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POSITION CONTROL OF AN INTERIOR PERMANENT MAGNET SYNCHRONOUS MOTOR BY USING ADAPTIVE BACK STEPPING ALGORITHM

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Abstract: This paper introduces a non linear position controller for an IPMSM (Interior Permanent Magnet Synchronous Motor). The system model equations provide the basis for the controller which is designed using adaptive back stepping technique. By recursive manner, virtual control states of the IPMSM drive have been identified and stabilizing control laws are developed subsequently using Lyapunov stability theory. Various system uncertainties are considered in the design. Using Lyapunov's stability theory, it is approved that the control variables are asymptotically stable. The complete model is then simulated using MATLAB/Simulink software. Performance of the controller is investigated at different dynamic operations such as command position change, sudden load change.

I. INTRODUCTION

The advances in power semiconductor technology, digital electronics, magnetic materials, and control theory have enabled modern ac motor drives to face challenging high-efficiency and high-performance requirements in the industrial sector. Among AC drives, the Interior Permanent Magnet Synchronous motor (IPMSM) is gaining popularity in industrial fields because of its advantages over other types of motors such as DC and Induction motors.
This is due to its:

- High power density,
- higher efficiency,
- low noise and,
- Robustness.[1]

These features are due to the incorporation of high-energy rare-earth alloys such as *neodymium–iron–boron* in its construction. In particular, the interior *permanent-magnet synchronous motor* (IPMSM) which has magnets buried in the rotor core exhibits certain good properties, such as mechanically robust rotor construction, rotor physical non saliency of the air gap, and small effective air gap. The rotors of these permanent magnet motors have complex geometry to ensure optimal use of the expensive permanent magnet material while maintaining a high magnetic field in the air gap. These features allow the IPMSM drive to be operated in high-speed mode by incorporating the field-weakening technique. Also, the mathematical model of the PMSM is less complex than that of an induction motor. However, the model is still non linear due to the coupling nature of the *d-q* axis currents (torque and flux components respectively) making the control task more complex.

Usually, high-performance motor drives require fast and accurate response, quick recovery from any disturbances, and insensitivity to parameter variations. The most effective method for high performance control of an IPMSM is vector control, or field oriented control[2]. This involves transforming the machine parameters from the standard *abc* frame to the synchronously rotating *d-q* frame using Park's transformation. Using the vector control technique, the torque and flux can be decoupled so that each can be controlled separately. This gives the IPMSM machine the highly desirable dynamic performance capabilities of the separately excited dc machine, while retaining the general advantages of the ac over dc motors. Originally, vector control was applied to the induction motor and a vast amount of research work has been devoted to this area. The vector control method is relevant to the IPMSM drive as the control is completely carried out through the stator, as the rotor excitation control is not possible.

However, precise speed control of an IPMSM drive becomes a complex issue owing to nonlinear coupling among its winding currents and the rotor speed as well as the nonlinearity present in the torque equation. The system nonlinearity becomes severe if the

IPMSM drive operates in the field weakening region where the direct axis current *id≠0*. This results in the appearance of a nonlinear term, which would have vanished under the existing vector control scheme with *id=0*.

There have been significant developments in non linear control theory applicable to electric motor drives. Interestingly, the $d - q$ transformation applicable to ac motors can be considered as a feedback linearization transformation. However, with the recent developments in nonlinear control theories, a modern control engineer has not only found a systematic approach in dealing with nonlinearities but has managed to develop approaches which had not been considered previously. The surge of such nonlinear control methods applicable to electromechanical systems include variable-structure systems[3], differential geometric approach [4], [5], and passivity theory [6].

As electrical drive systems possess well-defined nonlinear model characteristics, they have become good candidates for the application of newly developed nonlinear control techniques. If the knowledge of such nonlinearities can be included in the design of nonlinear controller, an enhanced dynamic behavior of the motor drive can be accomplished. One ingenious way of designing a nonlinear controller is adaptive back stepping (AB) [7]–[9].

Recently, the newly developed adaptive back stepping technique has been used in the design of speed and position controllers for dc, induction motors and permanent magnet motors [10]–[17].This technique allows the designer to incorporate most system nonlinearities and uncertainties in the design of the controller. In [12], [13], [15] the authors designed a nonlinear controller that achieves rotor angular speed and rotor flux amplitude tracking with uncertainties in the rotor resistance and load torque for an induction motor. Results show that tracking objectives are achieved with very little steady state error or overshoot. Zhou et. al [10] have developed a back stepping based controller for a DC motor and induction motor with uncertainties.

In this thesis, a nonlinear back stepping based controller is designed for position control of an IPMSM drive system. According to the proposed control technique, the controller is designed using the motor model equations incorporating various system uncertainties. The design method is similar to the one used in [17], however, the mechanical parameters such as motor inertia and load torque are estimated online instead of electrical parameters for better tracking response. The global system stability is verfied using Lyapunov stability theory. Digital simulations in MATLAB/Simulink verify the operation and stability of the controller and drive system.

II. MATHEMATICAL MODEL OF IPMSM

The mathematical model of the IPMSM can be given by the following equations in a synchronously rotating d-q reference frame as [19].

Voltage equations are given by:

$$v_d = Ri_d + L_d \frac{di_d}{dt} - P\omega_r L_q i_q \qquad 2.1$$

$$v_q = Ri_q + L_q \frac{di_q}{dt} + P\omega_r L_d i_d + P\omega_r \varphi_m \qquad 2.2$$

Arranging equations 2.1 and 2.2 in matrix form

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \begin{bmatrix} R + \frac{dL_d}{dt} & -P\varphi_r L_q \\ P\varphi_r & R + \frac{dL_q}{dt} \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} 0 \\ P\omega_r \varphi_m \end{bmatrix} \qquad 2.3$$

The developed motor torque is being given by

$$T_e = \frac{3P}{2}\left[\varphi_m i_q + (L_d - L_q) i_d i_q\right] \qquad 2.4$$

The mechanical Torque equation is given by

$$T_e = T_L + J\frac{d\omega_r}{dt} + B_m \qquad 2.5$$

$$\omega_r = \int \left(\frac{T_e - T_L - B\omega_r}{J}\right) dt \qquad 2.6$$

For convenience, the model is represented as follows.

$$\frac{di_d}{dt} = \frac{-Ri_d + P\omega_r L_q i_q}{L_d} + \frac{1}{L_d} v_d \qquad 2.7$$

$$\frac{di_q}{dt} = \frac{-Ri_q - P\omega_r L_d i_d - P\omega_r \varphi_m}{L_q} + \frac{1}{L_q} v_q \qquad 2.8$$

$$\frac{d\omega_r}{dt} = \frac{3P}{2J}\left[\varphi_m i_q + (L_d - L_q) i_d i_q\right] - \frac{T_L}{J} - \frac{B_m}{J}\omega_r \qquad 2.9$$

The above equations 2.7, 2.8, 2.9 are used to design the Simulink model of the machine.

Under balanced steady-state conditions, the electrical angular velocity of rotor ω_r is considered constant and equal to that of the synchronously rotating reference frame. In this mode of operation, with the time rate of change of all flux linkages neglected, the steady state versions of (2.7), (2.8) and (2.9) become

$$v_d = Ri_d - \omega_r L_q i_q \qquad 2.10$$

$$v_q = Ri_q + \omega_r (L_d i_d + \varphi_m) \qquad 2.11$$

$$v_o = Ri_0 = 0$$

Parks Transformation and Dynamic d q Modeling

The dynamic d q modeling is used for the study of motor during transient and steady state. It is done by converting the three phase voltages and currents to *dqo* variables by using Parks transformation [20]

Converting the phase voltages variables v_{abc} to v_{dqo} variables in rotor reference frame the following equations are obtained

$$\begin{bmatrix} v_d \\ v_q \\ v_o \end{bmatrix} = \frac{2}{3}\begin{bmatrix} \sin\theta_r & \sin(\theta_r - 120) & \sin(\theta_r - 120) \\ \cos\theta_r & \cos(\theta_r - 120) & \cos(\theta_r - 120) \\ 1/2 & 1/2 & 1/2 \end{bmatrix}\begin{bmatrix} v_d \\ v_q \\ v_o \end{bmatrix}$$

Convert V_{dqo} to V_{abc}

$$\begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix} = \begin{bmatrix} \cos\theta_r & \sin\theta_r & 1 \\ \cos(\theta_r - 120) & \sin(\theta_r) & 1 \\ \cos(\theta_r + 120) & \sin(\theta_r + 120) & 1 \end{bmatrix}\begin{bmatrix} v_q \\ v_d \\ v_o \end{bmatrix}$$

III. CONTROLLER DESIGN USING BACK STEPPING ALGORITHM

A. Idea of back stepping

By using newly developed back stepping algorithm we will design a controller. Back stepping is a recursive Lyapunov-based scheme proposed in the beginning of1990s. The technique was comprehensively addressed by Krstic, Kanellakopoulos and Kokotovic in [21]. The idea of back stepping is to design a controller recursively by considering some of the state variables as "virtual controls" and designing for them intermediate control laws. Back stepping achieves the goals of stabilization and tracking. The proof of these properties is a direct consequence of the recursive procedure, because a Lyapunov function is constructed for the entire system including the parameter estimates [21].

B. ***Controller design***

The foundation of back stepping is the identification of a virtual control variable and forcing it to become a stabilizing function. Thus, it generates a corresponding error variable which can be stabilized by proper selection input via Lyapunov's stability theory [17]. For position control, we regard the rotor speed and *d-q* axis currents as the virtual control variables. The block diagram of adaptive control scheme is shown below.

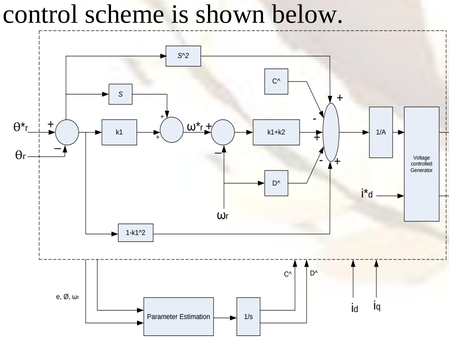


Fig3.1: Block diagram of adaptive control scheme

The following are the steps to design an adaptive controller

Step 1:

Define the position tracking error as

$e_1 = \theta^* - \theta$ 3.1

Where θ^* is the desired reference trajectory of the rotor angle. The position error dynamics is then

$\dot{e}_1 = \dot{\theta}^* - \dot{\theta} = \dot{\theta}^* - \omega_r$ 3.2

The stabilizing function is determined by differentiating the Lyapunov function: 2.10

$V_1 = \frac{1}{2}{e_1}^2$ to get 3.3

$\frac{dV_1}{dt} = e_1\dot{e}_1 = e_1(\dot{\theta}^* - \omega_r)$ 3.4 2.11

We now choose the first stabilizing function as

${\omega_r}^* = k_1 e_1 + \dot{\theta}^*$

Equation (3.4) indicates the desired velocity for position tracking. The next step is to design a speed controller so that the rotor speed will follow (3.4).

Step 2:

Now we Define the speed tracking error as

$e_2 = {\omega_r}^* - \omega = k_1 e_1 + \dot{\theta}^* - \omega_r$ 3.5

From equation (4.5), the position error dynamics can be written as

$\dot{e}_1 = -k_1 e_1 + e_2$ 3.6

The speed error dynamics is defined as

$\dot{e}_2 = \dot{\omega_r}^* - \dot{\omega}_r = -{k_1}^2 e_1 + k_1 e_2 + \ddot{\theta}^* - Ai_q - Bi_d i_q - C - D\omega_r$ 3.7

Now Define a new Lyapunov function as

$V_2 = \frac{1}{2}{e_1}^2 + \frac{1}{2}{e_2}^2$ 3.8

Differentiate to get

$\frac{dV_2}{dt} = e_1\dot{e}_1 + e_2\dot{e}_2$ 3.9

$= -k_1{e_1}^2 + e_2[(1 - {k_1}^2)e_1 + k_1 e_2 + \ddot{\theta}^* - Ai_q - Bidiq - C - D\omega r$ 3.10

Since i_d and i_q were identified as the virtual control variables, we Define the reference currents as

${i_q}^* = \frac{1}{A}[(1 - {k_1}^2)e_1 + (k_1 + k_2)e_2 + \ddot{\theta}^* - C - D\dot{\omega}_r]$ 3.11

${i_d}^* = 0$ 3.12

Substituting (3.11) and (3.12) back into equation (3.10) would yield

$\dot{V}_2 = -k_1{e_1}^2 - k_2{e_2}^2$ 3.13

Where $k_1, k_2 > 0$ are design constants.

Thus the virtual control is asymptotically stable. Since the parameters *C* and *D* are unknown we must use their estimated values $\hat{C}$ and $\hat{D}$ .Parameter *A* is assumed to be constant.

Thus equation (3.11)becomes

$\hat{i}_q^* = \frac{1}{A}\left[(1 - {k_1}^2)e_1 + (k_1 + k_2)e_2 + \ddot{\theta}^* - \hat{C} - \hat{D}\,\omega_r\right]$ 3.14

Step 3:

The goal now is to make i_d and i_q follow the reference trajectory i^*_d and i^*_q.

The final current error signals are defined as

$e_3 = \hat{i}_q^* - i_q$ 3.15

$e_4 = {i_d}^* - i_q$ 3.16

Using equations (3.15) and (4.16) the speed error dynamics can be represented by

$\dot{e_2} = -e_1 - k_2 e_2 + Ae_3 + Be_4 i_q + \tilde{C} + \tilde{D}\omega_r$ 3.17

Where $\tilde{C} = \hat{C} - C, and\ \tilde{D} = \hat{D} - D$ are the parameter estimation errors.
Now we define the current error dynamics as

$\dot{e_3} = \hat{i}_q^{*} - \frac{di_q}{dt}$ 3.18

$= \emptyset_4 + \tilde{A}\emptyset_5 + \tilde{C}\emptyset_6 + \tilde{D}\omega_r\emptyset_6 + Be_4 i_q\left(\emptyset_6 + \hat{D}\left(\frac{1}{A}\right)\right) - av_q$

3.19

And
$\dot{e_4} = -i_d = -\emptyset_2 - bv_d$ 3.20
Where $\emptyset_4, \emptyset_5\ and\ \emptyset_6$ are the known signals defined in the APPENDIX

Step 4:
The final Lyapunov function includes the current errors and parameter estimation errors.
$V_3 = \frac{1}{2}\left(e_1{}^2 + e_2{}^2 + e_3{}^2 + e_4{}^2 + \frac{1}{n_1}\tilde{C}^2 + \frac{1}{n_2}\tilde{D}^2\right)$ 3.21
Where n_1,n_2 are adaptive gains.
Now differentiate and substitute all error dynamic equations to get

$\dot{V_3} = e_1\dot{e_1} + e_2\dot{e_2} + e_3\dot{e_3} + e_4\dot{e_4} + \frac{1}{n_1}\tilde{C}\dot{\hat{C}} + \frac{1}{n_2}\tilde{D}\dot{\hat{D}}$ 3.22

$$= e_1(-k_1e_1 + e_2) + e_2(-e_1 - k_2e_2 + Ae_3 + Be_4 i_q + \tilde{C} + D\omega_r) + e_3(\emptyset_4 + \tilde{A}\emptyset_5 + \tilde{C}\emptyset_6 + \tilde{D}\omega_r\emptyset_6 - av_q) + e_4(-\emptyset_2 - bv_d) + \frac{1}{n_1}\tilde{C}\dot{\hat{C}} + \frac{1}{n_2}\tilde{D}\dot{\hat{D}}$$

$= -k_1e_1{}^2 - k_2e_2{}^2 - k_3e_3{}^2 - k_4e_4{}^2 + Ae_2e_3 + \tilde{C}\left(e_2 + \emptyset_6 e_3 + \frac{1}{n_1}\dot{\hat{C}}\right) + \tilde{D}\left(e_2\omega_r + \emptyset_6 e_3\omega_r + \frac{1}{n_2}\dot{\hat{D}}\right) + e_3(k_3e_3 + \emptyset_4 - av_q) + e_4\left(k_4e_4 + Bi_q\left(e_2 + \emptyset_6 + \right.\right.$

$\hat{D}$ $1A$−Ø2−bvq

The d-q axis reference voltages are chosen to be
$v_q{}^* = \frac{1}{a}(k_3e_3 + \emptyset_4)$ 3.23
$v_d{}^* = \frac{1}{b}\left(k_4e_4 + Bi_q\left(e_2 + \emptyset_6 + \hat{D}\left(\frac{1}{A}\right)\right) - \emptyset_2\right)$ 3.24
Where k_3, k_4> 0 are design constants.

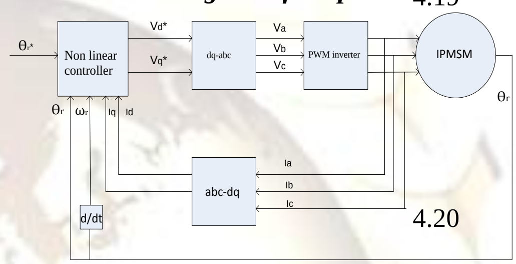
C. *Parameter estimation laws* 4.17
The update laws are defined as
$\dot{\hat{C}} = -n_1(e_2 + \emptyset_6 e_3)$ 3.25
$\dot{\hat{D}} = -n_2(e_2\omega_r + \emptyset_6 e_3\omega_r)$ 3.26
Substituting equation (3.22-3.24) into equation 3.21) would yield
$\dot{V_3} = -k_1e_1{}^2 - k_2e_2{}^2 - k_3e_3{}^2 - k_4e_4{}^2 + Ae_2e_3 < 0$
For sufficiently large k_2 and k_3 .Thus it is proven that the complete system is asymptotically stable.

IV. BLOCK DIAGRAM AND DRIVE SYSTEM

4 .18

A. *Block diagram of adaptive control scheme*

4.19



4.20

Fig.4.1.Block diagram of adaptive control scheme

B. *Drive system control*

Based on the control principle in chapter 4 the complete closed loop vector control scheme of the IPMSM is shown in Figure 4.1. First the command speed is calculated from the position error and derivative of the command position using equation3.4 as shown in chapter 3.The speed controller then generates the command *d-q* axis currents using equations (3.12) and (3.14) respectively. Parameters *C* and *D* are calculated using equations (3.24) and(3.25) and then the control voltages v_q and v_d are calculated using equations (3.22) and (3.23) respectively. Then they are converted to 3-phase voltages using Park's Transformation. The PWM signals are generated by comparing the 3-phasevoltages with high frequency triangular waveforms. The PWM logic signals operate the inverter switches which run the motor. The 3-phase currents i_{abc} are converted to *d-q* currents which are fed back into the controller along with the rotor speed and position which completes the closed loop system.

4.21

4.22

V. SIMULATION RESULTS AND DISCUSSION

A. Simulink Block of overall control system drive with sine wave trajectory

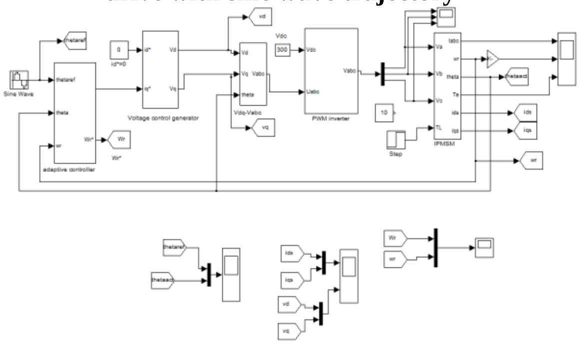


Fig.5.1. Simulink Block of overall control system drive with sine wave trajectory

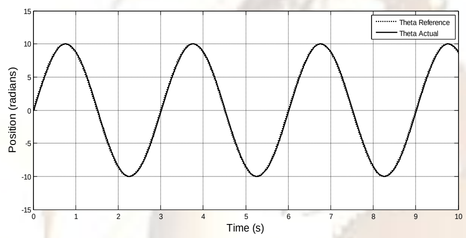


Fig.5.2. Command position and motor position

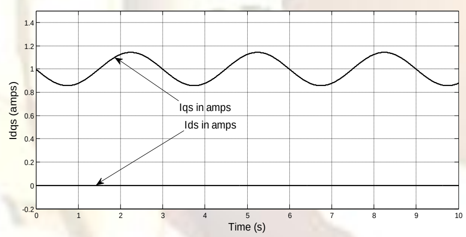


Fig.5.3. d-$q$axis currents i_{ds} and i_{qs}

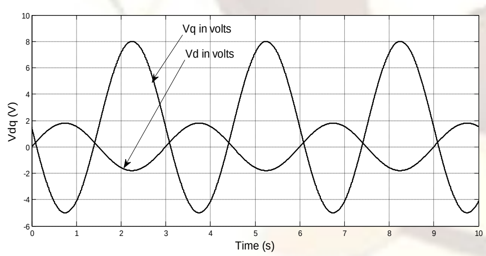


Fig.5.4. command voltages$V_d{}^*$and $V_q{}^*$

B. Simulink Block of overall control system drive with square wave trajectory

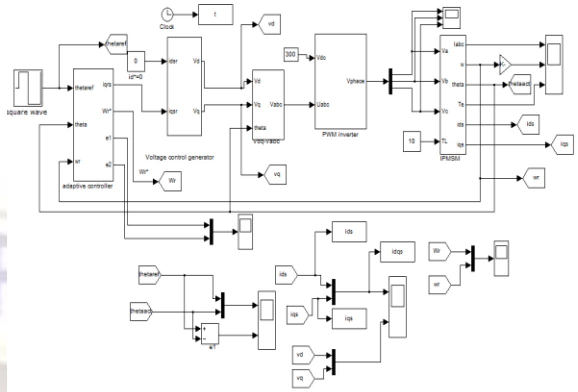


Fig.5.5.Simulink Block of overall control system drive with square wave trajectory

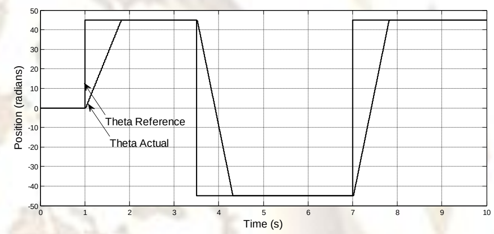


Fig.5.6 Command position and motor position

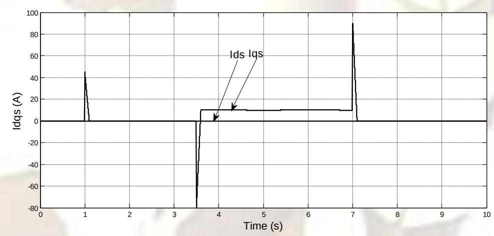


Fig.5.7.d-q axis currents i_{ds} and i_{qs}

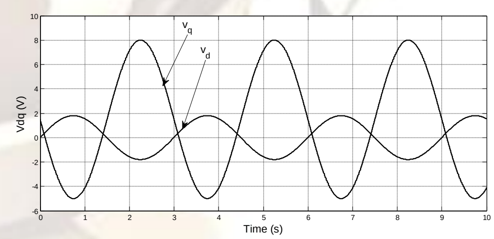


Fig.5.8. Command voltages$V_d{}^*$and $V_q{}^*$

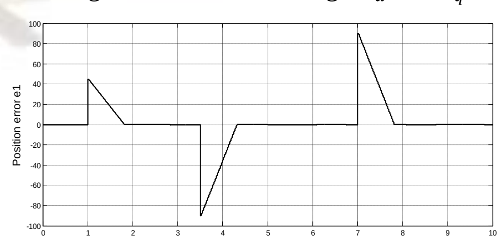


Figure 5.9: Corresponding position error e_1

COMMENTS:

In order to verify the effectiveness of the proposed adaptive scheme, digital simulations have been performed using MATLAB/Simulink software. The overall control bock diagram is shown in Figure 4.1 and a detailed block diagram of the adaptive controller is shown in Figure 3.1. Simulation work is done and some sample results are shown here.

Case1: Sine wave trajectory

Figure 5.1 shows the rotor position following the command position for a sinusoidal reference trajectory $\theta_r^* = 10\sin(2t)$.

The rotor position follows the reference trajectory with very little error and no overshoot as shown in fig.5.2. Figure 5.3 shows the *d-q* axis currents and figure 5.4 shows the command voltages V_d^*andV_q^*.

Case2: Square wave trajectory

Figure 5.5 shows the rotor position for a square wave trajectory. The actual position converges to the reference position in a short time with no overshoot and no steady state erroras shown in fig5.6. Fig.5.7 and 5.8 shows the corresponding currents and voltages. Fig 5.9 is the corresponding position error. The error between actual and command position is converge to zero. Since the control voltages guarantee asymptotic stability, the error values converge to zero.

By observing above *case1& and2,* we can say that using adaptive controller we can reduce the error between command and actual signals, and a good tracking can be achieved. So by using the adaptive control technique good tracking can be performed and accuracy of system will achieved.

V. CONCLUSION

Successful implementation of adaptive back stepping control for the position tracking of the IPMSM drive has been illustrated in this thesis. It has been shown that the IPMSM drive belongs to a class of nonlinear system for which adaptive back stepping technique can be effectively used. By recursive manner, virtual control states of the IPMSM drive have been identified and stabilizing control laws are developed subsequently using Lyapunov stability theory. The detailed derivation for the control laws has been provided. The controller was designed using adaptive back stepping with uncertainties of inertia, load torque and friction. As shown by the simulation results, position tracking was achieved with very little steady state error and overshoot. Global stability of the complete drive system has been verfied by Lyapunov stability theory. The validity of the proposed control technique has been established in simulation at different input conditions.

A. FUTURE SCOPE

In this thesis the control drive is implemented with adaptive controller for position tracking only. In future it can be implemented by considering speed also, and the hard ware implementation can be seen.

APPENDIX

$$\emptyset_1 = \frac{-Ri_q - P\omega_r L_d i_d - P\omega_r \varphi_m}{L_q}$$

$$\emptyset_2 = \frac{-Ri_d + P\omega_r L_q i_q}{L_d}$$

$$\emptyset_3 = (1 - k_1^{\,2})e_1 + (k_1 - k_2)e_2 + \ddot{\theta}^* - \hat{C} - \widehat{D}\omega_r$$

$$\emptyset_4 = \frac{1}{A}\Big[(1 - k_1^{\,2})(-k_1e_1 + e_2) + (k_1 + k_2)(-e_1 - k_2e_2) + \dddot{\theta}^* - \dot{\widehat{D}} - \widehat{D^2}\omega_r - \dot{\hat{C}} - \hat{C}\widehat{D} + (k_1 + k_2)e_3 - \widehat{D}i_q - \emptyset_1\Big]$$

$$\emptyset_5 = \frac{1}{A}\emptyset_3(k_1 + k_2)$$

$$\emptyset_6 = \frac{1}{A}(k_1 + k_2)$$

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