frac{e_A \cdot i_A}{\omega_m} + \frac{e_B \cdot i_B}{\omega_m} + \frac{e_C \cdot i_C}{\omega_m} \right) = K_E (f_a(\phi_e) \cdot i_A + f_b(\phi_e) \cdot i_B + f_c(\phi_e) \cdot i_C) \quad (3.12)$$

Where:

$$\begin{aligned} f_a(\phi_e) &= \sin(\theta_e) \\ f_b(\phi_e) &= \sin\left(\theta_e - \frac{2}{3}\pi\right) \\ f_c(\phi_e) &= \sin\left(\theta_e - \frac{4}{3}\pi\right) \end{aligned} \quad (3.13)$$

Combining all the above equations, the system in state-space form is;

$$\dot{x} = Ax + Bu \quad (3.14)$$

$$x = [i_A \quad i_B \quad i_C \quad \omega_r \quad \theta_e] \quad (3.15)$$

$$A = \begin{bmatrix} -\frac{R_s}{L_s} & 0 & 0 & -\frac{K_E(f_a(\theta_e))}{L_s} & 0 \\ 0 & -\frac{R_s}{L_s} & 0 & -\frac{K_E(f_b(\theta_e))}{L_s} & 0 \\ 0 & 0 & -\frac{R_s}{L_s} & -\frac{K_E(f_c(\theta_e))}{L_s} & 0 \\ \frac{K_E(f_a(\theta_e))}{J} & \frac{K_E(f_b(\theta_e))}{J} & \frac{K_E(f_c(\theta_e))}{J} & -\frac{D}{J} & 0 \\ 0 & 0 & 0 & \frac{P}{2} & 0 \end{bmatrix} \quad (3.16)$$

$$B = \begin{bmatrix} -\frac{1}{L_s} & 0 & 0 & 0 \\ 0 & -\frac{1}{L_s} & 0 & 0 \\ 0 & 0 & -\frac{1}{L_s} & 0 \\ 0 & 0 & 0 & -\frac{1}{J} \\ 0 & 0 & 0 & 0 \end{bmatrix} \quad (3.17)$$

$$u = [v_A \quad v_B \quad v_C \quad T_L]^t \quad (3.18)$$

3.1.2 Simulation of motor dynamics

The simulation of the BLDC motor was done using the software package MATLAB / SIMULINK. For this purpose, the motor's block diagram was constructed, as shown in Figure.3.5. After running the simulation, the speed, torque, current, input and output power waveforms were recorded. Figures.(5.1-5.5) shows the start-up process of the motor. The inverter was supplied with a voltage of 380V DC and the motor loaded with a rated torque of 6.45 N·m. Phase A voltage is a square wave of 190V DC Figure 5.1. However, notice how the phase current is distorted from the square wave shape to something between a sine-wave and a square wave as shown in Figure 5.4. This is a result of the inductance effect. After the initial start-up, the speed oscillates around 16.4 rad /s. show Figure 5.2., the load torque is a constant straight line of 6.45 N·m Figure 5.3.The electromagnetic

torque depends on phase currents, constant K , speed and electromotive force (recall equation 3.12). Therefore, all these quantities affect the look of its waveform. After the start-up process, the motor reaches steady-state at around 0.25 seconds. Table 3.1 lists its parameters in this state.

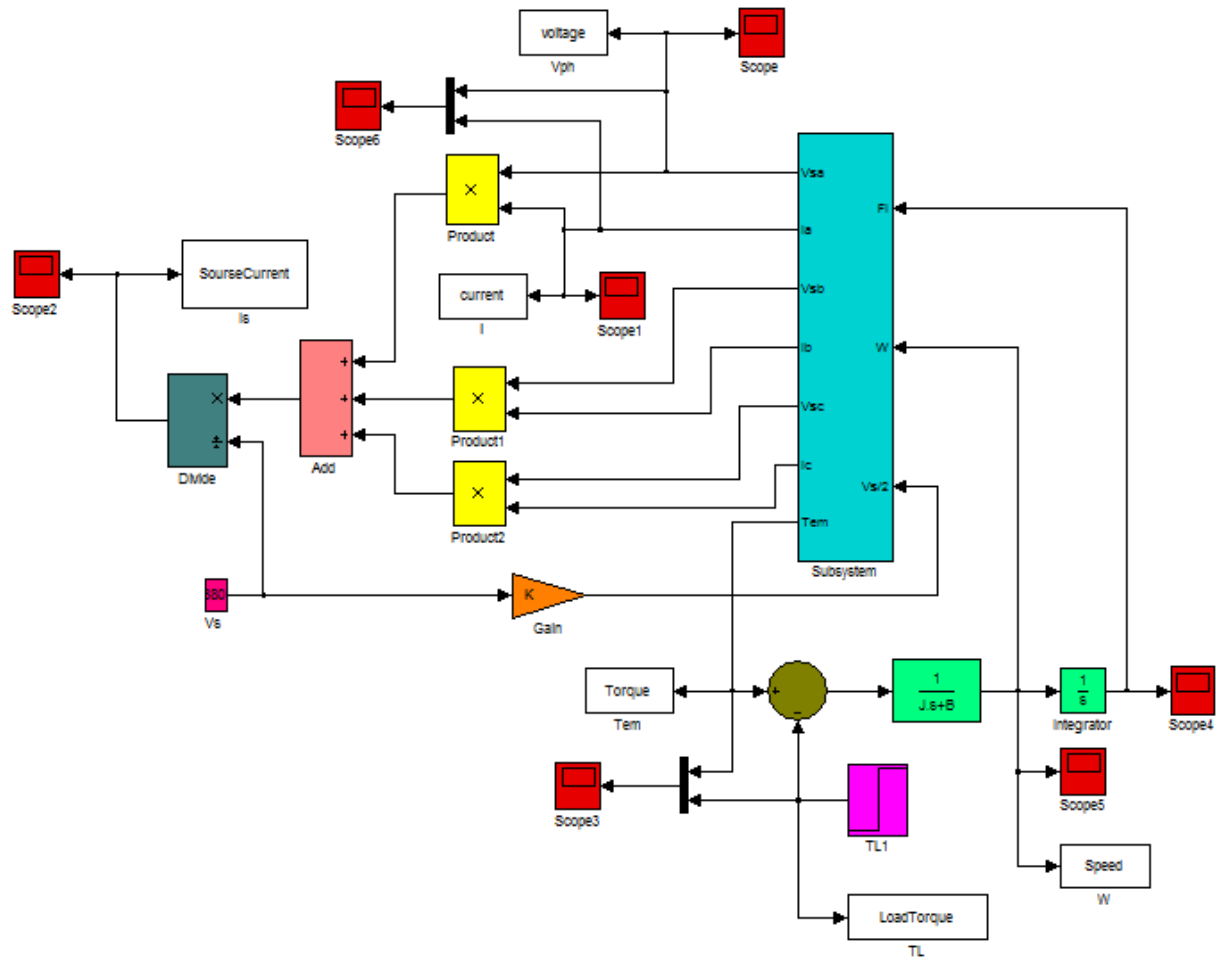


Fig. 3.5.a Block Diagram for BLDC motor operation.

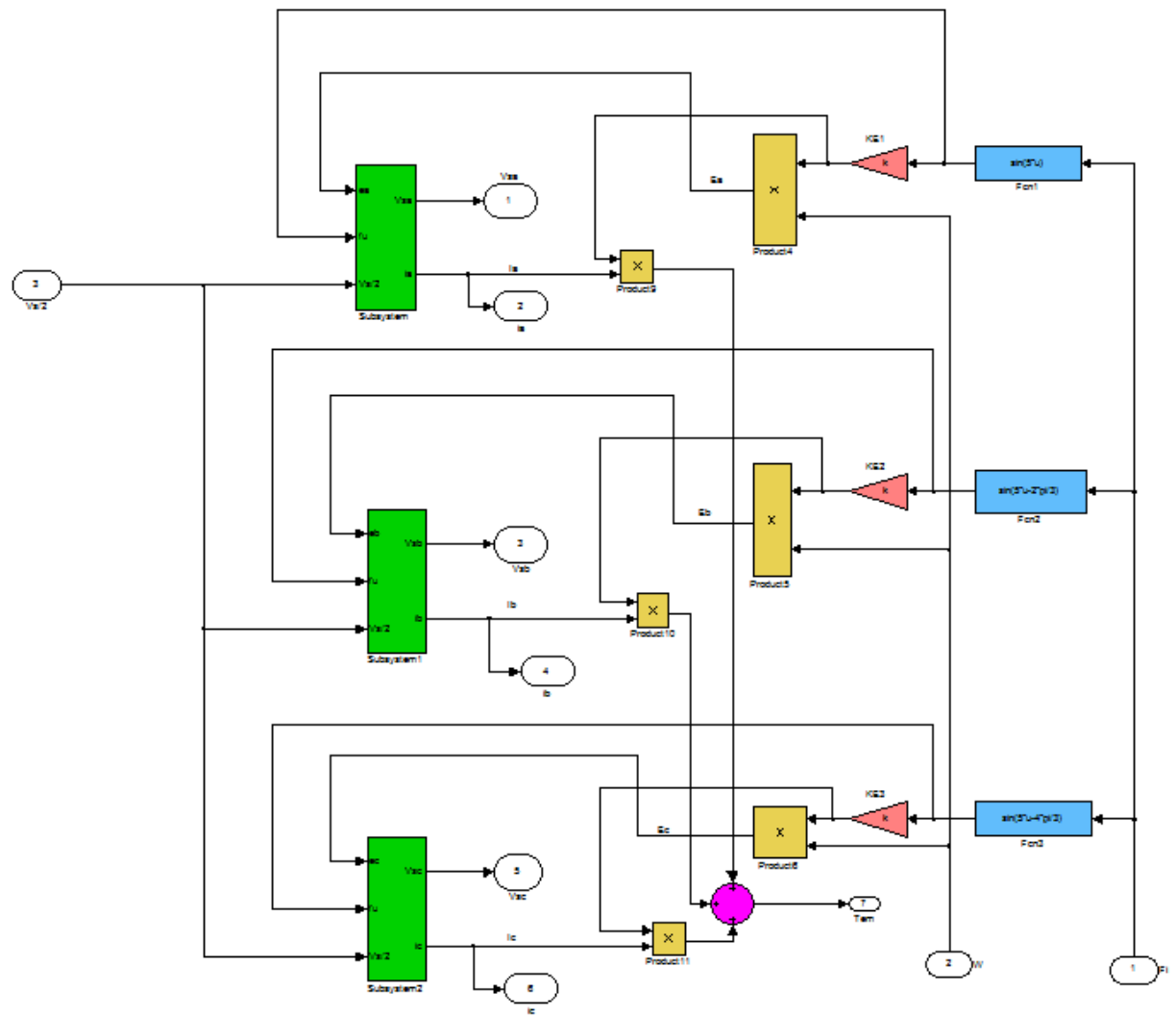


Fig. 3.5.b Block diagram behind subsystem shown in Fig.3.5.a

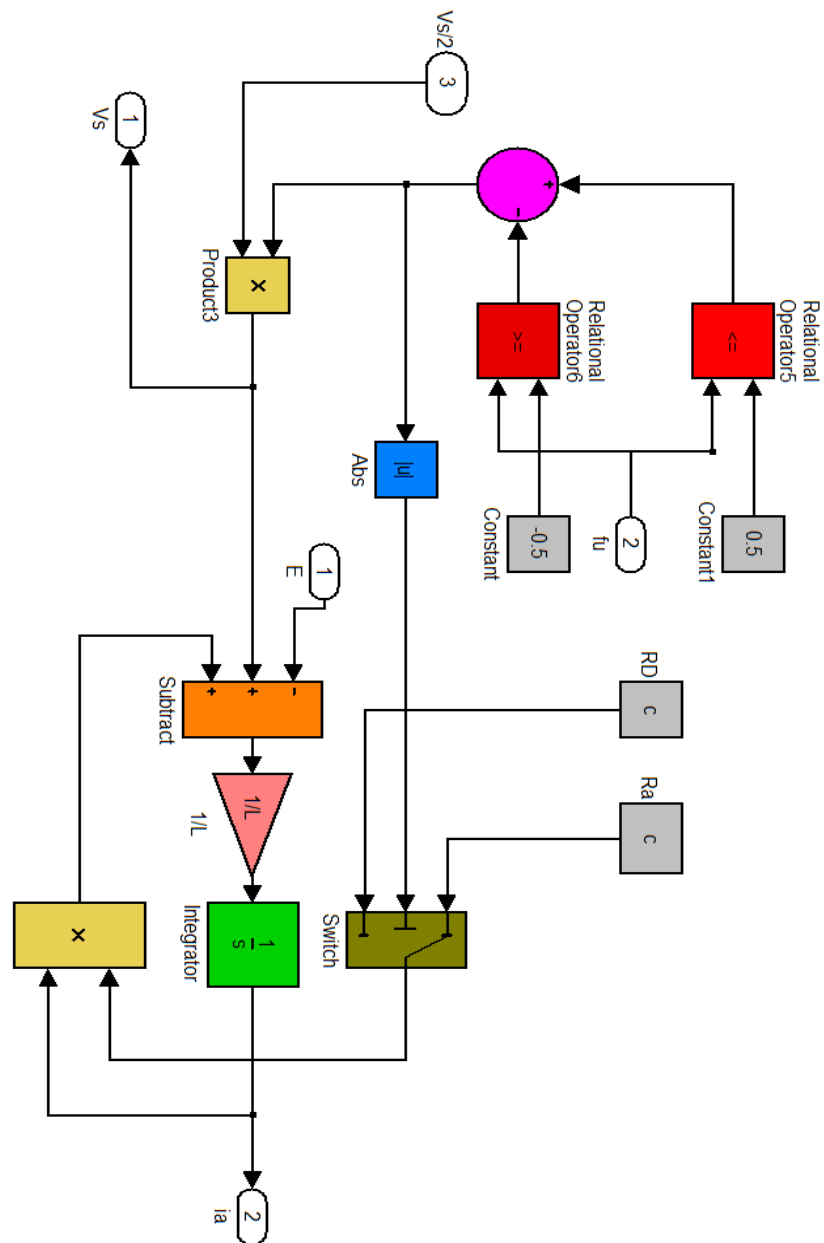


Fig. 3.5.c Block diagram behind subsystem shown in Fig.3.5.b

The motor efficiency was calculated as:

$$Eff(\%) = \frac{P_L}{P_{in}} \cdot 100 \quad (3.19)$$

where the load power on the motor shaft,

$$P_L = T_L \cdot \omega_m \quad (3.20)$$

and the input power of the inverter equal to the input power of the motor

$$P_{in} = V_s \cdot I_s \quad (3.21)$$

The above quantities were calculated as average values while the current as an *rms* value.

The waveforms were plotted using MATLAB/SIMULINK package.

Table 3.1 Electromechanical parameters of BLDC motor in steady-state.

Supply voltage	380 V DC
Output power	105.75 W
Input power	116.12 W
Load Torque	6.45 N.m
Efficiency	91.1 %
Electromagnetic Torque	6.5 N.m
Speed	16.4 rad/sec
Current	52.92A

3.2 BLDC motor performance in steady-state

The operation of the BLDC motor in steady-state may be regarded as that of a DC motor with commutators; hence the brush DC motor model may be used for analysis of the BLDC motor's performance in this state.

Section 3.2.1 explains the similarities and differences between these two types of motors.

3.2.1 BLDC Motor and Brush PM DC Motor: Similarities and Differences

The brush DC motor is excited by either field windings or permanent magnets. In both cases, they are placed on the stator. The armature winding, which is placed on the rotor, consists of a number of coils as shown in Figure.3.6. When the rotor turns, the current of the following coils that are approaching stationary brushes, are commutated. Due to the commutator, the resultant magnetic flux Φ_a produced by the coils is always perpendicular to the field flux although the current changes in the rotor coils [32]. As the name suggests, the brushless DC motor has no brushes and is electronically commutated with permanent magnets placed on the rotor.

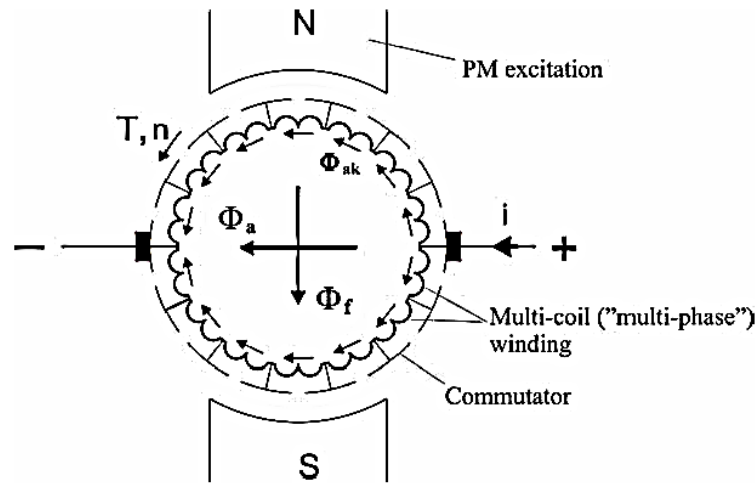


Fig. 3.6 Diagram of DC commutator motor, which visually explains its operation.

The commutator can be regarded as a mechanical inverter. This is seen clearly when the DC motor with three delta-connected coils or phases is considered as shown in Figure. 3.7a. At a particular time instant t_1 , coils A, B and C are supplied, generating the resultant flux, Φ_a perpendicular to the field flux (see Figure. 3.7).

The same position of Φ_a can be achieved if the coils are supplied from a DC source through a 3-phase inverter as shown in Figure.3.7b. This operation would be equal to that seen in the brushless DC motor.

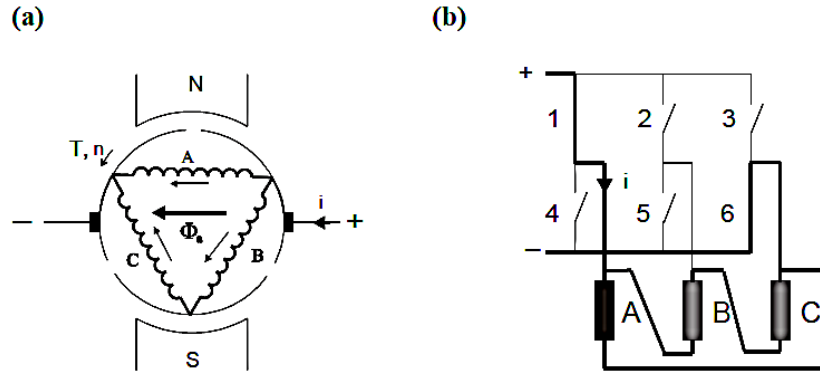


Fig. 3.7 Diagram of a DC motor with 3 coils or phases in the armature: (a) coils commutated by mechanical commutator, (b) coils commutated by electronic commutator (inverter).

In the case of star (wye) connected windings shown in Figure. 3.8, two coils are energized at any one time. Again, this can also be achieved by using a 3-phase inverter as shown in Figure. 3.8b. The above mentioned figure shows the position of the motor and coils energized by commutator and inverter at three different time instants. No changes are observed in the resulting flux position with respect to the stationary field. The only minute changes observed in the position of the flux, Φ_a are between two subsequent winding commutations (see Figure. 3.8) [32].

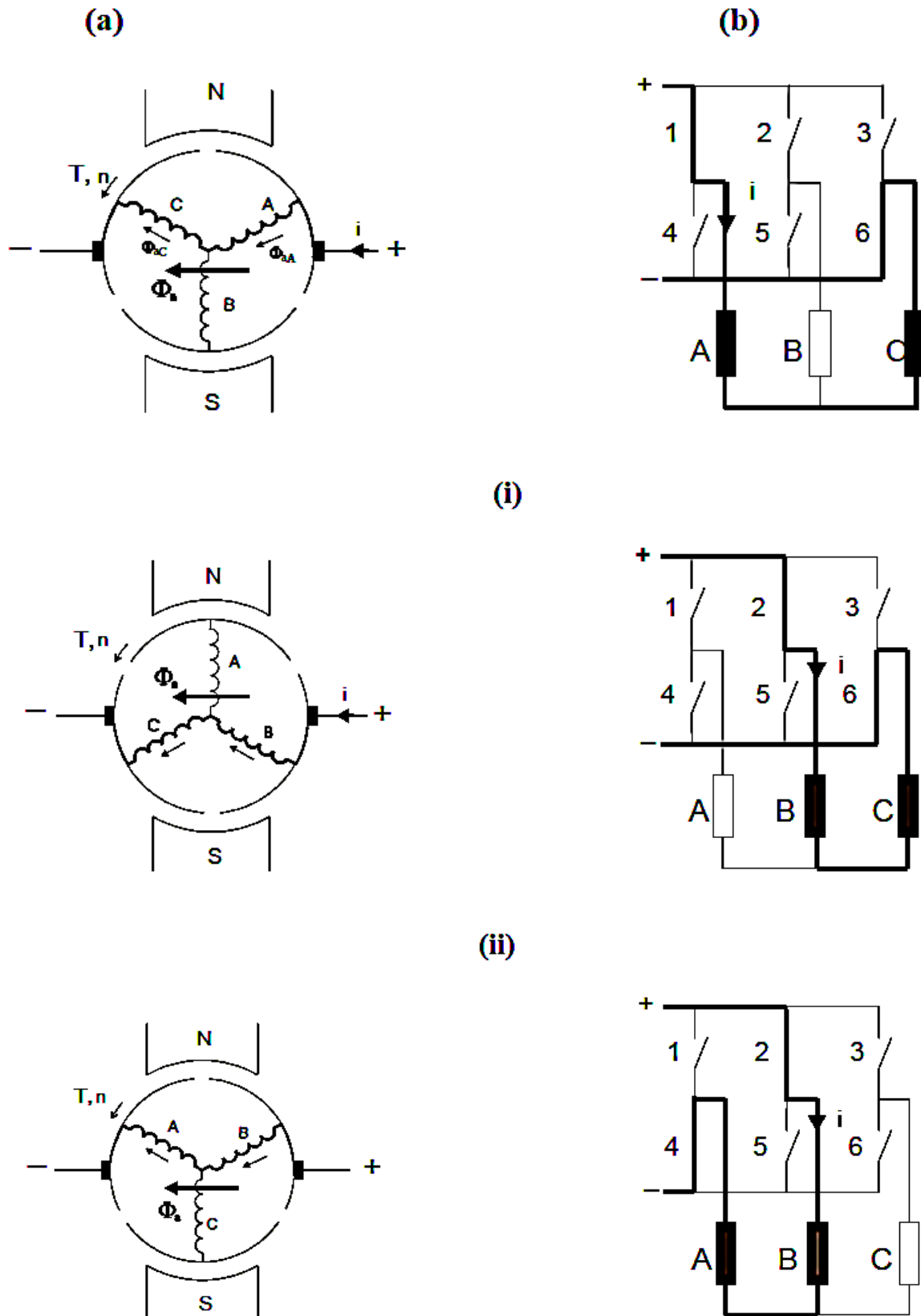


Fig. 3.8 Scheme of DC motor with 3-phase star connected winding: (a) with mechanical commutator (winding placed on rotor), (b) with inverter (winding is on stator and magnetic poles rotate).

Due to the change in position, i.e. $\Delta\theta = 60^\circ$ (see Figure. 3.9c), the resultant relations of stationary flux Φ_f and flux Φ_a , (also known as electromagnetic torque T_{em}) changes with time, producing some ripples in the torque (see Figure. 3.10). The more phases we have, the smoother the torque waveform will be [24]. Another important difference to be noted between the brush and brushless DC motors is as follows.

In a DC commutators motor, the armature winding, which is mechanically commutated, is placed on the rotor, while the field windings or permanent magnets are placed on the stator. In the BLDC motor, the reverse is true. This is because the armature winding in the former is a self commutated winding caused by the rotating commutator. On the other hand, the winding of the latter can be commutated by a stationary electronic inverter. The moment of commutation in the conventional DC motor is determined by the position of the coil with respect to the stationary brushes. In BLDC motors, it is determined by the position of the sensor signal. This, in turn, means that BLDC motors cannot operate without position sensors.

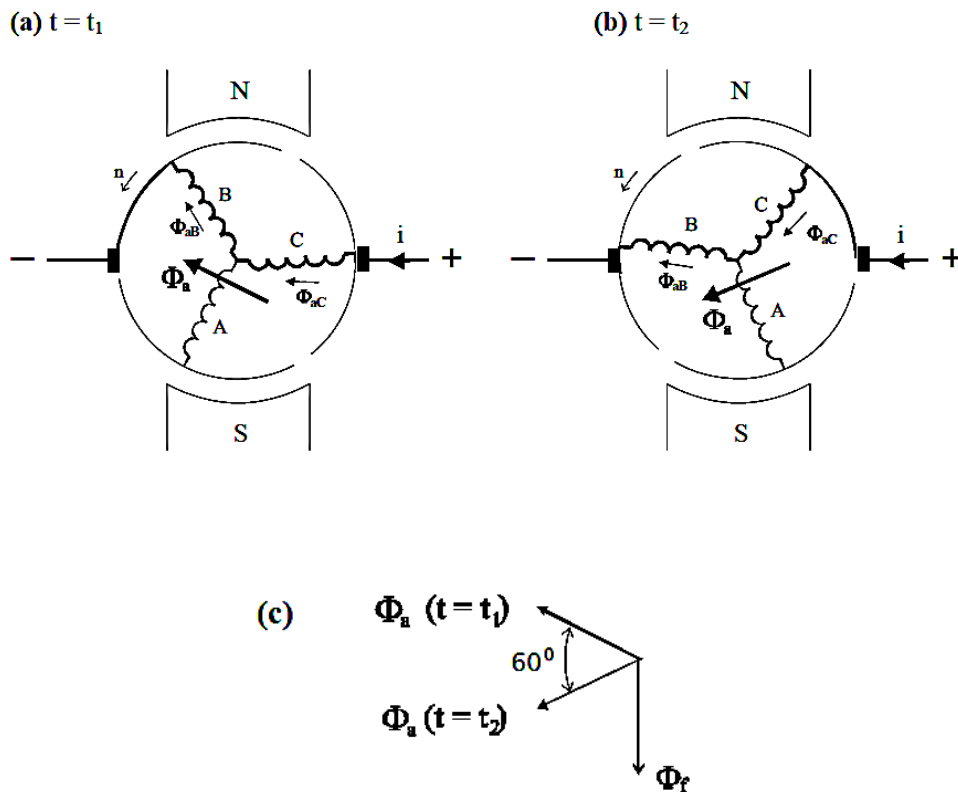


Fig. 3.9 Rotor positions at two subsequent instants: (a) at time t_1 , (b) at time t_2 and (c) mutual positions of the armature flux with respect to the field flux position at the two time instants

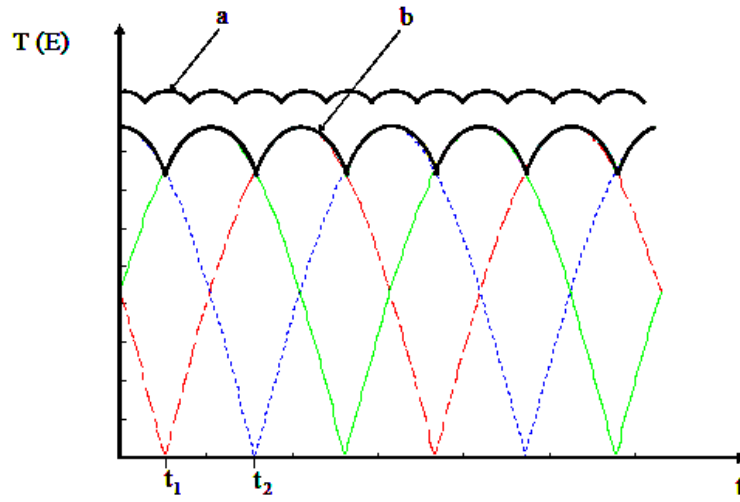


Fig. 3.10 The torque (T) and electromotive force (E) waveforms: (a) with more phases, (b) with a 3-phase motor.

3.2.2 Steady-state model of BLDC motor

When the BLDC motor is in steady-state, it is simply operating as normal brush DC motor. The model for this brush model, without any inductance, is shown in Figure. 3.11.

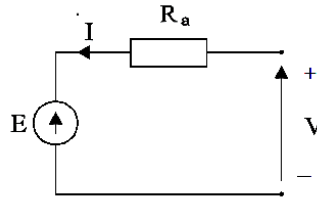


Fig. 3.11 Brush DC motor armature circuit diagram.

The equations that define this model are as follows.

$$T = K \cdot I - T_0 \quad (3.22)$$

$$E = K \cdot \omega_m \quad (3.23)$$

$$V = E + R \cdot I \quad (3.24)$$

where:

- torque on the shaft, T
- line-to-line electromotive force, E
- rotor angular speed, ω_m
- static friction torque, T_0
- constant, K
- armature resistance, R
- average armature current I
- source voltage, V

From equations 3.23 and 3.24 we obtain,

$$\omega_m = \frac{V - R \cdot I}{K} \quad (3.25)$$

Or:

$$\omega_m = \frac{V}{K} - R \left(\frac{T + T_o}{K^2} \right) \quad (3.26)$$

In the brush DC motor, at any time only a small part of the winding is commutated during the motor operation (see Figure. 3.8) and the inductance across this part does not affect the motor performance much. However, one negative effect occur by this commutator is that some sparks occur between the brushes and commutator.

In the BLDC motor with a 3-phase winding, at any time one third of the winding is commutated Figure. 3.9 and the current in the commutated part control the motor performance. The higher the motor speed, the shorter the commutation period and the faster the current changes in the commutated part of the winding. This change contributes to the relatively high voltage drop across this part, hence the inductance of this part (according to the equation $V_{com} = L_{com} \frac{di_a}{dt}$) may significantly influence the motor performance, in particular the speed-torque characteristics [24]. This means the winding inductance in the equivalent motor circuit should be considered.

If the motor inductance has to be included, then the circuit diagram will appear as follows.

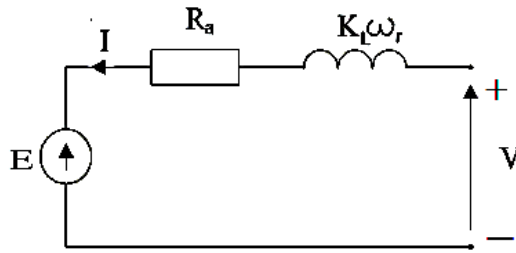


Fig. 3.12 Brush motor's equivalent circuit model with armature inductance

The voltage equation then changes to include the inductance. It is,

$$V = E + R \cdot I + K_L \cdot \Omega_m \cdot I^n \quad (3.27)$$

where K_L is a constant and the current is to the power n .

From equation (3.23) and (3.27), the speed is,

$$\omega_m = \frac{V - R \cdot I}{K_E + K_L \cdot I^n} \quad (3.28)$$

or expressed in terms of electromagnetic torque (recall torque $T_{em} = K \cdot I$),

$$\omega_m = \frac{V - R \cdot \frac{T_{em}}{K}}{K_E + K_L \cdot \left(\frac{T_{em}}{K}\right)^n} \quad (3.29)$$

All other equations remain the same as previously.

3.2.3 Performance characteristics in steady-state

To find out how the analyzed BLDC motor performs in steady-state and to determine which brush motor model may be used for analysis, the motor's electromechanical characteristics are first determined from the dynamic model. The characteristic that reflects the control of the inductance is speed-torque curve. This along with the current-torque curve is drawn in Figure.5.6. According to equation (3.26), the motor model with no inductance Figure. 3.11 gives the speed-torque characteristic as a straight line. For the motor model with inductance Figure. 3.12, the speed-torque characteristic according to equation (3.27) is non-linear (Figure.5.6).

On the basis of these characteristics, the particular steady-state model was selected and its parameters determined. The brush model with no inductance was chosen because, as Figure.5.6 shows, the speed-torque characteristics be similar to those of that model. Also, the straight line characteristic shows that inductance played no part in affecting its shape; otherwise it would have appeared as a curved line. The constant K_E was calculated using equation (3.22) or simply looking at Figure.5.6, calculating the slope of the current curve and taking its inverse. The static friction torque constant T_O was determined.

With the use of equations 3.22-3.28, the performance characteristics were calculated applying the MATLAB software (see m-file *brush1* in Appendix A). The characteristics were plotted in Figure. 5.7.

Figure. 5.7. shows the characteristics of both models, i.e. The selected brush DC motor and the BLDC motor determined after it reaches steady state. By plotting both characteristics, we are able to see how the BLDC motor (after it reached steady-state) performed compared to the brush motor. The brush motor's characteristics are those represented by dashed lines.

From these characteristics, we see that the brush motor model describes the BLDC motor in steady-state relatively well.

About the particular characteristic, as the torque increases, the motor slows down and hence the speed decreases.

The mechanical power was determined from the equation:

$$P_m = \omega_m \cdot T_L \quad (3.30)$$

The motor efficiency was calculated as follows:

$$Eff(\%) = \left(\frac{P_m}{P_{in}} \right) \cdot 100 \quad (3.31)$$

Where:

$$P_{in} = V \cdot I \quad (3.32)$$

Looking at Figure. 5.7, it is observed that at high torques, the speed-torque relation from the dynamic model becomes constant. This means that the inductance does have some effect (relatively small) on the motor performance. Since within the range of the rated torque (6.45N.m) both characteristics equal, the model without inductance can be applied to the analysis of this particular BLDC motor.

3.3 BLDC motor speed control

3.3.1. Simulink model of speed control of BLDC motor

This section details MATLAB simulink model of Permanent magnet brushless DC motor and its speed control using PI controller.

The Figure. 3.13 shows the Simulation model of BLDC motor with PI controller based on the mathematical modeling which is manually tuned to obtain the desired motor speed.

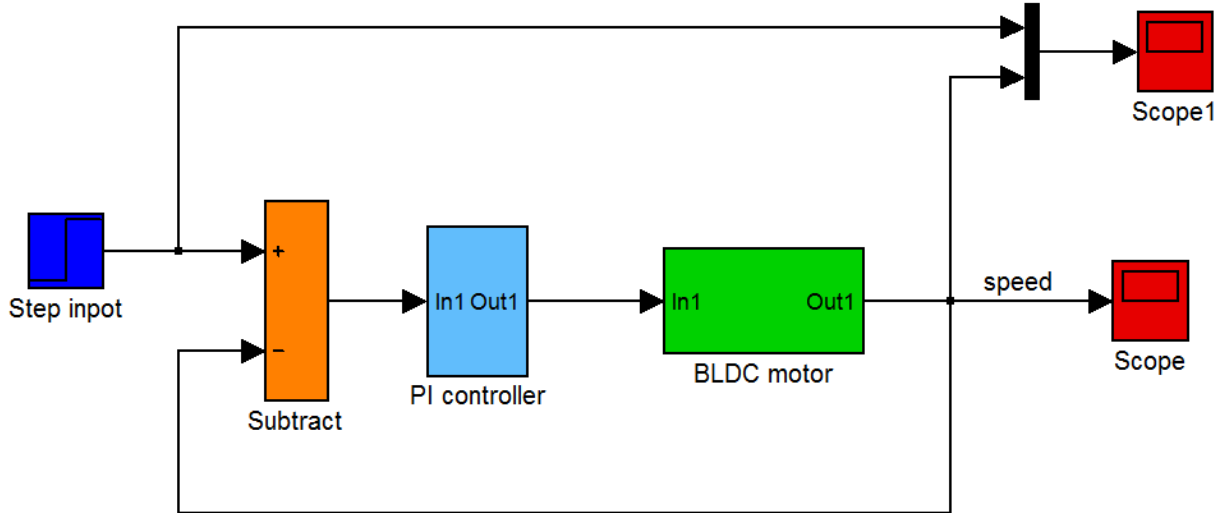


Fig. 3.13 Simulation model of BLDC motor with controller

The MATLAB simulink model of PM BLDC Motor has been developed as shown in Figure.3.5a,b,c.

The input of PI controller is the deference between the desired speed (ω_{ref}) and the motor speed (ω_m), and the output of the PI controller is applied to the BLDC Motor.

3.3.2. PI controller modeling and simulation

PI or two-term controllers are the most widely used controllers in industries today. PI control is applied in speed, current, and position control. The structure of PI controller will be as shown in Figure.3.14, and Figure 3.15. shows the MATLAB simulation of PI controller.

The expression of PI control in time domain is:

$$u(t) = k_p[R(t) - Y(t)] + k_i \int_0^t [R(t) - Y(t)] dt \quad (3.33)$$

$$u(t) = k_p e(t) + k_i \int_0^t e(t) dt \quad (3.34)$$

Where:

$u(t)$ - controllers output signal

$e(t)$ - controllers input error signal

k_p - proportional control gain

k_i - integral control gain

The Laplace transform of equation (3.33) .

$$U(s) = k_p(R(s) - Y(s)) + \frac{k_i}{s}(R(s) - Y(s)) \quad (3.35)$$

$$U(s) = k_p E(s) + \frac{k_i}{s} E(s) \quad (3.36)$$

$$U(s) = E(s) \left(k_p + \frac{k_i}{s} \right) \quad (3.37)$$

$$D(s) = \frac{U(s)}{E(s)} \quad (3.38)$$

$$D(s) = k_p + \frac{k_i}{s} \quad (3.39)$$

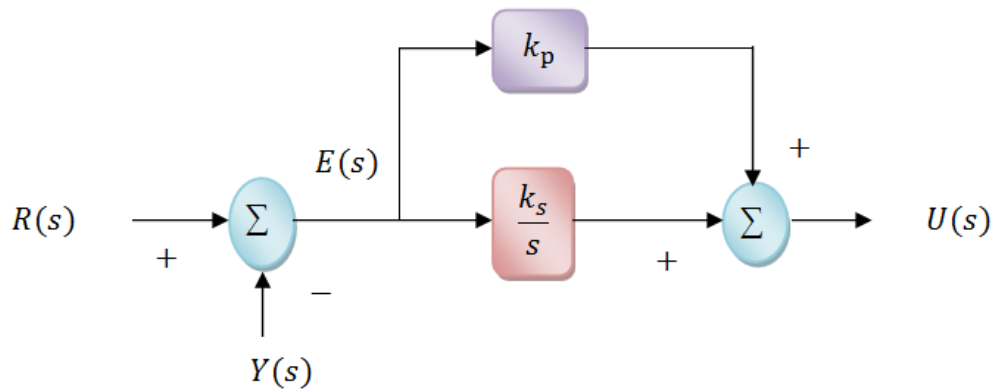


Fig.3.14. Structure of PI controller

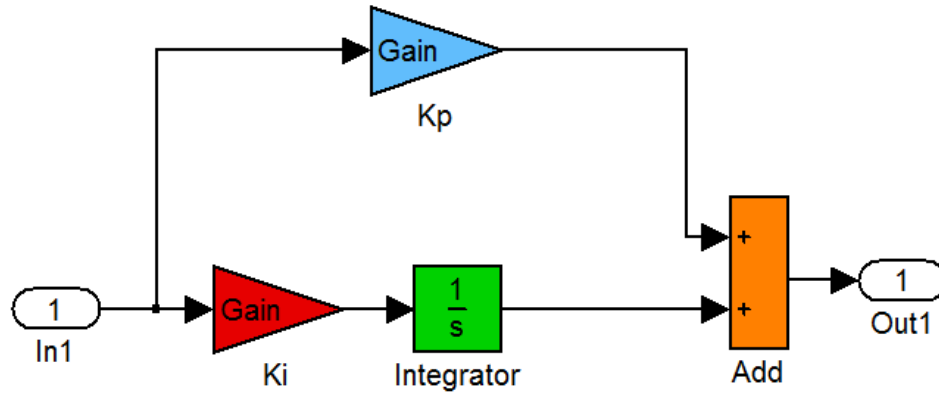


Fig 3.15. The MATLAB simulation of PI controller.

3.3.3. Tuning of PI controller and the results obtained

As mentioned in section (2.5.3) the system must remain online, the tuning method is to first set K_i and K_d values to zero. Increase the K_p until the output of the loop oscillates. Then increase K_i until any offset (overshoot, steady-state error, stability) is correct in sufficient time for the process. However, too much K_i will cause instability. Finally, increase K_d , if required , until the loop is acceptably quick to reach its reference after a load disturbance.

Figure.3.13 above shows the speed control system of BLDC motor with PI controller which is manually tuned to obtained the desired rotor speed. The PI values at which the speed is obtained at $K_p = 12$, $K_i = 6$ and. then graphical rotor speed with time after controller is shown in Figure 5.8. and Figure 5.9. show the improvement in the shape of the phase A current with controller .

4. Modeling , simulation and control of PMSM

This chapter presents the dynamic modeling and simulation of PMSM, focuses on the abc model and the second system known as dq model, discussed the synchronous motor in steady state model, so the permanent magnet synchronous motor control with dq model is present.

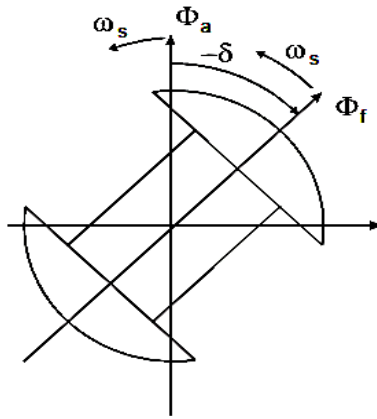
4.1 Dynamics of the synchronous motor

In this section we study the dynamic **abc** model of PMSM, and simulate the motor dynamics

4.1.1 Dynamic model of the synchronous motor

The 3-phase armature winding of the synchronous motor is connected to a 3-phase ac supply. The stator currents produce a rotating magnetic flux, as in 3-phase induction machines. The magnetic flux of the PM's is steady with respect to the rotor. To produce torque, these two magnetic fluxes cannot move with respect to one another. This means that the rotor should rotate with the same speed as the rotating flux produced by the stator Figure. 4.1. The rotor follows the stator rotating field by an angle $-\delta$ Figure. 4.1. The motor being supplied with rated frequency does not develop any torque at zero speed. For the motor to operate, the rotor should first reach a synchronous speed. This can be done in three ways: by driving the rotor to the synchronous speed (with an external machine), by starting the rotor (which has to be equipped with a starting cage) as in induction motors, or by supplying the stator winding with variable frequency, beginning from zero to the rated frequency. The last method can be used if the winding is supplied from a frequency converter.

In the experiment (to be discussed in detail later on in this chapter), a frequency converter was used (recall Figure.2.14) to try and bring the motor from standstill to its desired speed. The equivalent circuit for the dynamic model of the synchronous motor can be represented by two systems; namely the phase-a, phase-b, and phase-c (ABC) system, which involves three phases, and the direct polar axis, and quadrature or interpolar axis (dq) system which involves two phases (see details in section 4.2). The ABC system's model is very similar to that of the BLDC motor. Figure. 4.2 shows this ABC equivalent circuit diagram.



Φ_f -magnetic flux due to PM excitation .

Φ_a -flux due to stator current.

ω_s -synchronous speed.

Fig. 4.1 Explanation of synchronous motor operation.

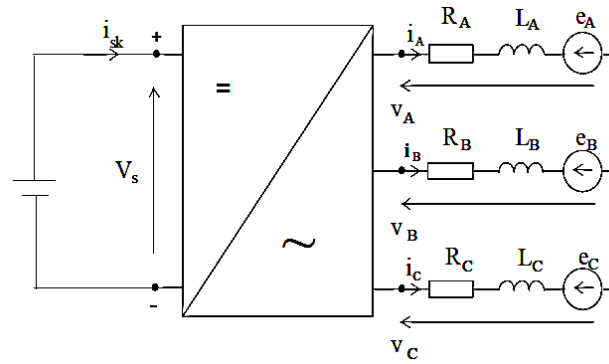


Fig. 4.2.a ABC equivalent circuit diagram.

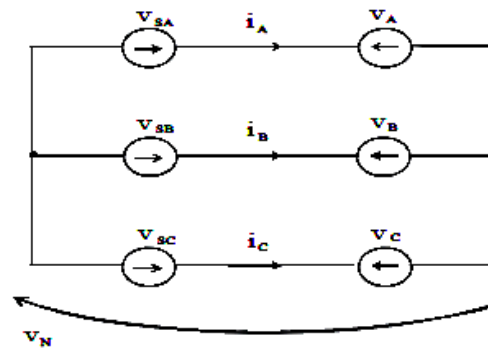


Fig. 4.2.b Motor side of Fig. 4.2.a.

The motion and voltage equations that define the model are the same as those of the BLDC motor (see equations 3.1 to 3.18). The only difference between the two models is the type of supply voltage (BLDC motor was supplied by dc supply, and PMSM was supplied by AC supply).

4.1.2 Simulation of the motor dynamics

The simulation of the synchronous motor was done using the software package MATLAB / SIMULINK. For this purpose, the motor's block diagram was constructed, as shown in Figure.4.3. Also keep in mind that section 4.3.1 offers another block diagram (dq system) to simulate the synchronous motor's performance. After running the simulation, the speed, torque, current, input and output power waveforms were recorded and analyzed using MATLAB / SIMULINK.

Figures. (5.10-5.16). shows the waveforms of the electrical and mechanical quantities after the stator was supplied with a sinusoidal voltage of 220V ,60Hz frequency and an initial speed oscillates around of 12.04 rad/s (the motor's rated speed). The motor was loaded with a rated torque of 6.45 N.m.

The current and voltage have a sinusoidal shape because, as Figure. 4.3 shows, the input was also sinusoidal. The motor efficiency is calculated in the same way as that of the BLDC motor (see equation 3.19). After the start-up process, the motor reaches steady-state at around 0.75 seconds.

After the motor reached steady-state, the electromechanical parameters were recorded in Table 4.1. As mentioned earlier, the angle δ refers to the common position of the supply voltage and electromotive force, for the same phase. If the stator resistance R_a is negligible, this angle is equal to the angle between the 'd' axis of the rotor and the stator flux.

During this motoring operation, this stator flux Φ_a pulls the rotor. The phasor diagram with the angle δ (and other quantities) is shown in Figure. 4.4. (E_f, i_A, V_A) .

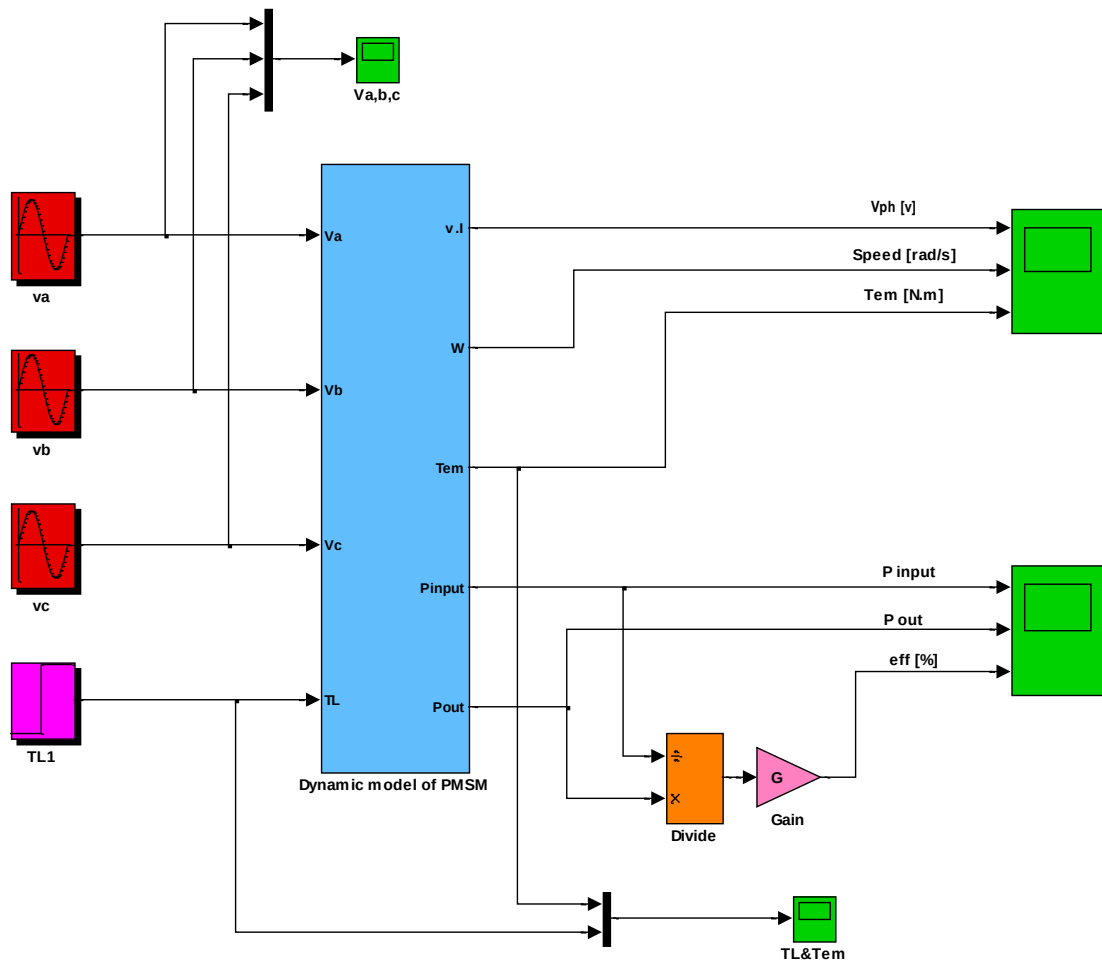


Fig. 4.3.a Block diagram of synchronous motor operation

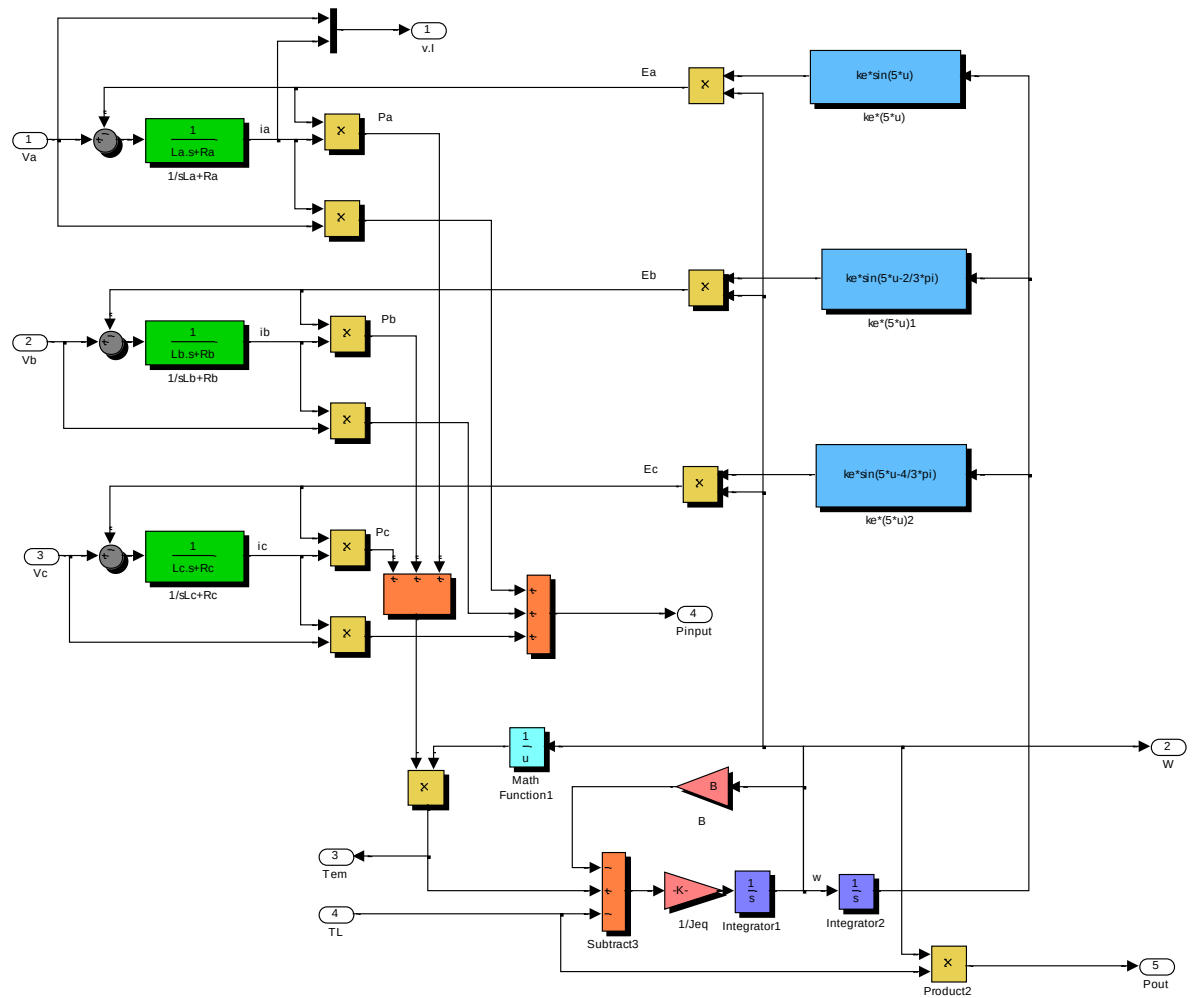


Fig. 4.3.b Block diagram behind subsystem shown in Fig.4.3 a.

Table 4.1 Electromechanical parameters of synchronous motor performed at rated conditions.

Supply voltage (Vs)	220 V AC
Output power (P_{out})	77.68 W
Input power (P_{input})	85.19 W
Load Torque (T_L)	6.45 N.m
Efficiency (<i>E_{ff}</i>)	91.18 %
Electromagnetic Torque (T_{em})	6.5 N.m
Motor Speed (ω_m)	12.04 rad/sec
Current (I)	48.98A

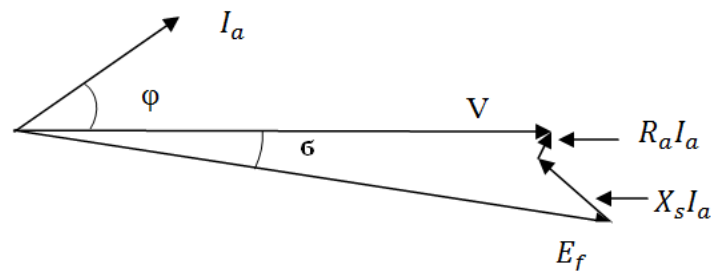


Fig.4.4.Phasor diagram showing angles, voltage and current.

4.2 Steady-state model of the synchronous motor

Just like the BLDC motor, the synchronous motor can be analyzed in steady-state without using a dynamic model. Discussed in this section, is its performance in this state; the motor is analyzed on the basis of the steady-state equivalent circuit and then its results compared to those obtained from the dynamic model.

4.2.1 Equivalent circuit model

The equivalent circuit model will be derived on a per-phase basis. The PM's produce the flux ϕ_f . The current I_a in the stator produces flux ϕ_a . Part of this, known as the leakage flux ϕ_{al} , does not link the rotor. A major part known as the armature reaction flux ϕ_{ar} is linked with the rotor. The resultant air gap flux ϕ is therefore the sum of two component fluxes ϕ_f and ϕ_a . Each flux component (and the resultant flux) induces a voltage component in the stator winding:

$$\phi_f \rightarrow E_f, \phi_{ar} \rightarrow E_{ar}, \phi_{al} \rightarrow E_{al}, \phi \rightarrow E$$

The excitation voltage E_f can be determined from the open circuit test. However, the armature reaction voltage, E_{ar} and the leakage flux voltage E_{al} depend on the armature current. Therefore they can be represented as voltage drops across the reactance's : X_{ar} —reactance of armature reaction and X_{al} — leakage reactance [34]. This is shown in the equivalent circuit in Figure. 4.5. The relationship between these voltages is:

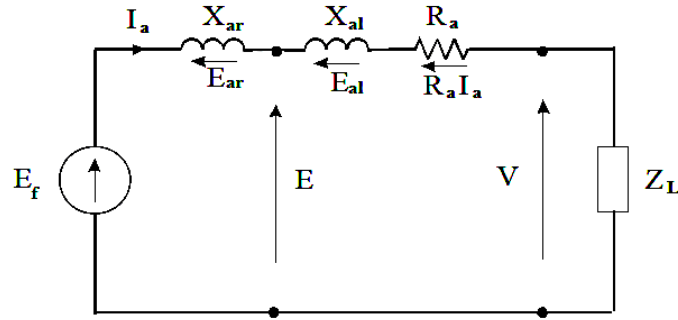
$$\underline{E} = \underline{E}_f - \underline{E}_{ar} \quad (4.10)$$

$$\underline{E} = \underline{E}_f - jX_{ar}I_a \quad (4.11)$$

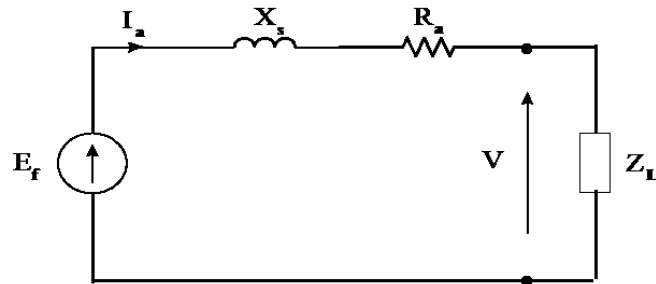
The voltage equation for the complete circuit (Figure.4.5.a) is:

$$\underline{V} = \underline{E}_f - \underline{R}_a I_a - jX_{al} I_a - jX_{ar} I_a \quad (4.12)$$

(a)



(b)



(c)

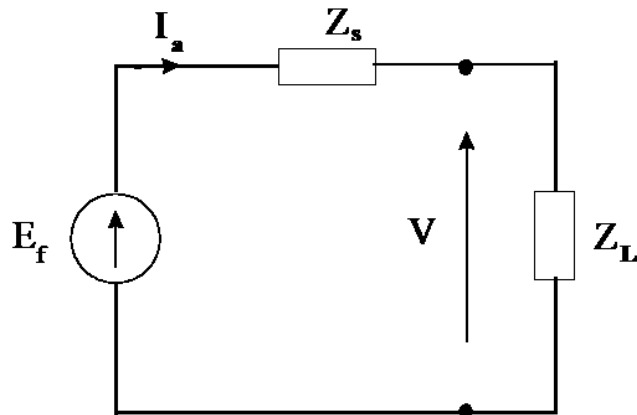


Fig. 4.5. Synchronous machine equivalent circuit: (a) armature reaction reactance X_{ar} , armature leakage reactance X_{al} , (b) synchronous reactance X_s , (c) armature impedance Z_s .

or (Figure.4.5.b)

$$\underline{V} = \underline{E}_f - \underline{R}_a \underline{I}_a - jX_s \underline{I}_a \quad (4.13)$$

or (Figure.4.5.c)

$$\underline{V} = \underline{E}_f - \underline{Z}_s \underline{I}_a \quad (4.14)$$

where:

$X_s = X_{ar} + X_{al}$ - synchronous reactance

$Z_s = R_a + jX_s$ - synchronous impedance

4.2.2 Power and torque characteristics

The complex power at the stator winding terminals is:

$$S = mVI_a^* \quad (4.15)$$

The stator current from the equivalent circuit (Figure.4.5.c):

$$I_a^* = \left(\frac{E_f - V}{Z_s} \right)^* = \frac{E_f^*}{Z_s^*} - \frac{V^*}{Z_s^*}$$

$$I_a^* = \frac{|E_f| \angle -\delta}{|Z_s| \angle -\varphi_s} - \frac{|V| \angle 0}{|Z_s| \angle -\varphi_s} \quad (4.16)$$

$$I_a^* = \frac{|E_f|}{|Z_s|} \angle (\varphi_s - \delta) - \frac{|V|}{|Z_s|} \angle \varphi_s$$

$$S = m \left(\frac{|V||E_f|}{|Z_s|} \angle (\varphi_s - \delta) - \frac{|V|^2}{|Z_s|} \angle \varphi_s \right) \quad (4.17)$$

The real power P and reactive power Q are:

$$P = m \left(\frac{|V||E_f|}{|Z_s|} \cos(\varphi_s - \delta) - \frac{|V|^2}{|Z_s|} \cos \varphi_s \right) \quad (4.18)$$

$$Q = m \left(\frac{|V||E_f|}{|Z_s|} \sin(\varphi_s - \delta) - \frac{|V|^2}{|Z_s|} \sin \varphi_s \right) \quad (4.19)$$

In large synchronous machines $R_a \ll X_s$, thus $Z_s \ll X_s$ and $\theta_s = 90^\circ$. From the above equations:

$$P = m \frac{|V||E_f|}{|Z_s|} \sin(\delta) \quad (4.20)$$

$$Q = m \left(\frac{|V||E_f|}{|Z_s|} \cos(\delta) - \frac{|V|^2}{|Z_s|} \right) \quad (4.21)$$

Equation 4.11 can be directly derived from the phasor diagram drawn for $R_a = 0$ (see Figure. 4.6)

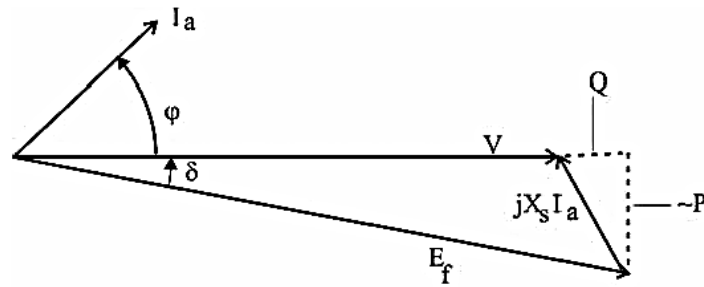


Fig. 4.6. Phasor diagram drawn for $R_a = 0$.

In general, the active power

$$P = mVI_a \cos \varphi \quad (4.22)$$

and reactive power

$$Q = mVI_a \sin \varphi \quad (4.23)$$

From the phasor diagram (Figure. 4.6),

$$X_s I_a \cos \varphi = E_f \sin \delta, \text{ and} \quad (4.24)$$

$$I_a \cos \varphi = \frac{E_f}{X_s} \sin \delta \quad (4.25)$$

Combining Equations 4.16 and 4.13, we get:

$$P = m \frac{V E_f}{X_s} \sin \delta \quad (4.26)$$

Because the stator losses are neglected in this analysis, the power developed at the terminals is also the air gap power. The electromagnetic torque developed by the machine is:

$$T = \frac{P}{\omega_s} \quad (4.27)$$

Inserting Equation 4.18 into 4.19:

$$T = \frac{m V E_f}{\omega_s X_s} \sin \delta \quad (4.28)$$

where m is the number of phases.

The torque-power angle characteristic drawn for the motor analyzed in this project is shown in Figure.4.7. The torque was calculated for the following data:

$$m = 3, P = 6$$

$$V_{rms} = \frac{1}{\sqrt{2}} V_{max} = 0.707 \times 220 = 155.54 \text{ V}$$

$$\omega_s = \frac{2\pi f}{P/2} = 20 \frac{\text{rad}}{\text{s}}$$

$$E_f = \frac{K_E}{\sqrt{2}} \omega_s = 0.707 \times 0.148 \times 20 = 2.093 \text{ V}$$

$$X_s = 2\pi f L_s = 2.61 \Omega$$

The calculations were done in MATLAB using m-file ***torque-delta*** (see Appendix B) as bellow. The maximum torque, known also as the pull-out torque is at $\delta = 90^\circ$. The machine will lose synchronism if $\delta > 90^\circ$. The pull-out torque can be increased by increasing the excitation flux Φ_f .

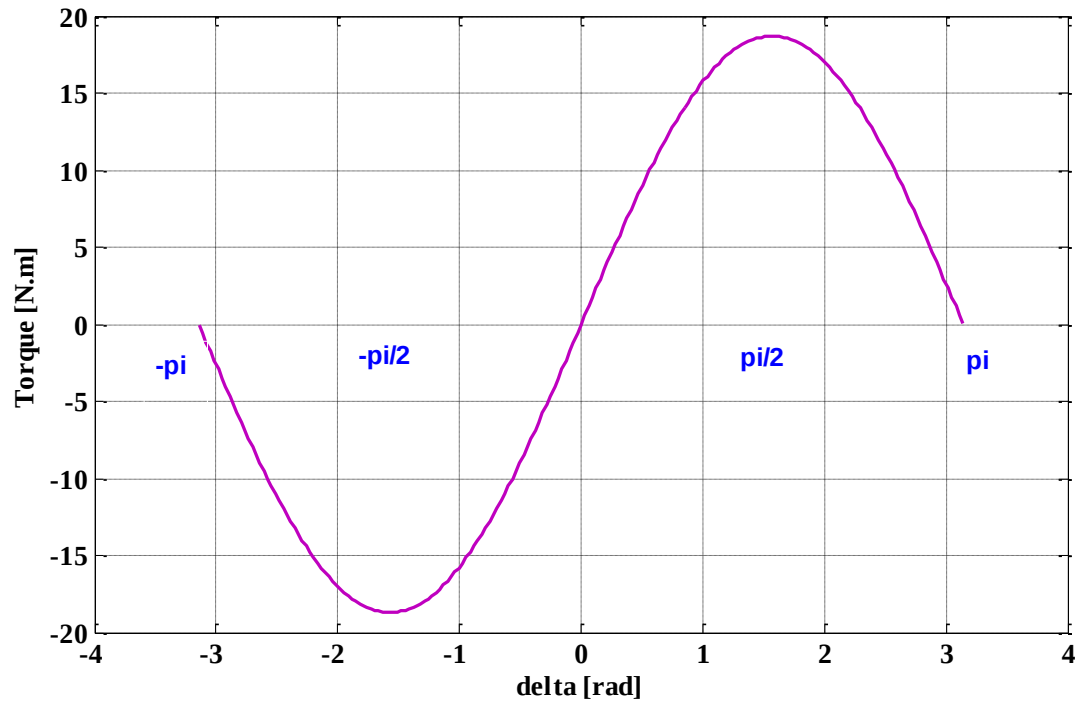


Fig. 4.7. Torque-power angle characteristic.

4.2.3 Performance characteristics of the motor in steady-state

To determine the electromechanical characteristics of the synchronous motor, calculations were carried out by means of MATLAB (see m-file *synchronousplot* in Appendix B), using the equations in sections 4.2.1 and 4.2.2. The parameters of the equivalent circuit for a supply frequency of 60 HZ are shown in Table 4.2.

Table 4.2 Parameters of the equivalent circuit

Supply voltage (Vs)	220 V AC
emf (E_f)	2.093 V AC
Synchronous reactance (Xs)	2.6 Ω
Armature resistance (Ra)	3.59 Ω

The performance characteristics are presented in Figure.5.17. The efficiency of the motor was calculated from the equation:

$$Eff = \frac{P_{out}}{P_{in}} \cdot 100\% \quad (4.29)$$

Where:

$\Delta P_{out} = P - \Delta P_m$, mechanical power loss.

$\Delta P_m = B \cdot \omega_s$ and $P_{in} = R_e(3 \cdot V \cdot I^*)$.

To verify the dynamic and steady-state models to some level, the simulation and calculation results obtained from both models respectively, are plotted on the same graph with the results from the dynamic model represented in form of points (see Figure. 5.17).

4.3 .Permanent magnet synchronous motor control

The complete drive scheme for the PMSM is shown in Figure.4.8. The motor voltage is being supplied through a three phase inverter. The PMSM drive has two control loops, (i) inner current control loop, and (ii) outer speed control loop. Only two phase motor currents are sensed and the third phase current is obtained by summing and inverting the two-phase currents because there is no neutral connection. The three phase reference currents are generated by outer speed control loop. The outer speed control loop requires actual rotor position and its speed.

The speed error is processed by the speed controller to obtain torque command.

The d-axis current of the motor is kept zero. The stator reference three phase current commands are obtained using Park's transformation of the rotor reference current commands I_q , I_d and rotor electrical angle θ .

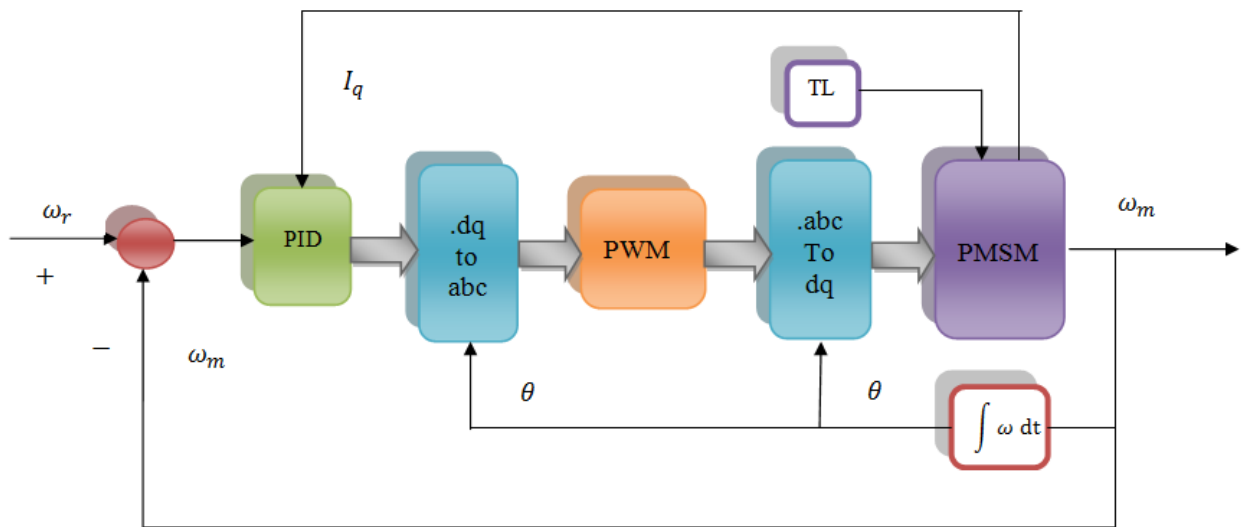


Fig.4.8. Permanent magnet synchronous motor drive system

4.3.1 .PMSM d-q modeling and simulation

Detailed modeling of PM motor drive system is required for proper simulation of the system. The **d-q** model has been developed on rotor reference frame as shown in Figure 4.9. At any time t , the rotating rotor d-axis makes an angle θ_r with the fixed stator phase axis and rotating stator mmf makes an angle α with the rotor d-axis. Stator mmf rotates at the same speed as that of the rotor.

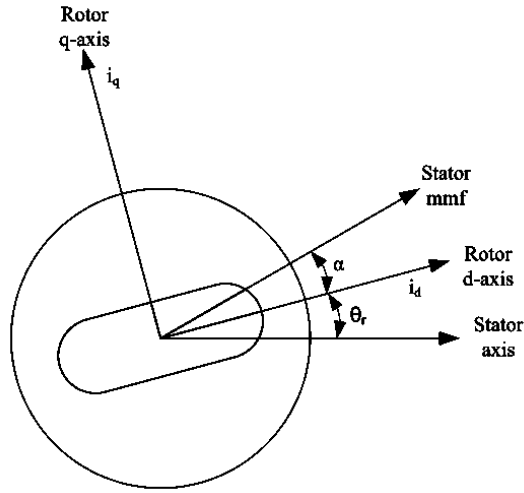


Fig.4.9. Motor axis.

The model of PMSM without damper winding has been developed on rotor reference frame using the following assumptions:

- 1) Saturation is neglected.
- 2) The induced EMF is sinusoidal.
- 3) Eddy currents and hysteresis losses are negligible.
- 4) There are no field current dynamics.

As mentioned earlier, the synchronous motor performance can be analyzed in two ways. Section 4.3.1 focuses on the second system known as dq , which normally stands for direct and quadrature axes as opposed to the conventional x , y and z axes. This means the system will now be 2-phase instead of 3-phase. Figure.4.10. shows how the two systems relate to each other.

The equations that relate the stator currents of the two systems are as follows [33].

$$\begin{bmatrix} i_{sd}(t) \\ i_{sq}(t) \end{bmatrix} = \underbrace{\sqrt{\frac{2}{3}} \begin{bmatrix} \cos(\theta_{da}) \cos\left(\theta_{da} - \frac{2\pi}{3}\right) \cos\left(\theta_{da} - \frac{4\pi}{3}\right) \\ -\sin(\theta_{da}) - \sin\left(\theta_{da} - \frac{2\pi}{3}\right) - \sin\left(\theta_{da} - \frac{4\pi}{3}\right) \end{bmatrix}}_{[T_s]_{abc \rightarrow dq}} \begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} \quad (4.30)$$

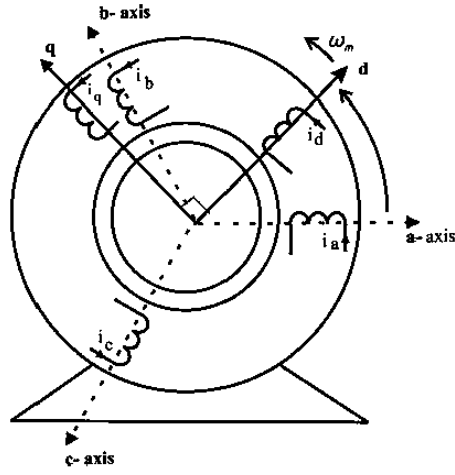


Fig.4.10. Scheme of ABC and dq system.

$$\begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} = \underbrace{\sqrt{\frac{2}{3}} \begin{bmatrix} \cos(\theta_{da}) & -\sin(\theta_{da}) \\ \cos\left(\theta_{da} + \frac{4\pi}{3}\right) & -\sin\left(\theta_{da} + \frac{4\pi}{3}\right) \\ \cos\left(\theta_{da} + \frac{2\pi}{3}\right) & -\sin\left(\theta_{da} + \frac{2\pi}{3}\right) \end{bmatrix}}_{[T_s]_{dq \rightarrow abc}} \begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} \quad (4.31)$$

where $[T_s]_{abc \rightarrow dq}$ is the transformation matrix to transform stator a-b-c phase winding currents to the corresponding dq winding currents and $[T_s]_{dq \rightarrow abc}$ is the transformation matrix in the reverse direction. The same transformation matrix relates the flux linkages λ and voltages.

Voltage equations are given by:

$$V_q = R_s i_q + \omega_r \lambda_d + \frac{d\lambda_q}{dt} \quad (4.32)$$

$$V_d = R_s i_d - \omega_r \lambda_q + \frac{d\lambda_d}{dt} \quad (4.33)$$

Let us assume $\rho = \frac{d}{dt}$, then

$$V_q = R_s i_q + \omega_r \lambda_d + \rho \lambda_q \quad (4.34)$$

$$V_d = R_s i_d - \omega_r \lambda_q + \rho \lambda_d \quad (4.35)$$

where the speed of the equivalent dq windings is $\omega_d = \omega_r$ (in electrical rad/s) in order to keep the d-axis always associated with the rotor magnetic axis.

The stator d and q - winding flux linkages can be written as follows:

$$\lambda_q = L_q i_q \quad (4.36)$$

$$\lambda_d = L_d i_d + \lambda_m \quad (4.37)$$

Substituting equations 4.36 and 4.37 into 4.34 and 4.35

$$V_q = R_s i_q + \omega_r (L_d i_d + \lambda_m) + \rho \lambda_q \quad (4.38)$$

$$V_d = R_s i_d - \omega_r \lambda_q + \rho (L_d i_d + \lambda_m) \quad (4.39)$$

Where:

V_q – q-axis voltage

V_d – d-axis voltage

R_s – stator resistance

i_q – q-axis current

i_d – d-axis current

ω_r – electrical speed

L_d – Direct-axis inductance

L_q – Quadrature-axis inductance

λ_q – flux linkage due q axis

λ_d – flux linkage due d axis

λ_m – Permanent magnet flux

λ – flux linkage

Arranging equations 4.38 and 4.39 in matrix form

$$\begin{bmatrix} V_q \\ V_d \end{bmatrix} = \begin{bmatrix} R_s + \rho L_q & \omega_r L_d \\ -\omega_r L_q & R_s + \rho L_d \end{bmatrix} \begin{bmatrix} i_q \\ i_d \end{bmatrix} + \begin{bmatrix} \omega_r \lambda_m \\ \rho \lambda_m \end{bmatrix} \quad (4.40)$$

Figure.4.11, shows a dynamic equivalent circuit of an PM synchronous machine based on Eqs. Note that in practice, magnetic circuits are subject to saturation as current increases. Especially, when I_q is increased, the value of L_q is decreased and λ_m and L_d is subject to armature reaction. Since I_d is maintained to zero or negative value (demagnetizing) in most operating conditions, saturation of L_d rarely occurs.

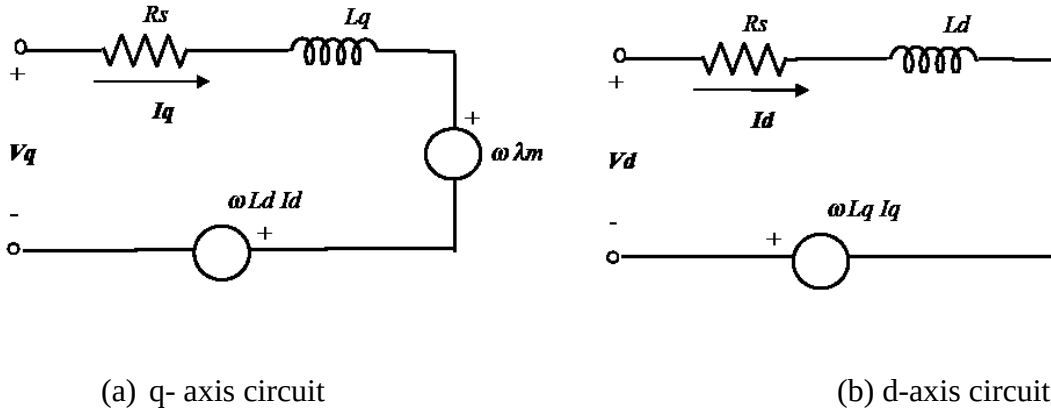


Fig.4.11.Equivalent circuit of a PM synchronous motor

The developed torque motor is being given by :

$$T_e = \frac{3}{2} \left(\frac{P}{2} \right) (\lambda_d i_q - \lambda_q i_d) \quad (4.41)$$

Substituting for flux linkages in the above equation for the non-salient 6-pole motor,

$$T_e = \frac{3}{2} \left(\frac{P}{2} \right) [(L_d i_d + \lambda_m) i_q - \lambda_q i_d] = \frac{3}{2} \left(\frac{P}{2} \right) \lambda_m i_q \quad (4.42)$$

The mechanical Torque equation is

$$T_e = T_L + B\omega_r + J \frac{d\omega_r}{dt} \quad (4.43)$$

$$T_e = K_t i_q + \frac{3}{2} \frac{P}{2} (L_d - L_q) i_d i_q \quad (4.44)$$

$$K_t = \frac{3}{2} \frac{P}{2} \lambda_m \quad (4.45)$$

The acceleration is determined by the difference of the electromagnetic torque and the load torque (including friction torque $T_{fr} = \omega_{mech} \cdot B$) acting on J_{eq} , the combined inertia of the load and the motor

$$\omega_r = \int \left(\frac{T_e - T_L - B\omega_r}{J} \right) dt \quad (4.46)$$

And

$$\omega_m = \omega_r \left(\frac{2}{P} \right) \quad (4.47)$$

In the above equations ω_r is the rotor electrical speed where as ω_m is the rotor mechanical speed.

Using the above equations the motor can be simulated in the simulink software as shown in the Figure.4.12.

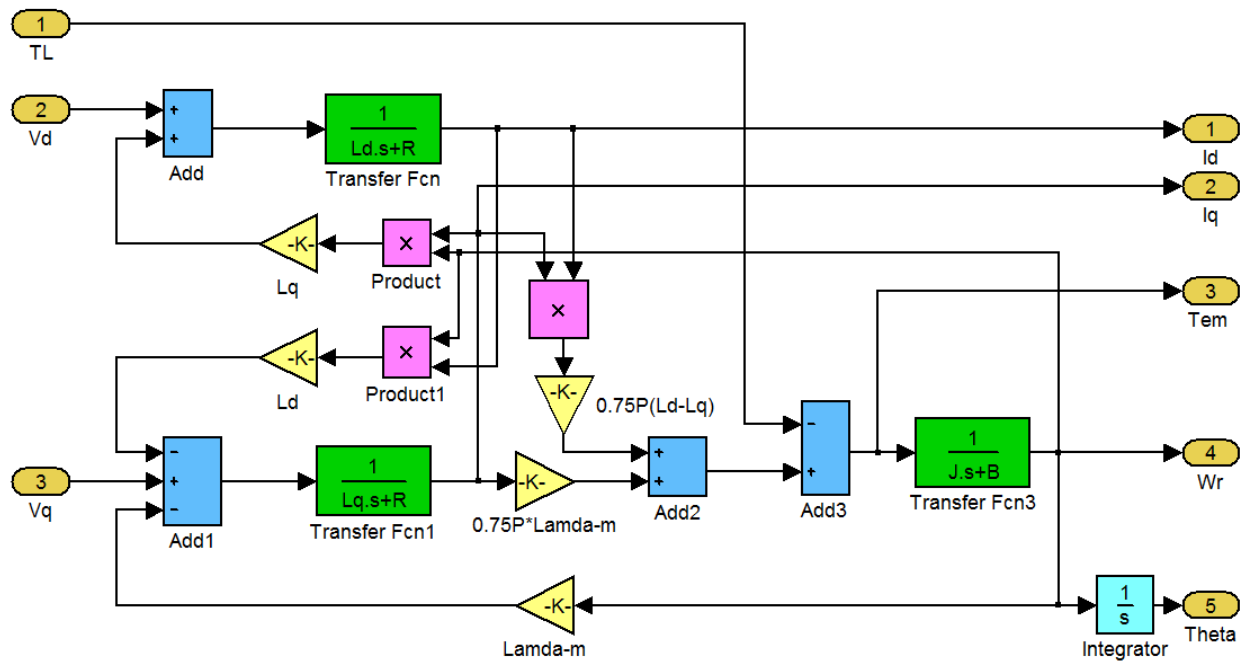


Fig.4.12. MATLAB Simulation of d-q model of PMSM

4.3.2. Inverter and modulator model

The inverter bridge is shown in Figure.4.13 , and its consist of three switches which connect each motor terminal to either the positive or the negative rail of a constant dc voltage source.

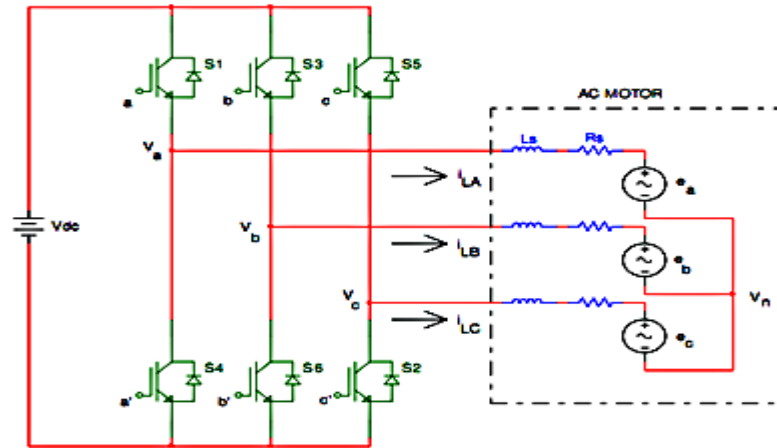


Fig.4.13. Three phase voltage source PWM inverter circuit.

These switches (power device) can be either BJT or IGBT and their maximum frequency of operation is an important factor in determining the modulation strategy to be use. In fact, in the variable voltage , variable frequency inverter drives, the dc link can be kept constant by a front end diode bridge and the fundamental output frequency of the output voltage can be controlled within the inverter by using Pulse Width Modulated (PWM) technique.

In SIMULINK, PWM waveform can be easily generated, using SOURCE, REPEATING SEQUENCE block, SUM and SWITCH blocks. The three phase SIN signals directly come from the source, and the triangle waveform must be changed from the REPEATING SEQUENCE block. The implementation of the three phase sinusoidal PWM inverter under SIMULINK environment is as shown in Figure.4.14.a. Input signals to the inverter are fundamental frequency, modulation index, and carrier frequency. Output signals from the inverter are three phase voltage.

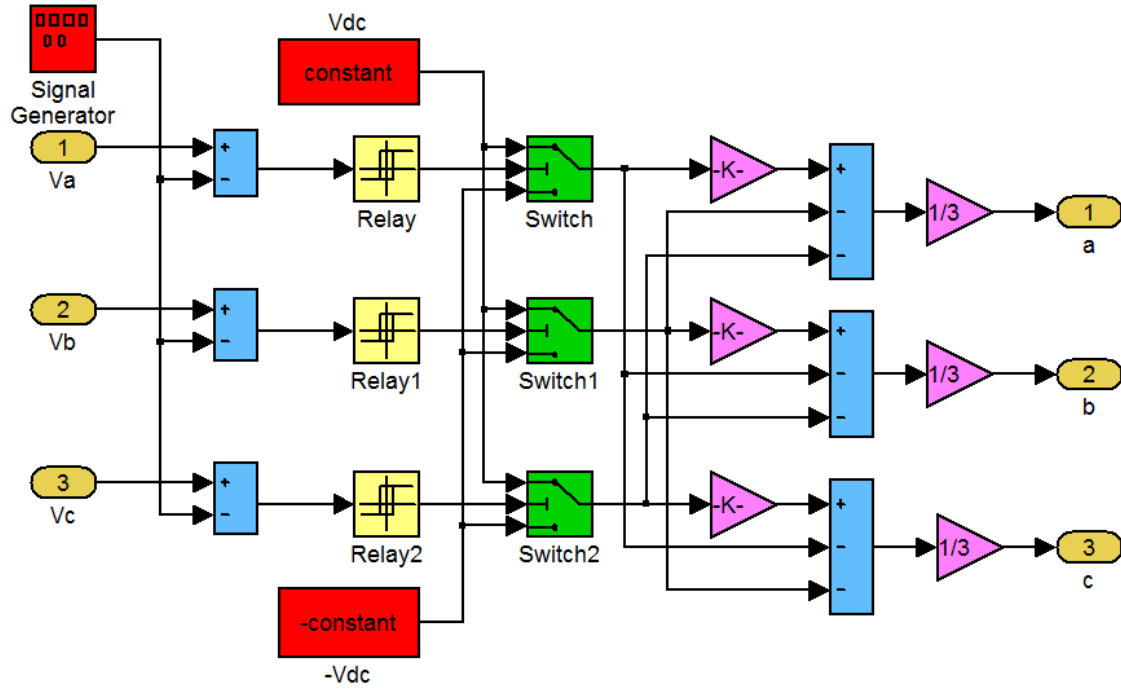


Fig.4.14.a. SIMULINK model of voltage source PWM inverter circuit.

For the purpose of the simplicity, only the PWM inverter part of the power circuit has been modeled in this investigation. Therefore, the dc link voltage V_{dc} is assumed constant.

$$V_{an} = \frac{(2V_{ao} - V_{bo} - V_{co})}{3} \quad (4.48)$$

$$V_{bn} = \frac{(2V_{bo} - V_{co} - V_{ao})}{3} \quad (4.49)$$

$$V_{cn} = \frac{(2V_{co} - V_{bo} - V_{ao})}{3} \quad (4.50)$$

The peak value and frequency of the sine wave represent the magnitude and frequency of the desired output. The ratio between the reference signal peak value V_r and the carrier signal peak value V_r is the modulation index M for the PWM inverter and is normally less than one when the fundamental frequency is less than base frequency.

The modulation index used in this simulation study was $M = 0.9$, and the switching frequency of the inverter was 10khz. After masking the PWM inverter, the block can be directly used with the adjusting of only two parameters that are, the modulation index M and rate of frequency.

Figure.4.14.b. Phase A voltage input and output of PWM inverter.

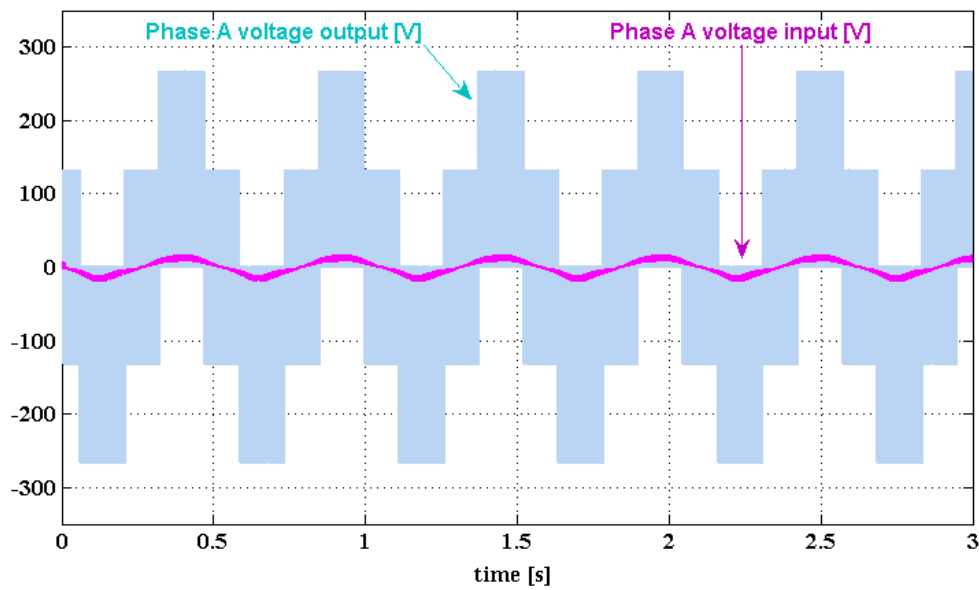


Fig.4.14.b. Phase A voltage input and output of PWM inverter.

4.3.3. PI controller simulation

For this controller the gain k_i is about an order of magnitude of 10 less than the gain k_p . The implementation of PI controller in MATLAB /SIMULINK environment is as shown in Figure.4.15.

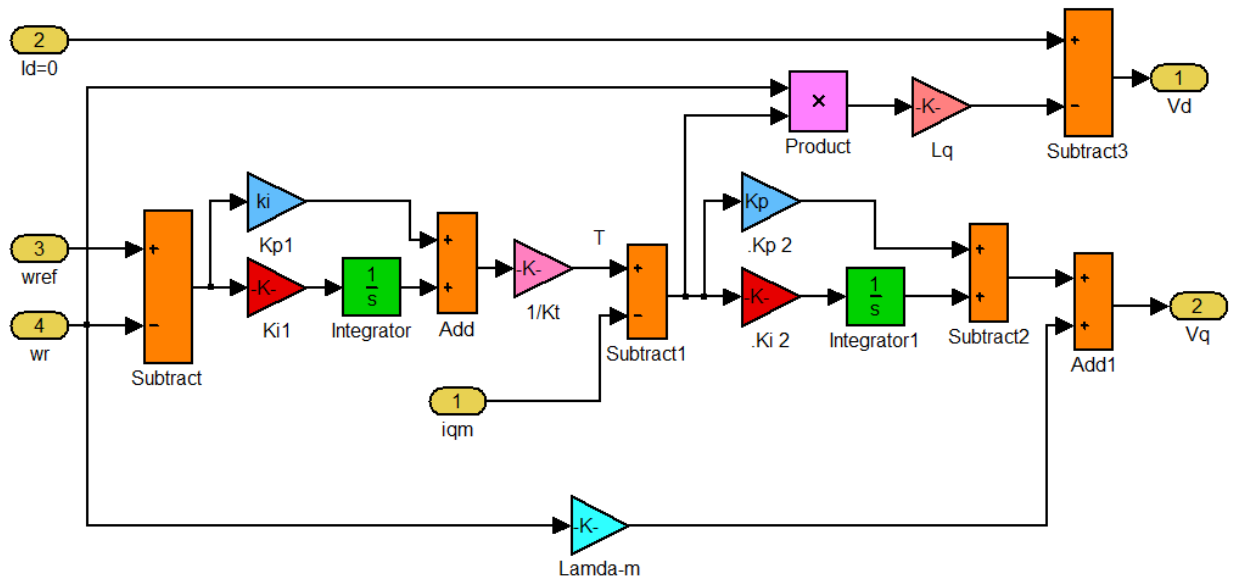


Fig.4.15. MATLAB /SIMULINK of PI controller

4.3.4. Electrical machine model transformation

The important transformations were used in the control system as shown bellow in the Figure.4.16 .

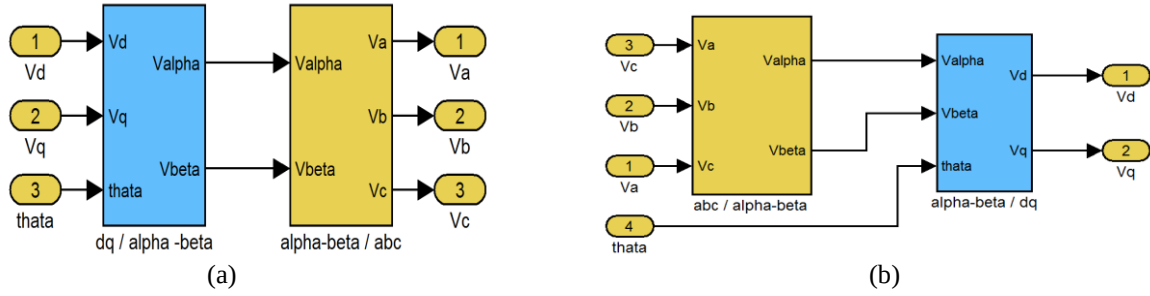


Fig.4.16. subsystem of (a) Inverse (Park & Clarke) transform (b) (Park & Clarke) transform

4.3.4 .1. Inverse Park transform

After the PI iteration you have two voltage component vectors in the rotating d-q axis. You will need to go through complementary inverse transforms to get back to the 3-phase motor voltage. First you transform from the two axis rotating d-q frame to the two axis stationary frame α - β . This transformation uses the Inverse Park Transform, as illustrated in Figure 4.17. and Figure.4.18. illustrate MATLAB /SIMULINK of Inverse Park Transform.

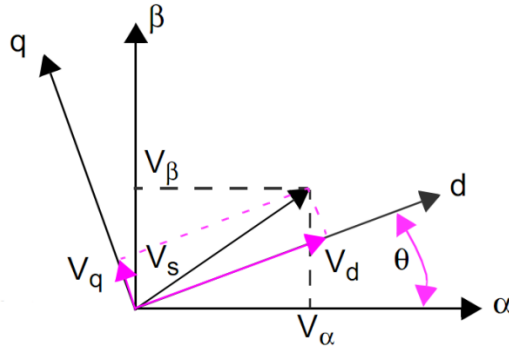


Fig 4.17. Inverse Park Transform

$$V_\alpha = V_d * \cos \theta - V_q * \sin \theta \quad (4.59)$$

$$V_\beta = V_d \sin \theta + V_q * \cos \theta \quad (4.60)$$

Where:

V_α – α -axis voltage V_β – β -axis voltage

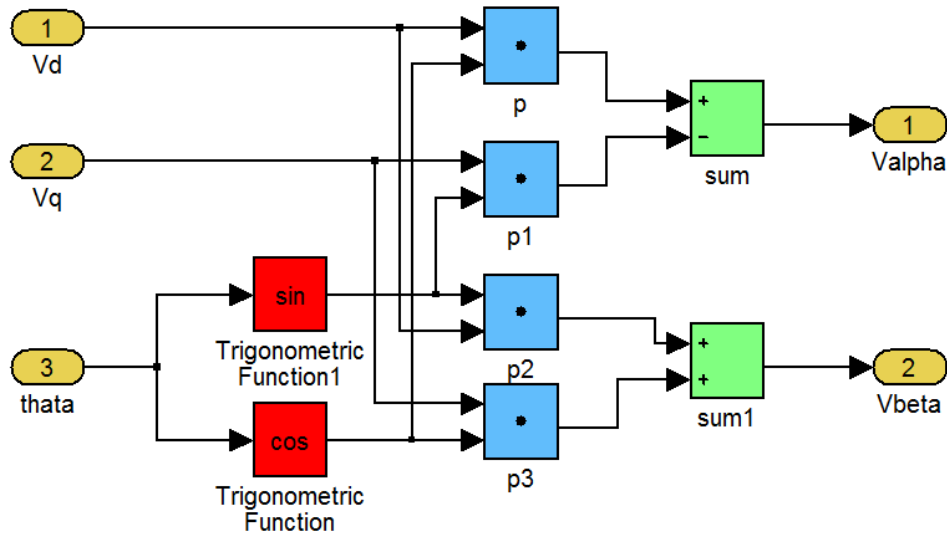


Fig 4.18. MATLAB /SIMULINK of Inverse Park Transform.

4.3.4 .2. Inverse Clarke transform

The next step is to transform from the stationary two axis α - β frame to the stationary three axis, 3-phase reference frame of the stator. Mathematically, this transformation is accomplished with the Inverse Clark Transform, as illustrated in Fig.4.19. and Fig 4.20. illustrate MATLAB /SIMULINK of Inverse Clark Transform.

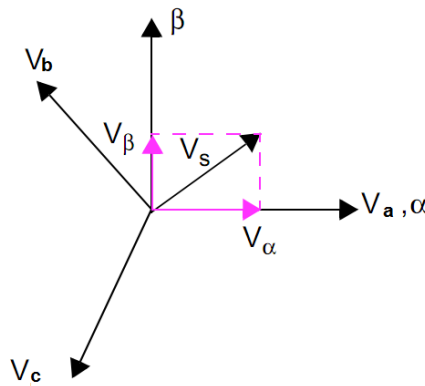


Fig.4.19. Inverse Clark Transform

$$V_a = V_\alpha \quad (4.61)$$

$$V_b = \frac{1}{2}(-\sqrt{3} * V_\beta - V_\alpha) \quad (4.62)$$

$$V_c = \frac{1}{2}(\sqrt{3} * V_\beta - V_\alpha) \quad (4.63)$$

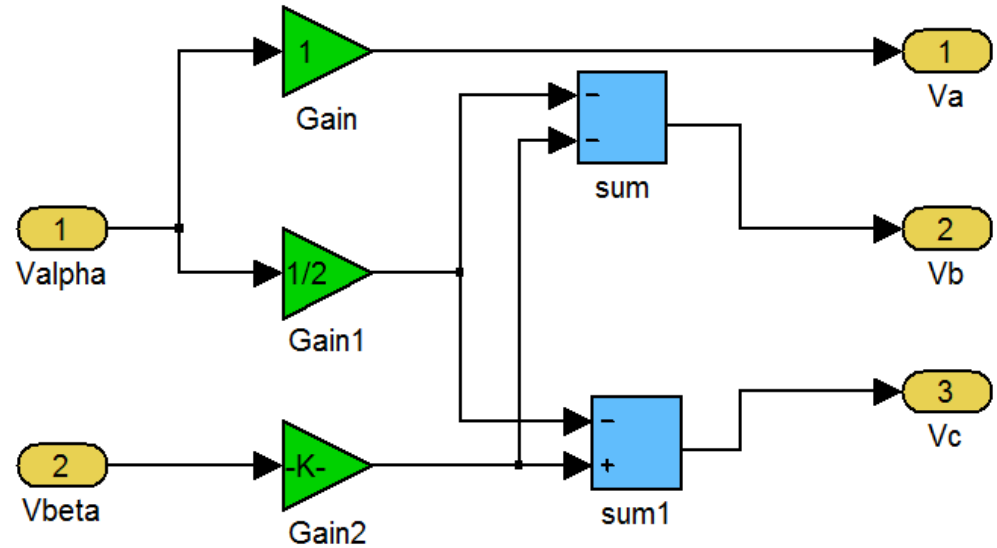


Fig 4.20. MATLAB /SIMULINK of Inverse Clark Transform.

4.3.4 .3.Clarke transform

The first coordinate transform, called the Clarke Transform, moves a three axis, two dimensional coordinate system, referenced to the stator, onto a two axis system, keeping the same reference (see Figure 4.21, where V_a , V_b and V_c are the individual phase currents).

Figure.4.22. illustrate MATLAB /SIMULINK of Clark Transform.

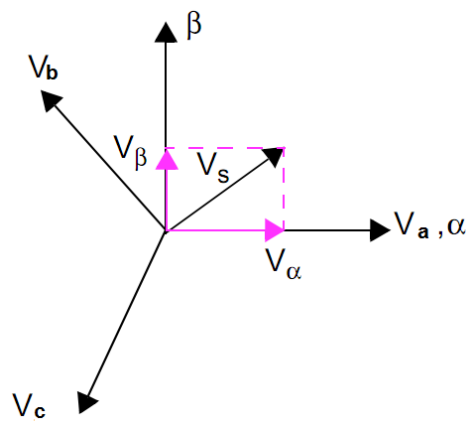


Fig.4.21.Clark Transform.

$$V_{\alpha} = V_a \quad (4.64)$$

$$V_{\beta} = \sqrt{3}(V_b - V_c) \quad (4.65)$$

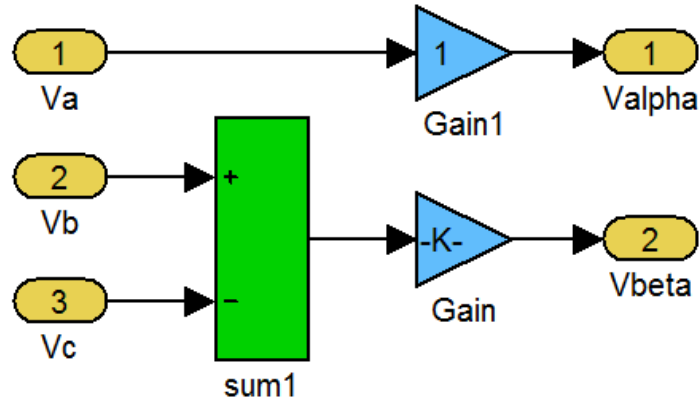


Fig 4.22. Illustrate MATLAB /SIMULINK of Clark Transform.

4.3.4 .4. Park transforms

At this point, you have the stator current represented on a two axis orthogonal system with the axis called α - β . The next step is to transform into another two axis system that is rotating with the rotor flux. This transformation uses the Park Transform, as illustrated in Figure.4.23. This two axis rotating coordinate system is called the ***d-q*** axis. θ represents the rotor angle. Figure.4.24. illustrate MATLAB /SIMULINK of Inverse Park Transform.

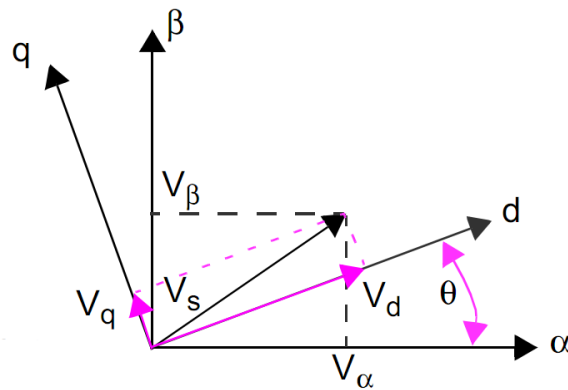


Fig 4.23. Park Transform

$$V_d = V_\beta * \cos \theta - V_\alpha * \sin \theta \quad (4.66)$$

$$V_q = V_\beta * \sin \theta + V_\alpha * \cos \theta \quad (4.67)$$

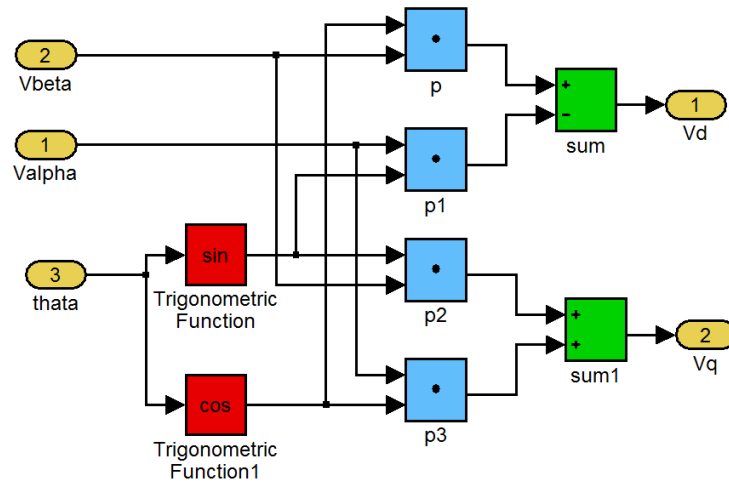


Fig 4.24. MATLAB /SIMULINK of Park Transform.

4.3.5 .Tuning of PID controller and the results obtained.

The PMSM drive simulation was built in several steps like ***dq*** variables transformation to ***abc*** phase, calculation torque and speed, control circuit, inverter and PMSM. The ***dq*** variables transformation to ***abc*** phase is built using the reverse Parks transformation. For simulation purpose the voltages are the inputs and the current are output. Using all the drive system blocks, the complete system block has been developed as shown in Figure.4.25.

The system built in Simulink for a PMSM drive system has been tested with the space vector modulation method at the constant torque region of operation ($TL = 0$).

The motor parameters used for simulation of ***dq*** model of PMSM are given in Table 2.5.

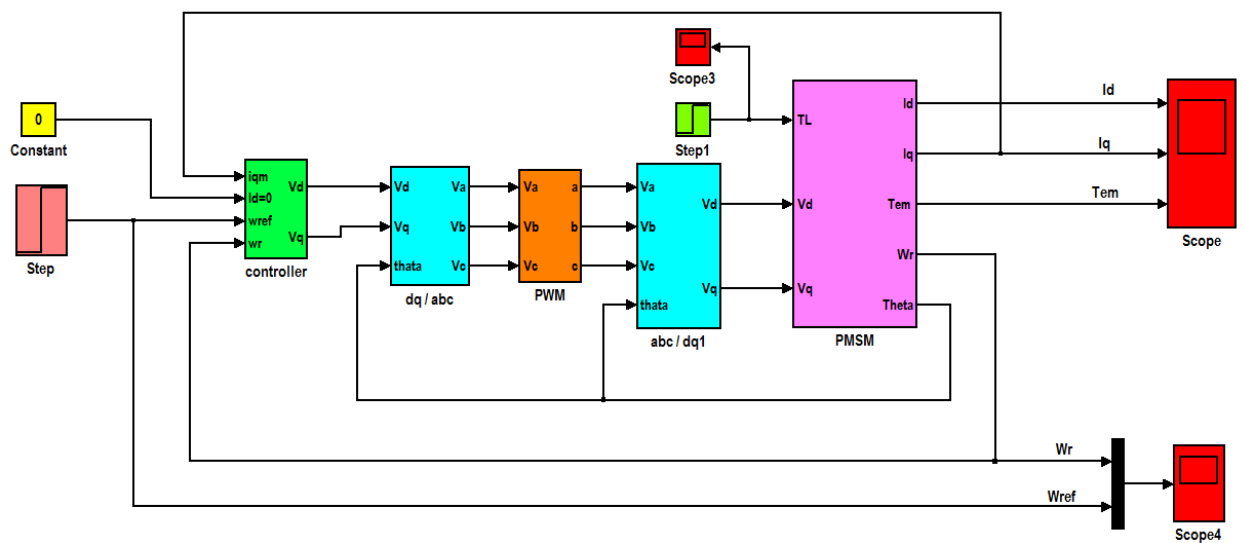


Fig.4.25. MATLAB /SIMULINK of PMSM drive system.

Speed control can be conveniently achieved in the PMSM drive by closing a speed feedback loop around the inner torque/current loop as illustrated in Figure.4.25. The torque request is generated by the speed controller in dependence on the speed error as in Figure.4.15. By keeping I_d current frozen at zero, the phasor of stator current will be placed just in the quadrature axis and the maximal driving torque will be achieved. This can be obtained by setting the demanded current I_d^* to zero. Then by tuning the parameters of (PI controller 1) and (PI controller 2) to obtain the request torque, and motor speed with minimum raise time, minimum steady state error, and no overshoot accord. The parameter was $K_p1 = 5$, $K_i1 = 0.0006$, $K_p2 = 75$, and $K_i2 = 0.005$. Subsequently graphical rotor speed with time before and after zoom in, is shown in Figure.5.18.(a,b).

5. Simulation results of PM motor.

In this chapter we illustrate the results obtained from the simulation of permanent magnet motor models (BLDC motor and PMSM), and offers a brief description for all figures separately.

5.1. Simulation results of permanent magnet BLDC motor.

After running the simulation, the Phase A Voltage and Current waveforms were recorded as shown in Figure.5.1. From the figure we show that the phase A voltage is rectangular and the current was distorted from the square wave shape to something between a sine-wave and a square wave, and Figure.5.4.show the current wave form with a big scale . This is a result of the inductance effect

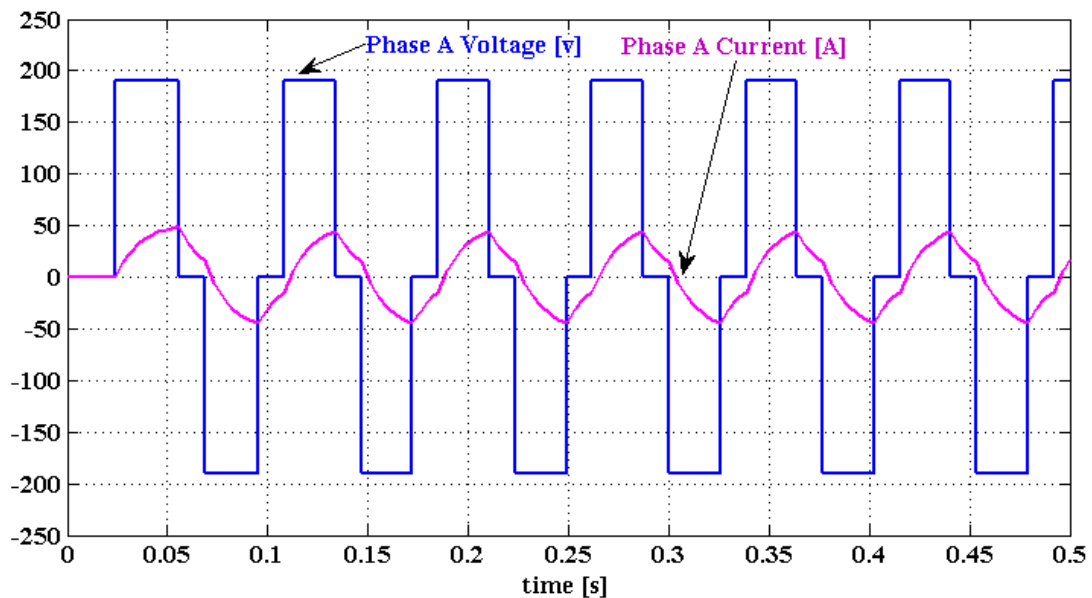


Fig. 5.1. Phase A voltage [v], and current [A]

Figure.5.2.show the speed response of PM BLDC motor, the speed of motor oscillates around 16.4 rad /s, and reach steady state at around 0.25 seconds.

Due to change of position the resultant relation of stationary flux ϕ_f and flux ϕ_a , change with time, producing some ripples in speed of motor.

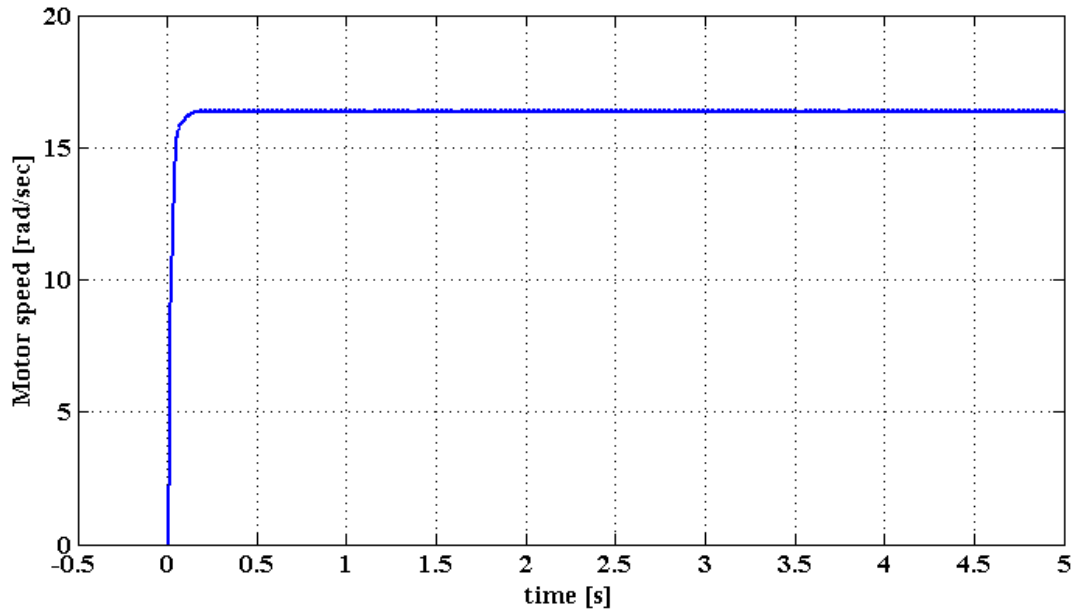


Fig. 5.2. BLDC motor, speed response.

Electromagnetic torque T_{em} as shown in Figure.5.3. responds with high value cumbered with Load Torque, then come to oscillates around 6.45 [N.m].

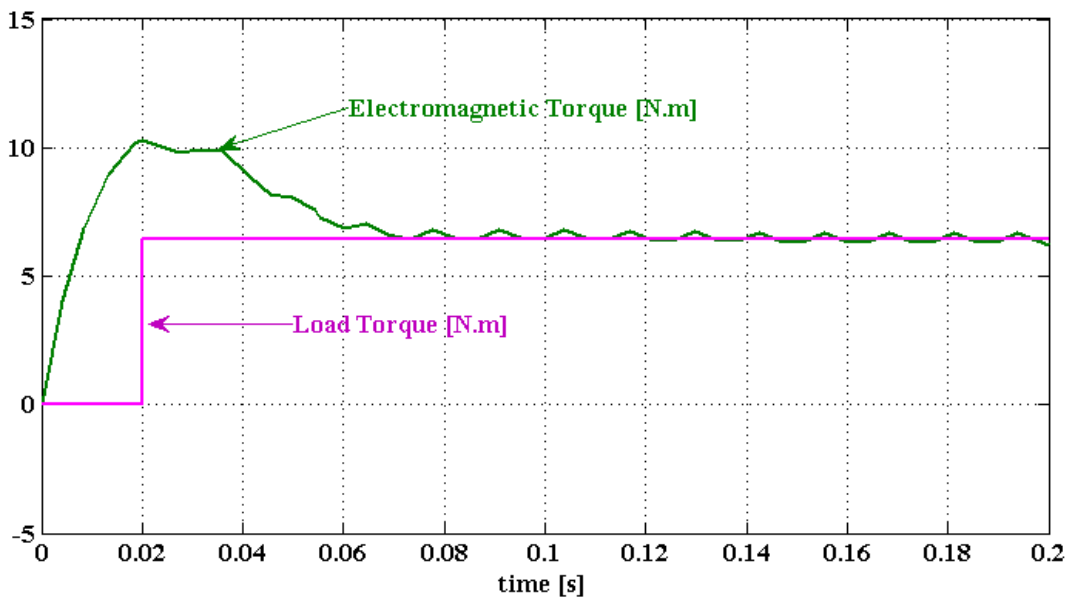


Fig. 5.3. BLDC motor Electromagnetic Torque with Load Torque.

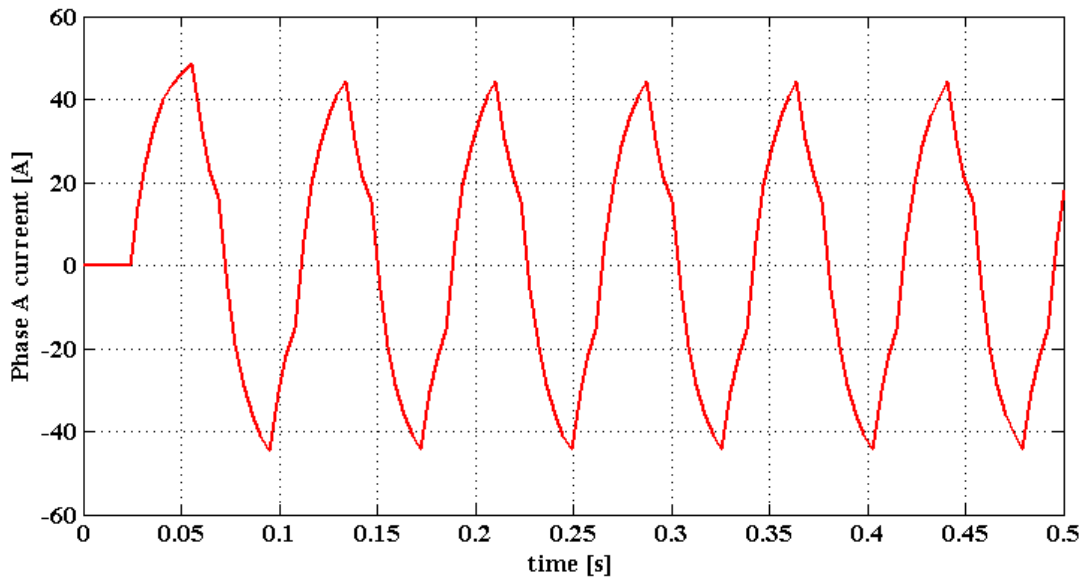


Fig. 5.4. BLDC motor Phase A current response.

Source current of BLDC motor was distorted as phase current by effect of inductance, and the increasing in amplitude of source current at the initial of time because the absence of load torque effect.

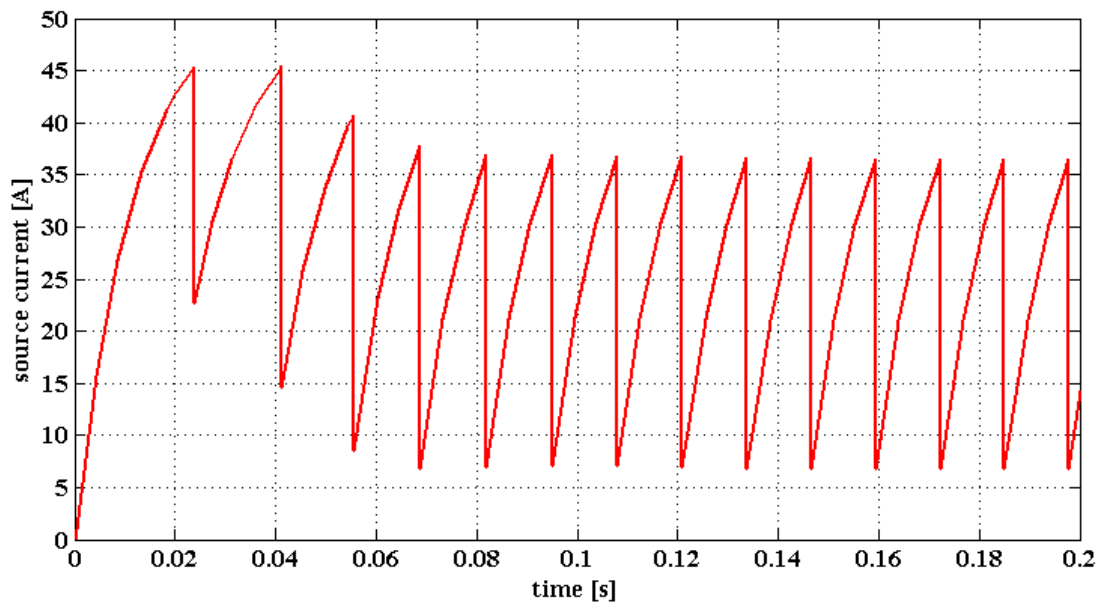


Fig. 5.5.BLDC motor source current [A].

Speed-torque curve, with the current-torque curve was drawn in Figure.5.6, and the figure show that the current was increase linearly with torque, and the speed decreased fast. When motor reaches study state the speed torque curve was straight line because inductance played no part in affecting. So Figure.5.6, show the efficiency curve. the BLDC motor has increasing efficiency as the simulation begins, it slowly decreases with increasing the load torque.

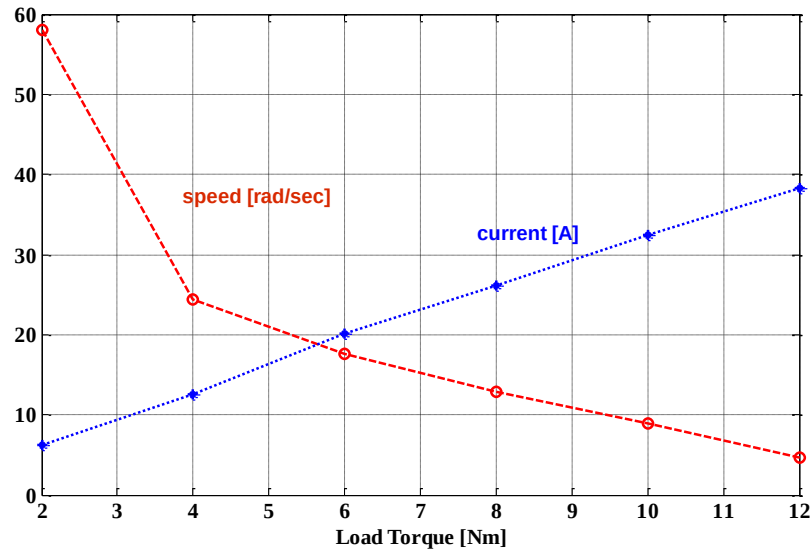


Fig.5.6. Speed and current-torque characteristics determined in steady-state, from the dynamic simulation.

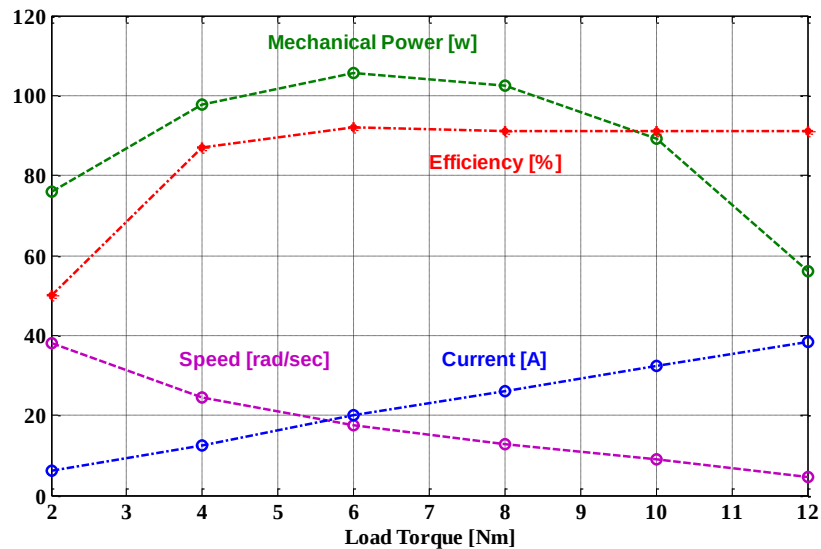


Fig. 5.7. Characteristics obtained from both the BLDC motor model after it reaches steady-state and the brush DC motor model.

5.2. Simulation Results of PMBLDC motor with PI Controller.

This section shows MATLAB simulink results of Permanent Magnet BLDC Motor and its speed response after controlling using PI controller. Figure.5.8, show the speed response with elimination of oscillation. So Figure.5.9, show the improvement in phase current.

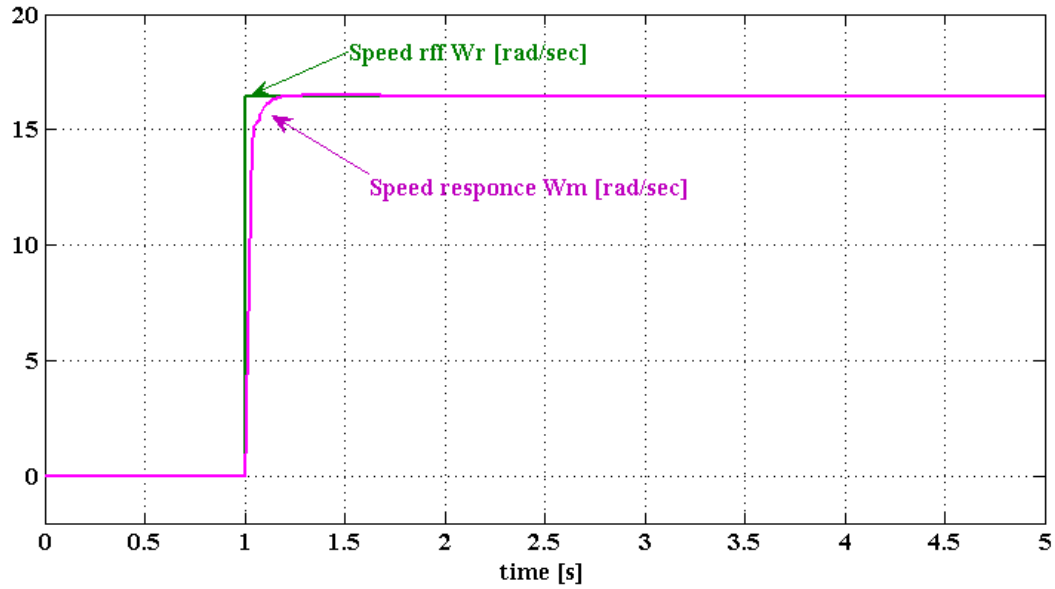


Fig 5.8. BLDC Motor speed with PID controller

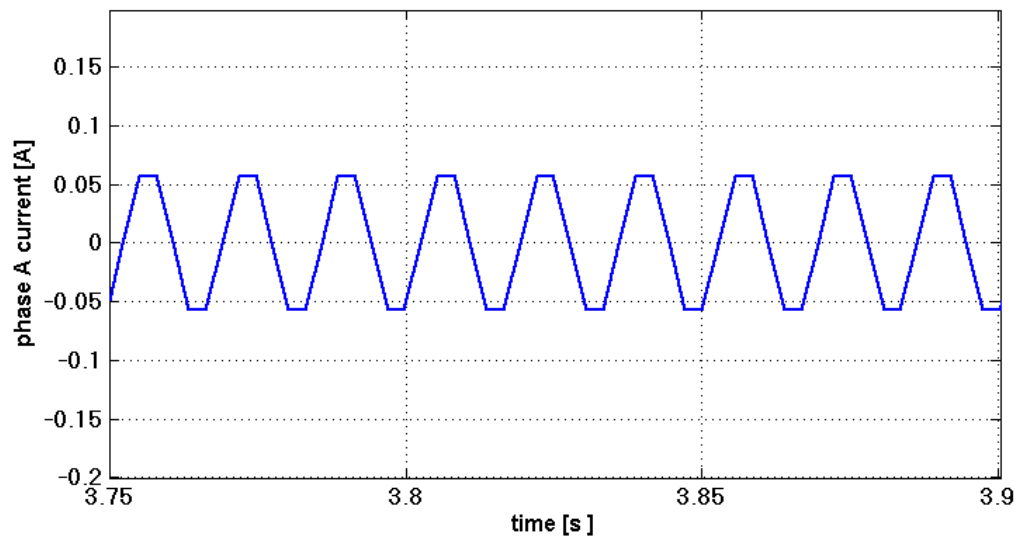


Fig 5.9. phase A current of motor with PID controller

5.3. Simulation Results of PMSM.

The PMSM was supplied with a sinusoidal voltage of 220V. Figure.5.10, show a three phase voltage source, and Figure5.11, show a phase A current and voltage of PMSM (single phase).

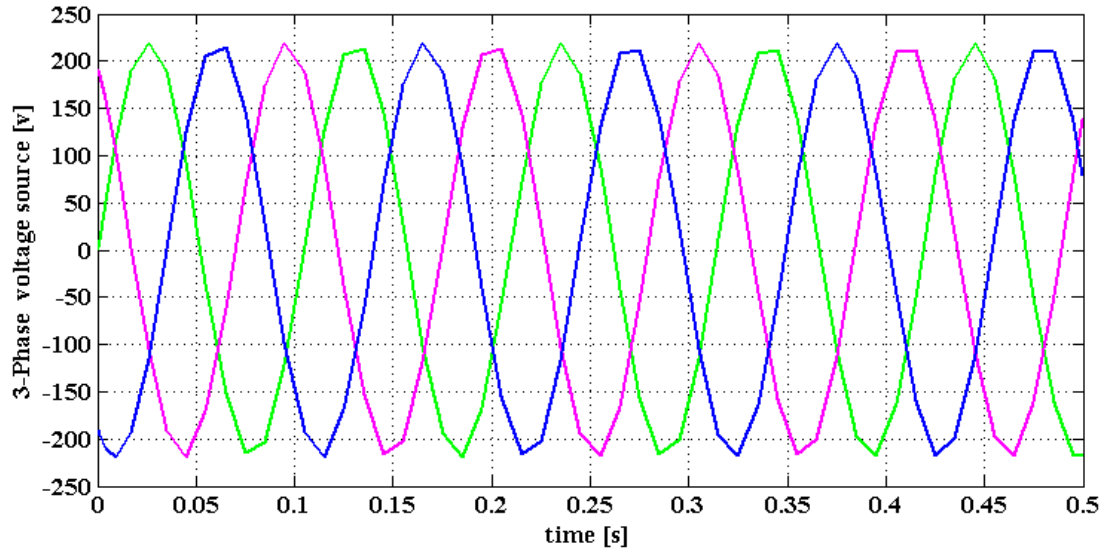


Fig 5.10. 3-Phase voltage source of PMSM.

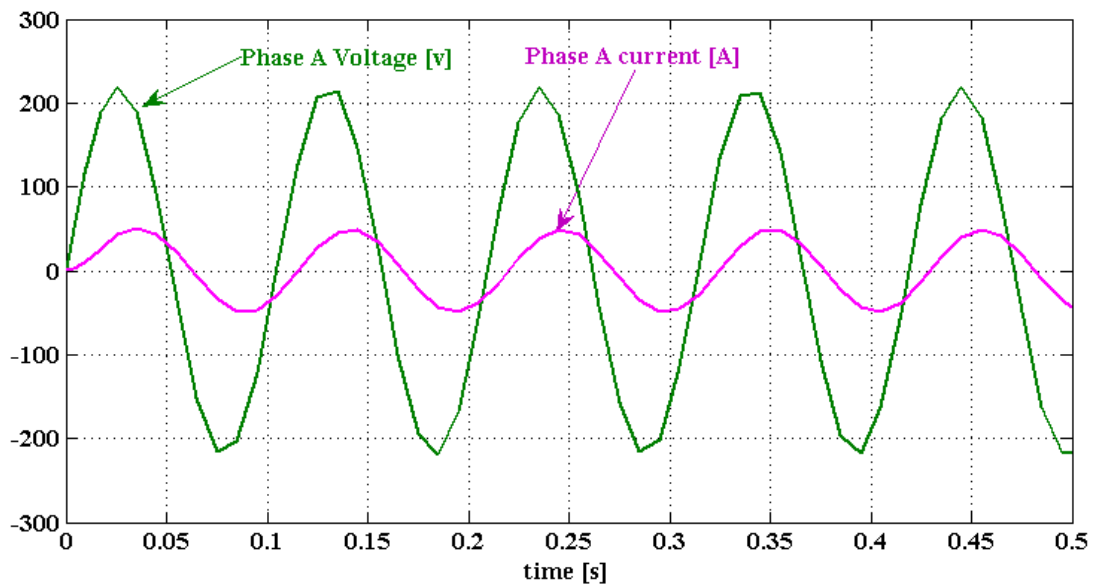


Fig 5.11. phase A current and voltage of PMSM

After running the simulation, the speed response of motor oscillates around of 12.04 rad/s with overshoot equal 27.48 ra/sec as shown in Figure.5.12, and the motor reaches steady-state at around 0.75 seconds. The motor was loaded with a rated torque of 6.45 N.m. So

the electromagnetic torque oscillates around 6.45 N.m, and need 0.75 seconds to reach steady state as shown in Figure.5.13.

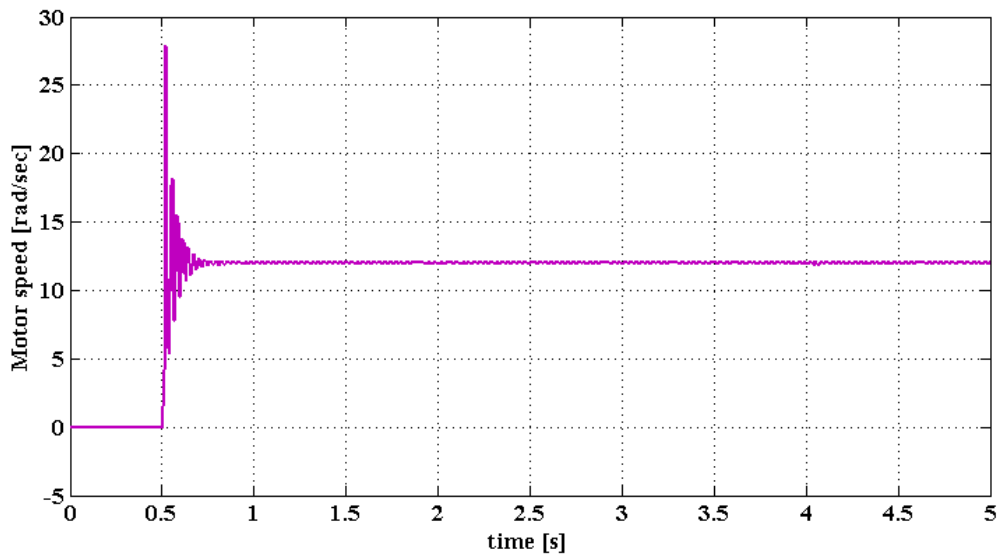


Fig 5.12. Motor speed response of PMSM.

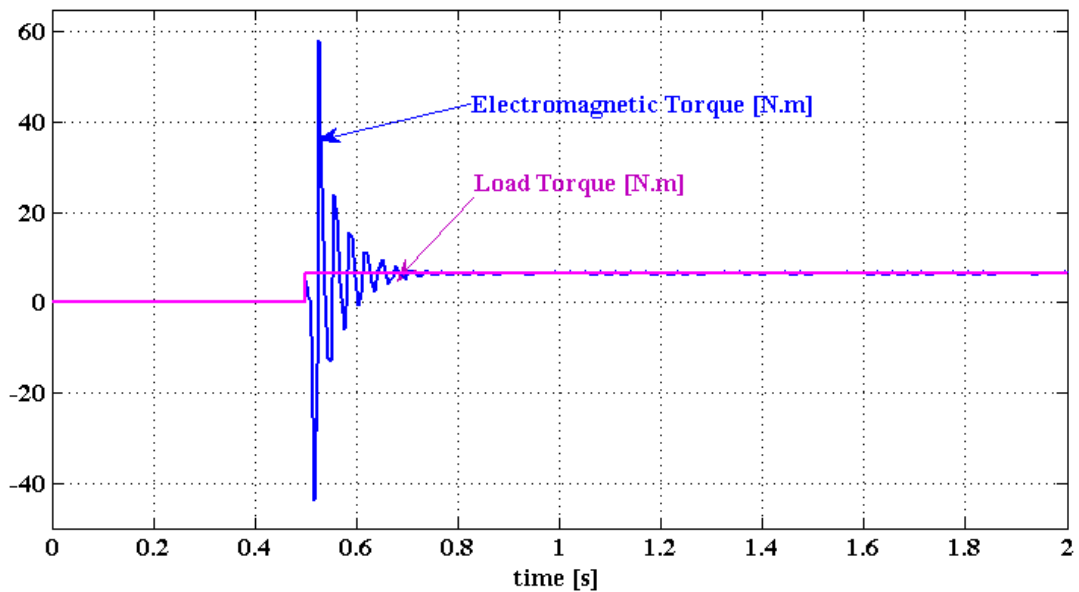


Fig 5.13. Electromagnetic Torque response with Load Torque of PMSM.

Figure.5.14, show the input power equal 85.19, and the Figure.5.15, show the output power of motor, because the output power depends on load torque T_L and motor speed ω_m (recall equation 3.20) we show the oscillation with overshoot before reach steady state.

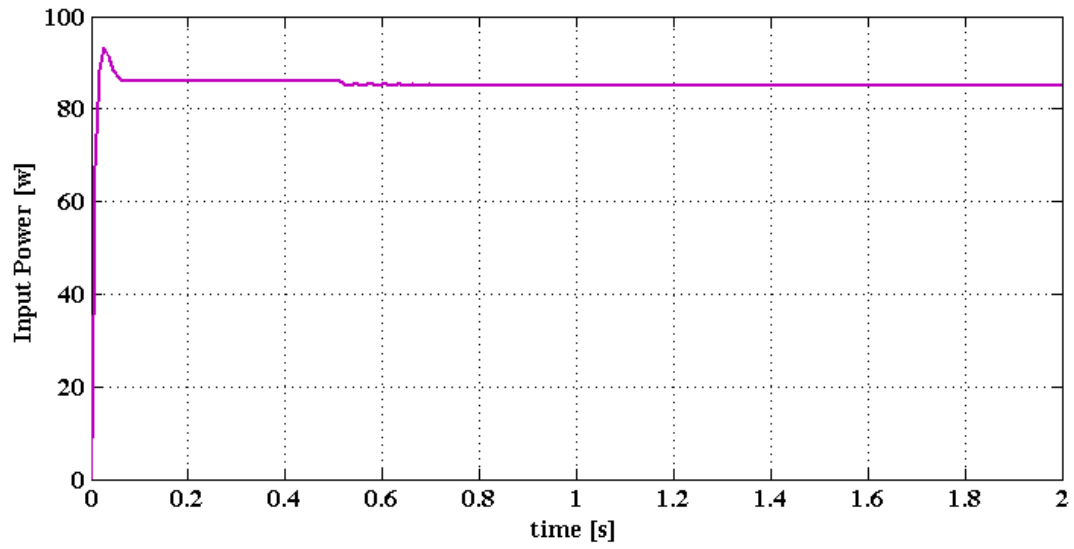


Fig 5.14. Input Power of PMSM.

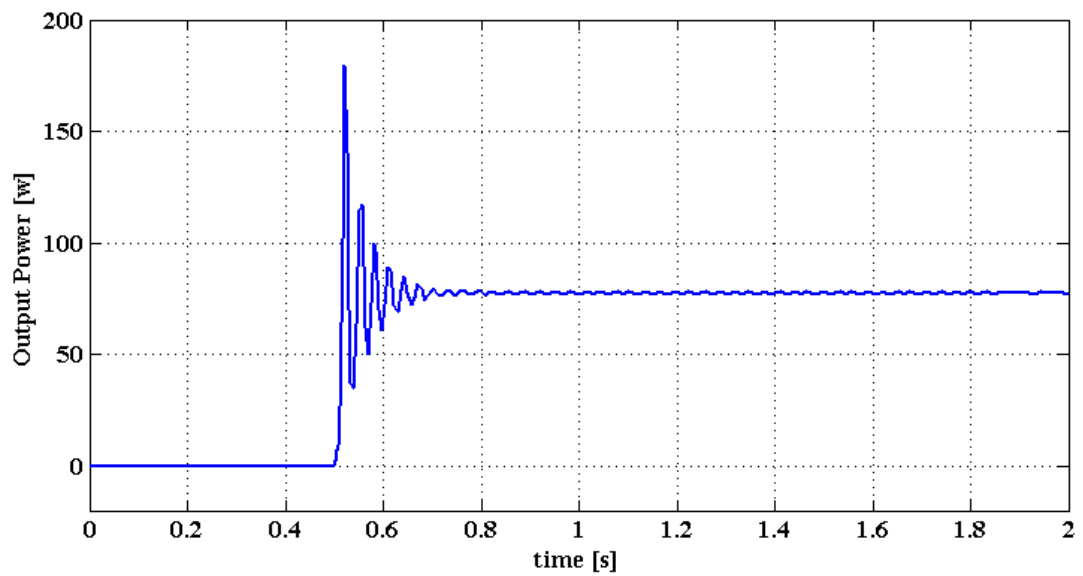


Fig 5.15. Output Power response of PMSM.

Duo to equation (3.19) the efficiency wave form was as shown in Figure 5.16, and at steady state it equal 91.18 %.

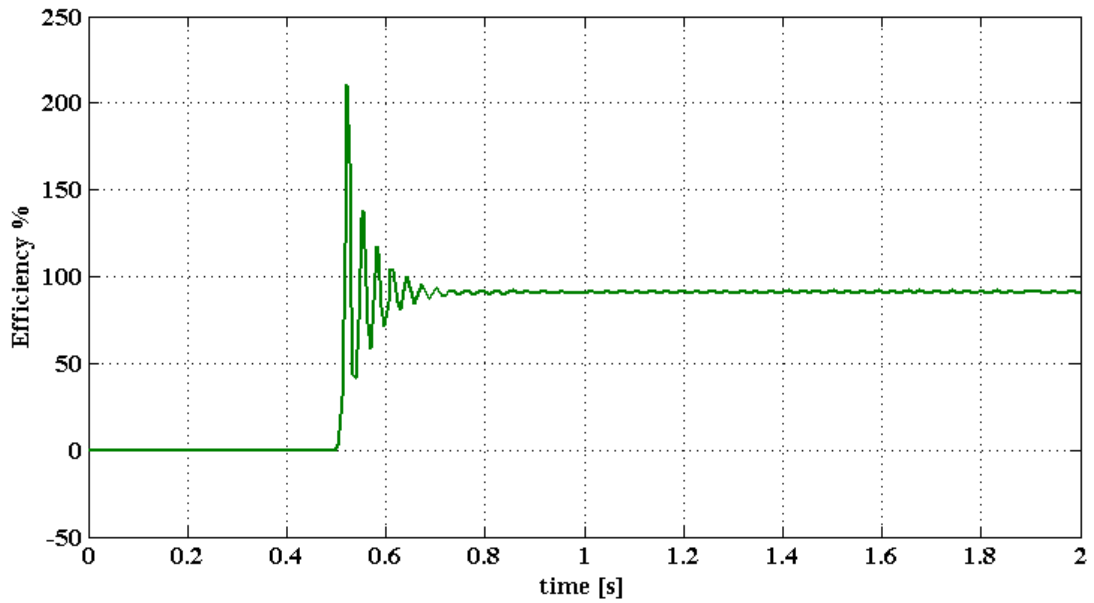


Fig 5.16.Efficiency waveform of PMSM.

The performance characteristics are presented in Figure.5.17. The efficiency of the motor increases step by step, and the speed stays constant in any case of load torque, but the current increasing as the torque increases.

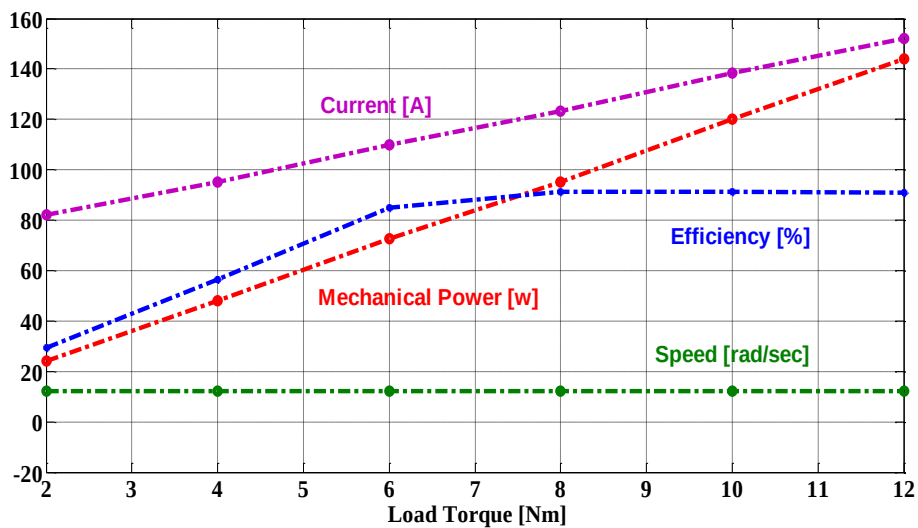


Fig. 5.17. Characteristics of steady-state and dynamic models for the synchronous motor.

5.4. Simulation Results of PMSM with PI Controller.

After running a simulation the speed response of motor was as shown in Figure.5.18a, so by increasing in scale of graph as shown in Figure 5.18b, the response was compensated (using PI controller) compared with Figure 5.12, there was no overshoot, and need 0.01second to reach steady state.

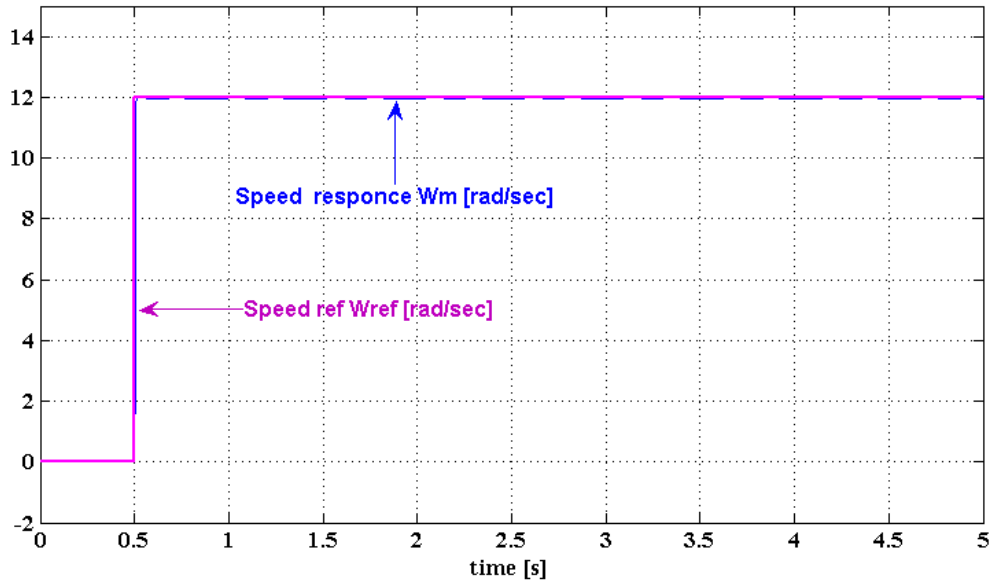


Fig 5.18a. Speed response of PMSM with PID controller.

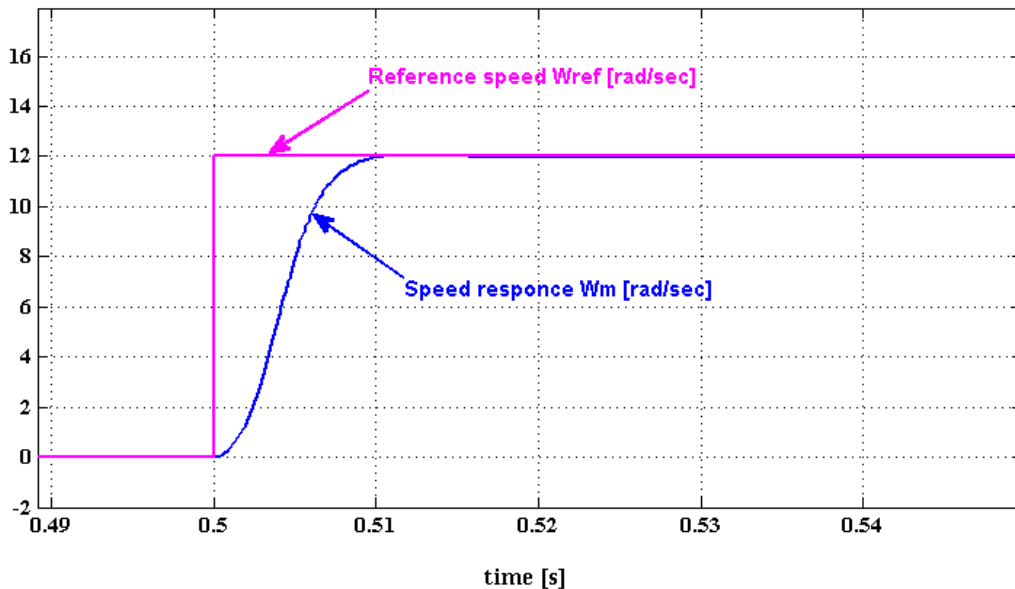


Fig 5.18b. Zoom in Speed Response of PMSM with PID controller.

6. CONCLUSIONS

The simulation of dynamic and transient performance of a 3-phase permanent magnet motor operating as a synchronous and BLDC motor was analyzed in this thesis. The motor performed quite differently under these two different schemes. To examine these differences, the software package MATLAB/SIMULINK was used to design the block diagrams and run the simulations. Models in both steady-state and dynamic conditions were taken into consideration. In addition, PI controller added to the synchronous and brushless dc motor; the results were then compared with those gotten in each case of model (BLDC motor, and PMSM) .

The BLDC motor, which operated with a self-controlled frequency, showed near consistencies when both of its models (in steady-state and dynamic conditions) were compared. The small discrepancies were attributed to several factors such as inductance.

As for the synchronous motor, both its models (steady-state and dynamic conditions) performance, also matched each other. However, the results a bit differed. These discrepancies were also attributed to several factors such as inductance.

Differences between the BLDC and synchronous motor (in their operation), were noted as follows.

1. The BLDC motor operates at a self-controlled frequency, meaning the motor itself imposes an appropriate value required for its actual speed. The motor should therefore be equipped with a speed or position sensor. The synchronous motor operates at set reference frequency and therefore has a constant speed.
2. At the transient response of the BLDC motor need 0.25sec to reach steady-state, and there is no overshoot occurred , but in PMSM model the transient response need 0.75sec to reach steady-state, and there is an overshoot occurred.
3. Looking at the characteristics, the efficiency of the BLDC motor first increases with a steep rise and then gradually decreases as the load torque is increased from no load. For the same condition. However, the synchronous motor's efficiency increases gradually. The equations that represent both motor models explain this difference. For example, the BLDC motor's efficiency depends on mechanical power and input power; at no load, the output power is zero (since it depends on load torque) leading to zero efficiency. As torque increases, both input and output power increase which leads to decreasing efficiency.

4. As mentioned before, the synchronous motor speed stays constant in any case of the conditions. However, particularly looking at the speed-torque characteristics, the BLDC motor speed decreases as the load torque is increased.
5. In building the SIMULINK models, the BLDC model had to be equipped with averaging blocks for the speed, power and torque; otherwise exact values could not have been read when the mentioned quantities reached steady-state (as the waveforms were disturbed). On the other hand, the synchronous motor needed no such block. When the power, torque and speed reached steady-state, the sinusoidal waveforms displayed definite values.

Although the two models operate quite differently, a few similarities were also noted as follows;

It is observed from the current-torque characteristics that both models have increasing current as the torque increases.

Conclusions drawn from this project are as follows.

1. The transient response for each particular abc model (BLDC, PMSM) show that the BLDC is faster than of PMSM to reach the steady-state.
2. The electromechanical characteristics of the BLDC and synchronous motor models show that regarding the average values, the motor behavior in steady-state is similar to that in dynamic conditions for each particular model. That means that the motors can be analyzed in this state, using the simpler steady-state model.
3. From the efficiency-torque characteristics, it can be concluded that in order to get increasing efficiency out of the machine (as the torque increases), the synchronous motor should be chosen. This is because, the BLDC motor has increasing efficiency as the simulation begins, it slowly decreases with increasing the load torque (as mentioned earlier).
4. About the synchronous motor operation, analysis of both the dq and ABC models shows that either one would yield the desired results and one's choice of which model to use would just depend on the application. In particular, the dq model should be used when the motor works with a control circuit. When operating in an open system, the ABC system can be used for analysis.
5. Based on the synchronous motor's behavior of performing at a constant speed and frequency, this motor model should then be chosen if one's application requires maintaining a constant speed on their motor.

6. When the motor works with a control circuit (PI controller), the synchronous motor's behavior is better than of BLDC motor behavior.

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APPENDIX A: M-FILES FOR BLDC MOTOR OPERATION

(a) M-file, *bldcplotter1* for plotting characteristics in Fig. 5.6.

```

TL=[2 4 6 8 10 12];
I=[6.25 12.5 20.1 26.2 32.5 38.25];
w=[58.02 24.4 17.59 12.79 8.907 4.676];
figure(1),
plot(TL,w,'ro',TL,I,'b*'),
xlabel(' Load Torque [Nm]'),gtext('speed [rad/sec]'),gtext('current [A]')

```

(b) M-file, *brush1* for plotting characteristics in Fig. 5.7.

```

TL=[2 4 6 8 10 12];
I=[6.25 12.5 20.1 26.2 32.5 38.25];
w=[38.02 24.4 17.59 12.79 8.907 4.676];
Pm=[76.03 97.6 105.54 102.32 89.07 56.112];
eff=[50 87.03 92 91.1 91 91 ];
figure(1),
plot(TL,w,'ro',TL,I,'ro',TL,Pm,'ro',TL,eff,'b*'),
xlabel(' Load Torque [Nm]'),gtext('Speed [rad/sec]'),gtext('Efficiency [%]'), gtext('Mechanical Power [w]'),
gtext('Current [A]')

```

APPENDIX B: M-FILES FOR SYNCHRONOUS MOTOR OPERATION

(a) M-file, *torque_delta* for plotting characteristics in Fig. 4.7.

```

Xs=2.61; V=155.54; Ef=2.093; w=20; delta=-pi:pi/100:pi;
T=(3*V*Ef*sin(delta))/(w*Xs);
figure(1),
plot(delta,T),grid
xlabel('delta [rad]'),
ylabel('Torque [N.m]'),

```

(b) M-file, *synchronousplot* for plotting characteristics in Fig. 5.17.

```

TL=[2 4 6 8 10 12];
W=[12 12.05 12.04 12.01 12 12.01 ];
I=[82.3 95.13 109.78 123.3 138.24 152.14];
Pm=[24 48.2 72.5 95.05 120 144.1];
eff =[29.3 56.6 84.85 91.1 91.2 91.08];
figure(1),
plot(TL,W,'ro',TL,I,'ro',TL,Pm,'ro',TL,eff,'b*')
xlabel(' Load Torque [Nm]'),gtext('Speed [rad/sec]'),gtext('Efficiency [%]'), gtext('Mechanical Power [w]'),
gtext('Current [A]')

```