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New Sensorless Sliding Mode Control of a Five-phase Permanent Magnet Synchronous Motor Drive Based on Sliding Mode Observer

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ABSTRACT

This paper proposes a sensorless sliding mode control (SMC) for a five phase permanent magnet synchronous motor (PMSM) based on a sliding mode observer (SMO). The stability of the proposed strategy is proved in the sense of the Lyapunov theory. The sliding mode controller is designed with an integral switching surface and the sliding mode observer is developed for the estimation of rotor position and rotor speed. The proposed sensorless control strategy exhibits good dynamic response to disturbances. Simulation results are provided to prove the effectiveness of the proposed strategy.

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1. INTRODUCTION

Multiphase drives have acquired an increasing interest in the last years due to their advantages compared to three phase motor drives. Their main benefits can be summarized as follows: reducing the torque ripples thus multiphase machine are attractive solutions for applications where lower vibration is required, reducing the phase current without the increase in stator phase voltage. The Multiphase drives offer smoother torque due to the lower torque ripple magnitude and increase of the frequency of the torque pulsations simultaneously [1]. PMSM have become competitive to induction motors due to their attractive features such as high efficiency, low inertia, and power density [2]-[3]. One of the most popular control strategies of permanent magnet synchronous motor drive is the vector control proposed by Blascke. However, this control technique is sensitive to external disturbance and parameters uncertainties. To overcome this limitation, numerous nonlinear control strategies have been proposed in the literature such the backstepping control [4], input–output linearization control [2] and sliding mode control [5]-[7] and so on.

Recently, the sliding mode control technique has received worldwide interest. Indeed, this control strategy is robust to uncertainties and disturbance and also offer a stable control strategy and fast dynamic response [7]. Several methods of applying the SMC control to PMSM motor drives have been developed in the literature [7]-[8]. In [7], a new adaptive sliding mode control is designed to achieve the synchronization control of multiple three phase PMSM. A new adaptive law is proposed to alleviate the chattering despite the uncertainties and disturbances. In [8], a new sliding mode control is proposed for stabilization of multiple motor control. The proposed SMC proves its robustness to disturbances and parameters variation and the stability of the closed loop system is proved in the context of Lyapunov theory. The sliding mode control is applied to PMSM with good performance [9]-[10].

The synthesis of the sliding mode control requires the rotor speed and rotor position information. Thus, the installation of position and speed sensors is required. However, these transducers are expensive and sensitive to harsh environment conditions such as temperature and vibration. To discard this limitation, many research activities have focused on sensorless control techniques to estimate the speed and rotor position of the PMSM. The sensorless control strategy of PMSM drives are mainly classified into categories: the first type deals with high frequency injection and the second category is based on observer. Recently, the techniques based observer have gained an increasing interest such as SMO [11]-[16], Kalman filtering [17] and model reference system [2].

Compared to these approaches, the sliding mode observer has several features such as robustness to disturbance, order reduction control and simple algorithm implementation [18]. However, the main disadvantage of SMO is the chattering. To cope with limitation, many methods have been investigated [11]-[13]. In [11], a saturation function is used to replace the sign function but the insertion of additional position compensation and low pass filter causes phase delay which make the control of high performance applications unsatisfactory. To avoid the introduction of additional position compensation and low pass filter, some methods have been proposed [12]-[13]. In [12], an iterative SMO for sensorless vector control for a three-phase PMSM is proposed. The proposed SMO improves the performance in estimating the angle and the rotor speed by iteratively applying the observer in the sensorless PMSM Control. In [13], a novel SMO using the sigmoid function with variable boundary layer as switching function instead of the sign function to achieve high speed sensorless control for PMSM. The SMO proposed in [12]-[13] don't use the additional position compensation and low pass filter and the rotor position is estimated according to back EMF which may influence the accuracy of rotor position estimation.

In this paper, a new sliding mode observer is proposed to achieve the a new sliding mode control for the estimation of rotor position and rotor speed for a five phase PMSM under the assumption that only the stator voltages and stator currents are available for measurement. The controller of five phase PMSMS is proposed based sliding mode control strategy. The main contribution of this paper is the introduction of an appropriate combination of sliding mode control and sliding mode observer and its application to multiphase motor drive, indeed:

a. A new sliding mode control for a five phase PMSM is designed.
b. A new sliding mode observer of five phase PMSM is proposed.
c. The combination of these two methods.

This paper is organized as follows: the first section is devoted to introduction. Section 2 describes the PMSM system model, section 3 and section 4 deals with the development of sliding mode controller and sliding mode observer for five phase PMSM respectively. Section 5 is devoted to simulations results and finally, some conclusions are presented in the last section.

2. MODEL OF A FIVE-PHASE PMSM

The stator voltage of the five-phase PMSM in a natural stator frame is are given by [19]:

$$[V_s] = R_s [I_s] + \frac{d}{dt}[\Phi_s] + [E] \tag{1}$$

Where

$[V_s] = [v_1 \quad v_2 \quad v_3 \quad v_4 \quad v_5]^T$ is the stator voltage vector

$[I_s] = [i_1 \quad i_2 \quad i_3 \quad i_4 \quad i_5]^T$ is the stator current vector

$[\Phi_s] = [\phi_{s1} \quad \phi_{s2} \quad \phi_{s3} \quad \phi_{s4} \quad \phi_{s5}]^T$ is the stator flux vector

$[E] = [e_1 \quad e_2 \quad e_3 \quad e_4 \quad e_5]^T$ is the back electromotive force (EMF) vector

R_s is the stator resistance.

The control of five phase PMSM in a natural frame is not evident due to the magnetic coupling between phases. To simplify the control of five phase PMSM, the machine model can be transformed into a system of decoupled equations in a new frame. By applying the Concordia transformation and Park transformation, the equivalent model of the symmetrical five phase PMSM under assumptions about balancing operating conditions and linear magnetic circuits is presented in a rotating reference frame $\left(d_p q_p - d_s q_s\right)$ as [20]:

$$\begin{cases} \dfrac{dI_{dp}}{dt} = -\dfrac{R_s}{L_p} I_{dp} + \omega_e I_{qp} + \dfrac{1}{L_p} v_{dp} \\ \dfrac{dI_{qp}}{dt} = -\dfrac{R_s}{L_p} I_{qp} - \omega_e I_{dp} - \dfrac{\sqrt{\frac{5}{2}}\Phi_f \omega_e}{L_p} + \dfrac{1}{L_p} v_{qp} \\ \dfrac{dI_{ds}}{dt} = -\dfrac{R_s}{L_s} I_{ds} + 3\omega_e I_{qs} + \dfrac{1}{L_s} v_{ds} \\ \dfrac{dI_{qs}}{dt} = -\dfrac{R_s}{L_s} I_{qs} - 3\omega_e I_{ds} + \dfrac{1}{L_s} v_{qs} \\ \dfrac{d\Omega}{dt} = \dfrac{\sqrt{\frac{5}{2}} P \Phi_f}{J} I_{qp} - \dfrac{1}{J} T_L - \dfrac{f}{J}\Omega \end{cases} \Rightarrow \begin{cases} \dfrac{dI_{dp}}{dt} = f_{dp} + \dfrac{1}{L_p} v_{dp} \\ \dfrac{dI_{qp}}{dt} = f_{qp} + \dfrac{1}{L_p} v_{qp} \\ \dfrac{dI_{ds}}{dt} = f_{ds} + \dfrac{1}{L_s} v_{ds} \\ \dfrac{dI_{qs}}{dt} = f_{qs} + \dfrac{1}{L_s} v_{qs} \\ \dfrac{d\Omega}{dt} = f_\Omega \end{cases} \quad (2)$$

Where

$$\begin{cases} f_{dp} = a_1 I_{dp} + \omega_e I_{qp} \\ f_{qp} = a_1 I_{qp} - \omega_e I_{dp} + a_2 \omega_e \\ f_{ds} = a_3 I_{ds} + 3\omega_e I_{qs} \\ f_{qs} = a_3 I_{qs} - 3\omega_e I_{ds} \\ f_\Omega = a_4 I_{qp} + a_5 T_L + a_6 \Omega \end{cases} \quad (3)$$

With:

$$a_1 = -\frac{R_s}{L_p}; a_2 = -\frac{\sqrt{\frac{5}{2}}\Phi_f}{L_p}; a_3 = -\frac{R_s}{L_s}; a_4 = \frac{\sqrt{\frac{5}{2}} P \Phi_f}{J}; a_5 = -\frac{1}{J}; a_6 = -\frac{f}{J}$$

Where (I_{dp}, I_{qp}, I_{ds} , I_{qs}) and (v_{dp}, v_{qp}, v_{ds}, v_{qs}) are the stator currents and stator voltages in the $\left(d_p q_p - d_s q_s\right)$ frame. Φ_f is the flux linkage of permanent magnet and ω_e and Ω are the electrical and mechanical speed respectively.

In this new frame, the five phase PMSM can be expressed in two 2D planes $\left(d_p q_p\right)$ and $\left(d_s q_s\right)$ and one axis corresponding to the zero consequence component. This zero component is omitted as a result of star connection of the stator winding used in this work. L_p is the stator inductance of the main fictitious machine, L_s is the stator inductance of the secondary fictitious machine.

3. SLIDING MODE CONTROL

The sliding mode control is well adopted to nonlinear systems. The basic idea of sling mode control consist to choose a sliding surface and to force the system to converge to this surface. To achieve this, a

discontinuous control is used to maintain the dynamic of the state on the defined sliding surface despite uncertainties and disturbances. The design of sliding mode control occurs in three steps:
a. Selection of sliding surface.
b. Development of control law.
c. Determination of convergence conditions.

3.1. Sliding surface design

The sliding surface with integral action can be chosen as [6]:

$$s(t) = e(t) + q_i \int_0^t e(\tau)dt \tag{4}$$

Where *e(t)* and q_i (i=1,2,3,4,5) are the variable error and positive constant respectively. In this paper, one designs the currents and speed sliding mode controllers using the mathematical nonlinear five phase PMSM model described in (2). Using (4), five sliding surfaces with integral actions are considered as follows:

$$s_\Omega = e_\Omega + q_1 \int_0^t e_\Omega(\tau)dt \tag{5}$$

$$\begin{cases} s_{dp} = e_{dp} + q_2 \int_0^t e_{dp}(\tau)dt \\ s_{qp} = e_{qp} + q_3 \int_0^t e_{qp}(\tau)dt \\ s_{ds} = e_{ds} + q_2 \int_0^t e_{ds}(\tau)dt \\ s_{qs} = e_{qs} + q_5 \int_0^t e_{qs}(\tau)dt \end{cases} \tag{6}$$

Where e_Ω the speed error, and e_{dp}, e_{qp}, e_{ds},e_{qs} the d_p, q_p, d_s, q_s stator current errors components respectively, are given by:

$$\begin{cases} e_\Omega = \Omega - \Omega^* \\ e_{dp} = I_{dp} - I_{dp}^* \\ e_{qp} = I_{qp} - I_{qp}^* \\ e_{ds} = I_{ds} - I_{ds}^* \\ e_{qs} = I_{qs} - I_{qs}^* \end{cases} \tag{7}$$

Where ‘*’’indicates the corresponding variables are reference values.

3.2. Control law

In the next section, a new sliding mode control of a five phase PMSM is proposed based on the SMC built in [6] for three-phase induction motor. The synthesis of the sliding mode controller occurs in two steps:
a. Step 1: Speed controller
b. Step2: current controllers

3.2.1. Speed controller

The speed controller is designed to achieve the convergence of the speed error toward zero by constraining the convergence of the system to the sliding surface ($S_\Omega = 0$)

$$s_\Omega = 0 \Rightarrow \dot{s}_\Omega = \dot{e}_\Omega + q_1 e_\Omega = 0 \tag{8}$$

Using (2), (3) and (8), one obtains:

$$\begin{aligned} \dot{s}_{\Omega} &= \dot{\Omega} - \dot{\Omega}^{*} + q_1 e_{\Omega} \\ &= f_{\Omega} - \dot{\Omega}^{*} + q_1 e_{\Omega} \\ &= a_4 I_{qp} + a_5 T_L + a_6 \Omega - \dot{\Omega}^{*} + q_1 e_{\Omega} \end{aligned} \tag{9}$$

From (8) and (9), the equivalent I_{qp} stator current component can be deducted as follows:

$$I_{qpeq}{}^{*} = \left(\dot{\Omega}^{*} - a_5 T_L - a_6 \Omega - q_1 e_{\Omega}\right) / a_4 \tag{10}$$

In order to achieve suitable control performance despite uncertainties on the dynamic system, a discontinuous function called 'reaching controller' is required to be added to control part to cope with uncertainties across the sliding surface [7]. The reaching control is given by:

$$I_{qpn}^{*} = -k_1 \operatorname{sgn}(s_{\Omega}) \tag{11}$$

Where k_1 is a positive constant. Finally, the command q_{p} current component is given by:

$$I_{qp}{}^{*} = \left(\dot{\Omega}^{*} - a_5 T_L - a_6 \Omega - q_1 e_{\Omega}\right) / a_4 - k_1 \operatorname{sgn}(s_{\Omega}) \tag{12}$$

In this paper, five phase PMSM with sinusoidal back electromotive fore (EMF) is considered. That means that the electromagnetic torque is produced only by I_{qp} stator current component. So that d_{p}, d_s and q_s axis current components are controlled to be null [21].

$$\begin{cases} I_{dpeq}{}^{*} = 0 \\ I_{dseq}{}^{*} = 0 \\ I_{qseq}{}^{*} = 0 \end{cases} \tag{13}$$

3.2.2. Currents controllers

The current controllers are designed such a way to bring the system to follow the trajectory defined by the sliding surfaces $\left(s_{dp} = 0; s_{qp} = 0; s_{ds} = 0; s_{qs} = 0\right)$

$$\begin{cases} s_{dp} = 0 \\ s_{qp} = 0 \\ s_{ds} = 0 \\ s_{qs} = 0 \end{cases} \Rightarrow \begin{cases} \dot{s}_{dp} = 0 \\ \dot{s}_{qp} = 0 \\ \dot{s}_{ds} = 0 \\ \dot{s}_{qs} = 0 \end{cases} \tag{14}$$

Using (2), (3) and (6), one obtains:

$$\begin{cases} \dot{s}_{dp} = f_{dp} + \dfrac{1}{L_p} v_{dp} - \dot{I}^*_{dp} + q_2\left(I_{dp} - I^*_{dp}\right) \\ \dot{s}_{qp} = f_{qp} + \dfrac{1}{L_p} v_{qp} - \dot{I}^*_{qq} + q_3\left(I_{qp} - I^*_{qp}\right) \\ \dot{s}_{ds} = f_{ds} + \dfrac{1}{L_p} v_{ds} - \dot{I}^*_{ds} + q_2\left(I_{ds} - I^*_{ds}\right) \\ \dot{s}_{qs} = f_{qs} + \dfrac{1}{L_p} v_{qs} - \dot{I}^*_{qs} + q_3\left(I_{qs} - I^*_{qs}\right) \end{cases} \tag{15}$$

From (14) and (15), the equivalent stator voltages reference can be given by:

$$\begin{cases} v_{dpeq} = L_p\left(\dot{I}^*_{dp} - f_{dp} - q_2 e_{dp}\right) \\ v_{qpeq} = L_p\left(\dot{I}^*_{qp} - f_{qp} - q_3 e_{qp} - a_4 e_{\Omega}\right) \\ v_{dseq} = L_s\left(\dot{I}^*_{ds} - f_{ds} - q_4 e_{ds}\right) \\ v_{qseq} = L_s\left(\dot{I}^*_{qs} - f_{qs} - q_5 e_{qs}\right) \end{cases} \tag{16}$$

In order to cope with uncertainties, reaching controllers should be added to control part:

$$\begin{cases} v^*_{dpn} = -k_2 \operatorname{sgn}(s_{dp}) \\ v^*_{qpn} = -k_3 \operatorname{sgn}(s_{qp}) \\ v^*_{dsn} = -k_4 \operatorname{sgn}(s_{ds}) \\ v^*_{dsn} = -k_5 \operatorname{sgn}(s_{qs}) \end{cases} \tag{17}$$

Where k_i (i=2,3,4,5) are positive constants. Finally, the total reference voltages can be formulated from (16) and (17) as follow:

$$\begin{cases} v_{dp}{}^* = v_{dpeq}{}^* + v^*_{dpn} \\ v_{qp}{}^* = v_{qpeq}{}^* + v^*_{qpn} \\ v_{ds}{}^* = v_{dseq}{}^* + v^*_{dsn} \\ v_{qs}{}^* = v_{qseq}{}^* + v^*_{qsn} \end{cases} \tag{18}$$

Hence, the stator voltages reference can be expressed as:

$$\begin{cases} v_{dp}{}^* = L_p\left(\dot{I}^*_{dp} - f_{dp} - q_2 e_{dp}\right) - k_2 \operatorname{sgn}(s_{dp}) \\ v_{qp}{}^* = L_p\left(\dot{I}^*_{qp} - f_{qp} - q_3 e_{qp} - a_4 e_{qp}\right) - k_3 \operatorname{sgn}(s_{qp}) \\ v_{ds}{}^* = L_s\left(\dot{I}^*_{ds} - f_{ds} - q_4 e_{ds}\right) - k_4 \operatorname{sgn}(s_{ds}) \\ v_{qs}{}^* = L_s\left(\dot{I}^*_{qs} - f_{qs} - q_5 e_{qs}\right) - k_5 \operatorname{sgn}(s_{qs}) \end{cases} \tag{19}$$

3.3. Stability analysis

The objective in this section is the study of the stability of sliding mode control in closed loop. The aim is to achieve the convergence of the speed Ω and the stator current components I_{dp}, I_{qp}, I_{ds} and I_{qs} in

$\left(d_p q_p \text{ - } d_s q_s\right)$ frame to their reference values. The dynamic of the speed error is given by:

$$\dot{e}_\Omega = \dot{\Omega} - \dot{\Omega}^* = f_\Omega - \dot{\Omega}^* \tag{20}$$

The derivative of current components with respect to time can be given by:

$$\begin{cases} \dot{e}_{dp} = \dot{I}_{dp} - \dot{I}^*_{dp} = f_{dp} + \dfrac{1}{L_p} v_{dp} - \dot{I}^*_{dp} \\ \dot{e}_{qp} = \dot{I}_{qp} - \dot{I}^*_{qp} = f_{qp} + \dfrac{1}{L_p} v_{qp} \text{ - } a_4 e_{qp} - \dot{I}^*_{qp} \\ \dot{e}_{ds} = \dot{I}_{ds} - \dot{I}^*_{ds} = f_{ds} + \dfrac{1}{L_s} v_{ds} - \dot{I}^*_{ds} \\ \dot{e}_{qs} = \dot{I}_{qs} - \dot{I}^*_{qs} = f_{qs} + \dfrac{1}{L_s} v_{qs} - \dot{I}^*_{qs} \end{cases} \tag{21}$$

Using (12), the command q_p current can be written as:

$$I_{qp}{}^* = \left(\dot{\Omega}^* - a_5 T_L - a_6 \Omega - q_1 e_\Omega - k_\Omega \operatorname{sgn}(s_\Omega)\right) / a_4 \tag{22}$$

Where $k_\Omega = a_4 k_1$ Using (20) and (3), the derivative of speed error can be rewritten as:

$$\dot{e}_\Omega = a_4 I_{qp} + a_5 T_L + a_6 \Omega - \dot{\Omega}^* \tag{23}$$

Using (12) and (23), one obtains:

$$\dot{e}_\Omega = a_4 e_{qp} - q_1 e_\Omega - k_\Omega \operatorname{sgn}(s_\Omega) \tag{24}$$

From (19), the stator voltages can be rewritten as follows:

$$\begin{cases} v_{dp} = L_p\left(\dot{I}^*_{dp} \text{ - } f_{dp} - q_2 e_{dp} - k_{dp} \operatorname{sgn}(s_{dp})\right) \\ v_{qp} = L_p\left(\dot{I}^*_{qp} \text{ - } f_{qp} - q_3 e_{qp} \text{ - } a_4 e_{qp} - k_{qp} \operatorname{sgn}(s_{qp})\right) \\ v_{ds} = L_s\left(\dot{I}^*_{ds} \text{ - } f_{ds} - q_4 e_{ds} - k_{ds} \operatorname{sgn}(s_{ds})\right) \\ v_{qs} = L_s\left(\dot{I}^*_{qs} \text{ - } f_{qs} - q_5 e_{qs} - k_{qs} \operatorname{sgn}(s_{qs})\right) \end{cases} \tag{25}$$

With $k_{dp} = \dfrac{k_2}{L_p}, k_{qp} = \dfrac{k_3}{L_p}, k_{ds} = \dfrac{k_4}{L_p}$ and $k_{qs} = \dfrac{k_5}{L_p}$

Using (21) and (25), the derivative of current components with respect to time can be rewritten as:

$$\begin{cases} \dot{e}_{dp} = -q_2 e_{dp} - k_{dp} \operatorname{sgn}(s_{dp}) \\ \dot{e}_{qp} = -q_3 e_{qp} \text{ - } a_4 e_{qp} - k_{qp} \operatorname{sgn}(s_{qp}) \\ \dot{e}_{ds} = -q_4 e_{ds} - k_{ds} \operatorname{sgn}(s_{ds}) \\ \dot{e}_{qs} = -q_5 e_{qs} - k_{qs} \operatorname{sgn}(s_{qs}) \end{cases} \tag{26}$$

The study of this stability is carried out using the Lyapunov function. Let's consider the Lyapunov function to prove the stability of the proposed sliding mode controller.

$$V_c = \frac{1}{2}\left(e_\Omega^2 + e_{dp}^2 + e_{qp}^2 + e_{ds}^2 + e_{qs}^2\right) \tag{27}$$

The stability condition is assured under two conditions: -The Lyapunov function should be positive definite which is proved.- its derivative is be negative ($\dot{V}_c < 0$). The derivative of V_c with respect to time is given by:

$$\begin{aligned}\dot{V}_c &= e_\Omega\left(a_4 e_{qp} - q_1 e_\Omega - k_\Omega \operatorname{sgn}(s_\Omega)\right) + e_{dp}\left(-q_2 e_{dp} - k_{dp}\operatorname{sgn}(s_{dp})\right)\\ &\quad + e_{qp}\left(-a_4 e_\Omega - q_3 e_{qp} - k_{qp}\operatorname{sgn}(s_{qp})\right) + e_{ds}\left(-q_4 e_{ds} - k_{ds}\operatorname{sgn}(s_{ds})\right)\\ &\quad + e_{qs}\left(-q_5 e_{qs} - k_{qs}\operatorname{sgn}(s_{qs})\right)\end{aligned} \tag{28}$$

Then, the derivative of V_c can be expressed as:

$$\begin{aligned}\dot{V}_c &= -q_1 e_\Omega^2 - q_2 e_{dp}^2 - q_3 e_{qp}^2 - q_4 e_{ds}^2 - q_5 e_{qs}^2 - k_\Omega \operatorname{sgn}(s_\Omega) - k_{dp}\operatorname{sgn}(s_{dp})\\ &\quad - k_{qp}\operatorname{sgn}(s_{qp}) - k_{ds}\operatorname{sgn}(s_{ds}) - k_{qs}\operatorname{sgn}(s_{qs})\end{aligned} \tag{29}$$

To ensure the stability of $\dot{V}_c$, q_1, q_2, q_3, q_4, q_5 should be chosen such a way:

$$\begin{cases} q_1 >>> \left|k_\Omega \operatorname{sgn}(s_\Omega)\right| \\ q_2 >>> \left|k_{dp} \operatorname{sgn}(s_{dp})\right| \\ q_3 >>> \left|k_{qp} \operatorname{sgn}(s_{qp})\right| \\ q_4 >>> \left|k_{ds} \operatorname{sgn}(s_{ds})\right| \\ q_5 >>> \left|k_{qs} \operatorname{sgn}(s_{qs})\right| \end{cases} \tag{30}$$

So that, (30) can be written as follows:

$$\dot{V}_c < -q_1 e_\Omega^2 - q_2 e_{dp}^2 - q_3 e_{qp}^2 - q_4 e_{ds}^2 - q_5 e_{qs}^2 \tag{31}$$

It is to be noted that the sign function causes chattering so that it is replaced by sat function as defined in [18]. In the next section, a sliding mode observer will be developed for the estimation of rotor speed and rotor position for a five phase permanent magnet synchronous motor.

4. SLIDING MODE OBSERVER

The synthesis of sliding mode control requires the knowledge of speed and position and thus mechanical sensors such as resolvers or encoders must be installed. However, the presence of these transducers increases the cost of the system and decreases its reliability. In this section, we will propose an algorithm based on sliding mode observer for the estimation of rotor speed and rotor position under the assumptions of only the stator voltages and currents are measured. The principle of SMO applied to five phase PMSM is based on the stator current errors generated from their measured and estimated values that should converge to zero via selected sliding surface.

4.1. Design of the sliding mode observer

Based on (2), a sliding mode observer of five phase PMSM is designed from the sliding mode observer built in [15] for three-phase PMSM as follows:

$$\begin{pmatrix} \hat{I}_{dp} \\ \hat{I}_{qp} \\ \hat{I}_{ds} \\ \hat{I}_{qs} \end{pmatrix} = \begin{pmatrix} -\frac{R_s}{L_p} & \hat{\omega}_e & 0 & 0 \\ -\hat{\omega}_e & -\frac{R_s}{L_p} & 0 & 0 \\ 0 & 0 & -\frac{R_s}{L_s} & 3\hat{\omega}_e \\ 0 & 0 & 3\hat{\omega}_e & -\frac{R_s}{L_s} \end{pmatrix} \begin{pmatrix} \hat{I}_{dp} \\ \hat{I}_{qp} \\ \hat{I}_{ds} \\ \hat{I}_{qs} \end{pmatrix} - \begin{pmatrix} 0 \\ \frac{\sqrt{\frac{5}{2}}\Phi_f \hat{\omega}_e}{L_p} \\ 0 \\ 0 \end{pmatrix} + \begin{pmatrix} \frac{1}{L_p} & 0 & 0 & 0 \\ 0 & \frac{1}{L_p} & 0 & 0 \\ 0 & 0 & \frac{1}{L_s} & 0 \\ 0 & 0 & 0 & \frac{1}{L_s} \end{pmatrix} \begin{pmatrix} v_{dp} \\ v_{qp} \\ v_{ds} \\ v_{qs} \end{pmatrix} + K_o S + \Phi sign(S) \tag{32}$$

Where $\begin{pmatrix} \hat{I}_{dp} & \hat{I}_{qp} & \hat{I}_{ds} & \hat{I}_{qs} \end{pmatrix}^T$: the estimated current and $\hat{\omega}_e$ the estimated rotor speed. Φ and K_o are the matrices gains of the observer and sign(x) is the sign function. '.' is used to design the derivative with respect to time. The matrices of gains Φ and K_o are given by:

$$K_o = \begin{pmatrix} k_{o1} & 0 & 0 & 0 \\ 0 & k_{o2} & 0 & 0 \\ 0 & 0 & k_{o3} & 0 \\ 0 & 0 & 0 & k_{o4} \end{pmatrix} \text{ and } \Phi = \begin{pmatrix} \phi_1 & 0 & 0 & 0 \\ 0 & \phi_2 & 0 & 0 \\ 0 & 0 & \phi_3 & 0 \\ 0 & 0 & 0 & \phi_4 \end{pmatrix}$$

S is the sliding surface given by:

$$S = \begin{pmatrix} s_1 \\ s_2 \\ s_3 \\ s_4 \end{pmatrix} = \begin{pmatrix} I_{dp} - \hat{I}_{dp} \\ I_{qp} - \hat{I}_{qp} \\ I_{ds} - \hat{I}_{ds} \\ I_{qs} - \hat{I}_{qs} \end{pmatrix}$$

4.2. Stability analysis

The stability of the sliding mode observer is achieved by its convergence toward the sliding surface. To prove the stability of the proposed sliding mode observer, let us choose the Lyapunov function as [15]:

$$V = \frac{1}{2} S^T S + \frac{1}{2} \frac{\tilde{\omega}_e^2}{\eta_1} \tag{33}$$

Where $\tilde{\omega}_e$ is the speed estimation error given by:

$$\tilde{\omega}_e = \omega_e - \hat{\omega}_e \text{ and } \eta_1 > 0$$

According to theorem of Lyapunov stability. The sliding mode condition is achieved to ensure the condition bellow:

$$\dot{V} < 0 \tag{34}$$

Where $\dot{V}$ is the derivative with respect to time of V. Assuming that the rotor speed is constant in a small sampling interval, the derivative of (33) is given by:

$$\dot{V} = S^T \dot{S} - \frac{\tilde{\omega}_e}{\eta_1} \dot{\hat{\omega}}_e \tag{35}$$

Where

$$\dot{S} = \begin{pmatrix} \dot{I}_{dp} - \dot{\hat{I}}_{dp} \\ \dot{I}_{qp} - \dot{\hat{I}}_{qp} \\ \dot{I}_{ds} - \dot{\hat{I}}_{ds} \\ \dot{I}_{qs} - \dot{\hat{I}}_{qs} \end{pmatrix} = \begin{pmatrix} -\frac{R_s}{L_p} & \omega_e & 0 & 0 \\ -\omega_e & -\frac{R_s}{L_p} & 0 & 0 \\ 0 & 0 & -\frac{R_s}{L_s} & 3\omega_e \\ 0 & 0 & -3\omega_e & -\frac{R_s}{L_s} \end{pmatrix} \begin{pmatrix} I_{dp} \\ I_{qp} \\ I_{ds} \\ I_{qs} \end{pmatrix} - \begin{pmatrix} 0 \\ \frac{\sqrt{\frac{5}{2}}\Phi_f \omega_e}{L_p} \\ 0 \\ 0 \end{pmatrix} + \begin{pmatrix} \frac{1}{L_p} & 0 & 0 & 0 \\ 0 & \frac{1}{L_p} & 0 & 0 \\ 0 & 0 & \frac{1}{L_s} & 0 \\ 0 & 0 & 0 & \frac{1}{L_s} \end{pmatrix} \begin{pmatrix} v_{dp} \\ v_{qp} \\ v_{ds} \\ v_{qs} \end{pmatrix}$$

$$- \begin{pmatrix} -\frac{R_s}{L_p} & \hat{\omega}_e & 0 & 0 \\ -\hat{\omega}_e & -\frac{R_s}{L_p} & 0 & 0 \\ 0 & 0 & -\frac{R_s}{L_s} & 3\hat{\omega}_e \\ 0 & 0 & -3\hat{\omega}_e & -\frac{R_s}{L_s} \end{pmatrix} \begin{pmatrix} \hat{I}_{dp} \\ \hat{I}_{qp} \\ \hat{I}_{ds} \\ \hat{I}_{qs} \end{pmatrix} + \begin{pmatrix} 0 \\ \frac{\sqrt{\frac{5}{2}}\Phi_f \hat{\omega}_e}{L_p} \\ 0 \\ 0 \end{pmatrix} - \begin{pmatrix} \frac{1}{L_p} & 0 & 0 & 0 \\ 0 & \frac{1}{L_p} & 0 & 0 \\ 0 & 0 & \frac{1}{L_s} & 0 \\ 0 & 0 & 0 & \frac{1}{L_s} \end{pmatrix} \begin{pmatrix} v_{dp} \\ v_{qp} \\ v_{ds} \\ v_{qs} \end{pmatrix}$$

$$- K_o S - \Phi sign(S)$$

Hence:

$$\dot{S} = S^T (A - K_o) S - \tilde{\omega}_e \begin{pmatrix} \hat{I}_{qp} \\ -\hat{I}_{dp} - \sqrt{\frac{5}{2}} \frac{\Phi_f}{L_p} \\ 3\hat{I}_{qs} \\ -3\hat{I}_{ds} \end{pmatrix} - S^T \Phi sign(S) \tag{36}$$

Where

$$A = \begin{pmatrix} -\frac{R_s}{L_p} & \omega_e & 0 & 0 \\ -\omega_e & -\frac{R_s}{L_p} & 0 & 0 \\ 0 & 0 & -\frac{R_s}{L_s} & 3\omega_e \\ 0 & 0 & -3\omega_e & -\frac{R_s}{L_s} \end{pmatrix}$$

Setting (36) in (35), one obtains:

$$\dot{V} = S^T\left(A-K_o\right)S - \tilde{\omega}_e\left(\hat{I}_{qp}\tilde{I}_{dp} - \hat{I}_{dp}\tilde{I}_{qp} - \sqrt{\frac{5}{2}}\frac{\Phi_f}{L_p}\tilde{I}_{qp} - \frac{\tilde{\omega}_e}{\eta_1} + 3\hat{I}_{qs}\tilde{I}_{ds} - 3\hat{I}_{ds}\tilde{I}_{qs}\right) - S^T\Phi sign\left(S\right) \quad (37)$$

To guarantee the stability of V , $\dot{V}$ should be negative, so that from (37):

$$S^T\left(A-K_o\right)S < 0 \quad (38)$$

$$S^T\Phi sign\left(S\right) > 0 \quad (39)$$

$$\hat{I}_{qp}\tilde{I}_{dp} - \hat{I}_{dp}\tilde{I}_{qp} - \sqrt{\frac{5}{2}}\frac{\Phi_f}{L_p}\tilde{I}_{qp} - \frac{\tilde{\omega}_e}{\eta_1} + 3\hat{I}_{qs}\tilde{I}_{ds} - 3\hat{I}_{ds}\tilde{I}_{qs} = 0 \quad (40)$$

From (39), the gains ϕ_1, ϕ_2, ϕ_3 and ϕ_4 in the matrix Φ should be positive. (38) stipulates that the eigenvalues of $\left(A-K_o\right)$ must be in the left half plane. So, the gains $k_{1o}, k_{2o}, k_{3o}, k_{4o}$ in the matrix K are selected using the pole placement method outlined in the next section. From (40), the rotor speed adaptation law mechanism can be obtained as follows:

$$\hat{\omega}_e = \eta_1\int\left(\hat{I}_{qp}\tilde{I}_{dp} - \hat{I}_{dp}\tilde{I}_{qp} + 3\hat{I}_{qs}\tilde{I}_{ds} - 3\hat{I}_{ds}\tilde{I}_{qs} - \sqrt{\frac{5}{2}}\frac{\Phi_f}{L_p}\tilde{I}_{qp}\right)dt \quad (41)$$

Typically, a PI mechanism is used in the speed adaptation law and the estimated speed can be given by:

$$\begin{aligned}\hat{\omega}_e = &K_p\left(\hat{I}_{qp}\tilde{I}_{dp} - \hat{I}_{dp}\tilde{I}_{qp} + 3\hat{I}_{qs}\tilde{I}_{ds} - 3\hat{I}_{ds}\tilde{I}_{qs} - \sqrt{\frac{5}{2}}\frac{\Phi_f}{L_p}\tilde{I}_{qp}\right)\\ &+ K_i\int\left(\hat{I}_{qp}\tilde{I}_{dp} - \hat{I}_{dp}\tilde{I}_{qp} + 3\hat{I}_{qs}\tilde{I}_{ds} - 3\hat{I}_{ds}\tilde{I}_{qs} - \sqrt{\frac{5}{2}}\frac{\Phi_f}{L_p}\tilde{I}_{qp}\right)dt\end{aligned} \quad (42)$$

Where K_p and K_i are the gains of the PI mechanism. The estimated rotor position can be deducted as:

$$\hat{\theta}_e = \int\hat{\omega}_e dt \quad (43)$$

4.3. Gain selection

The gain selection designed by the classical approach allows to select the observer poles proportional to motor poles. This approach enables the observer to be dynamically faster than the motor but is susceptible to noise. To avoid this problem, the observer poles are designed with imaginary parts identical to poles of motor but in the left in the complex plane [15]. From (2), the motor poles, which are the eigenvalues of A, are given by

$$A=\begin{pmatrix} -\frac{R_s}{L_p} & \omega_e & 0 & 0 \\ -\omega_e & -\frac{R_s}{L_p} & 0 & 0 \\ 0 & 0 & -\frac{R_s}{L_s} & 3\omega_e \\ 0 & 0 & -3\omega_e & -\frac{R_s}{L_s} \end{pmatrix}$$

So that the motor poles are given by:

$$\begin{cases} \lambda_{m1}=\lambda_{m2}=-\frac{R_s}{L_p}\pm i\omega_e \\ \lambda_{m3}=\lambda_{m4}=-\frac{R_s}{L_s}\pm 3i\omega_e \end{cases} \tag{44}$$

Where $\lambda_{m1,2}$ and $\lambda_{m3,4}$ are the poles of the main fictive machine and secondary fictive machine respectively. Moreover, the observer poles are governed by the eigenvalues of $\left(A-K_o\right)$.Taking into account that the poles of the observer are shifted in the left by K_o for the main fictive machine and secondary fictive machine respectively. The gains in the matrix K can be given by: The observer poles governed by the eigenvalues of $\left(A-K_o\right)$ and the eigenvalues are given by:

$$\begin{cases} \lambda_{mo1}=\lambda_{mo2}=\left(-\frac{R_s}{L_p}-k_{gain1}\right)\pm i\omega_e \\ \lambda_{mo3}=\lambda_{mo4}=\left(-\frac{R_s}{L_s}-k_{gain2}\right)\pm 3i\omega_e \end{cases} \tag{45}$$

Where $k_{gain1}=k_{o1}=k_{o2}$ and $k_{gain2}=k_{o3}=k_{o4}$ In order to reduce the chattering, the “sign(s)” term, used previously, is replaced by a term with softer variation related to saturation function “sat(s)” defined by:

$$sat(s)=\begin{cases} 1 \quad if\ s>\lambda \\ \frac{s}{\lambda} \quad if\ |s|\le\lambda \\ -1 \quad if\ s<\lambda \end{cases} \tag{46}$$

where, λ is a small positive constant representing the thickness of the boundary layer. The proposed sliding mode control based on sliding mode observer is shown in Figure 1 where I_{dp}, I_{qp}, I_{ds} and $I_{qs}$are the currents in the rotating frame, $\hat{\omega}_e, \hat{\theta}_e$ are the rotor speed and rotor angle respectively. ‘* ’ is used to indicate that the variables are reference values. The estimated speed obtained from the SMO and the reference speed are processed in the speed sliding mode controller to determine the referenceI_{qp}current. The commanded I_{dp}, I_{ds} and I_{qs} currents components are fixed to zero. Then, the reference and actual currents components in

(d_p q_p− d_s q_s) rotating frame are processed in the current sliding mode controllers to obtain as outputs the equivalent commanded voltages.

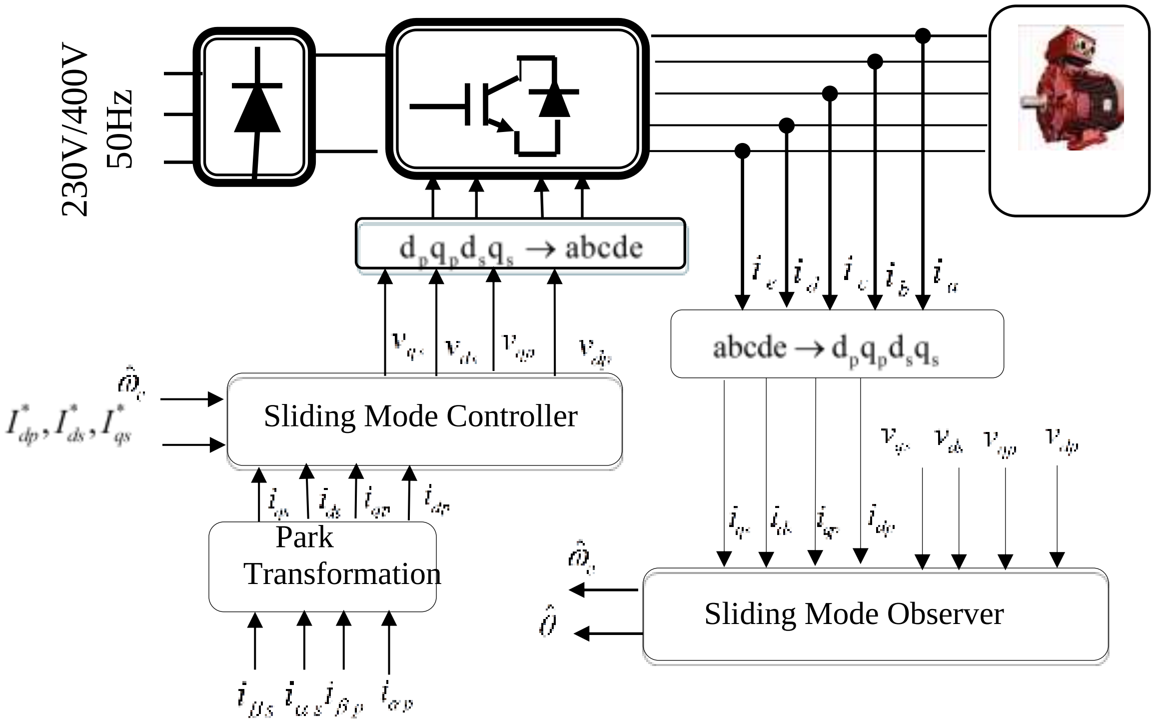


Figure 1. The block diagram of sensorless siding mode control of five-phase PMSM based on sliding mode observer

5. SIMULATIONS RESULTS

To confirm the validity of the proposed sensorless sliding mode control of five phase PMSM based on sliding mode observer. The overall block diagram shown in Figure 1 has been modeled using Matlab/Simulink Computer program. The motor parameters are shown in Table 1 [20]. The sliding mode control and sliding mode observer parameters are summarized in Table 2 and Table 3 respectively.

To highlight the performance of the proposed feedback sensorless control developed in the previous sections, the sliding mode control of five phase PMSM has been tested in three several situation. The first scenario was devoted to speed inversion and torque change. The second situation analyses the performance of feedback sensorless control when the motor is running at low speed. The last situation illustrates the response of the motor under step speed (sudden speed).

5.1. Scenario 1: Speed inversion and load torque change

In this situation, the response of the motor is illustrated under two different profiles. In the first case, the performance of five phase PMSM to change in reference speed is considered. Indeed, the reference speed is a pulse which is increased from standstill to rated value 1500 rpm and then it is increased to reach -1500 rpm at t=1s. The corresponding results are shown in Figure 2. The second case shows the drive performance following the application of a variable load torque under a fixed speed set at 1500 rpm. The load is a variable step which is equal to 10 N.mat t=0.5s and 6 N.m at t=1.5s. The waveforms of the response of the system to torque disturbance are illustrated in Figure 3.

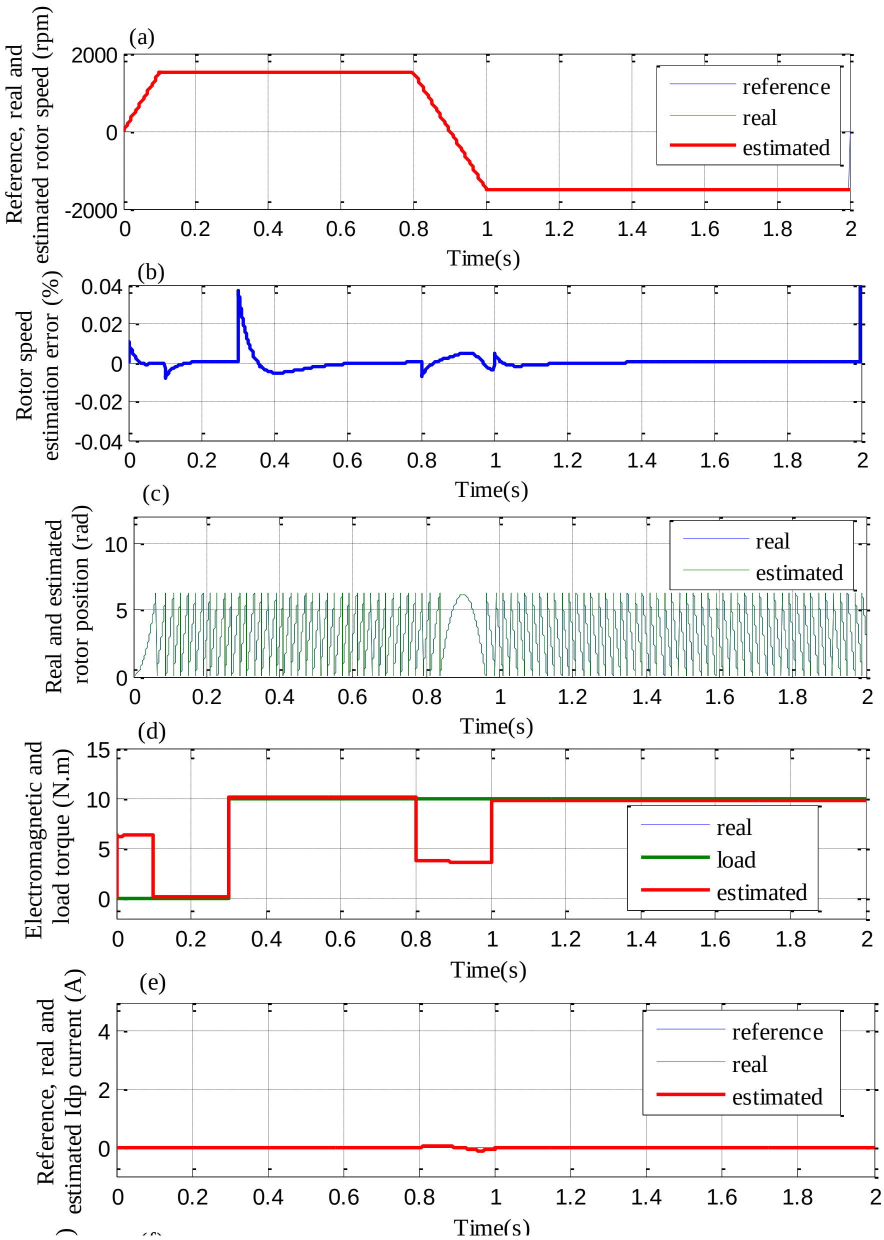
From Figure 2 and Figure 3, it is to be noted that the proposed sliding mode controller exhibits good dynamics tracking for speed and d_p, q_p, d_s,q_s stator currents components. Indeed, the speed and currents converge perfectly to their reference values in both steady state and dynamic transient which show good quality response. The SMO shows good speed control in the estimation of the rotor position and speed and the loading does not have significant effect on the tracking performance. Hence, the maximum speed estimation error in transient is approximately 0.04% as illustrated in Figure 2(b) and 0.05% as illustrated in Figure 3(b) for inversion speed and load torque change respectively. The speed estimation errors converge toward zero in steady state.

5.2. Scenario 2: low Speed

Figure 4 shows the performance of senseorless feedback control under low speed. The reference speed is a step fixed at 60 rpm. Figure 4 show that the SMC force the speed and currents to converge to their reference values perfectly and with excellent dynamics. Figure 4(a) shows the reference, actual and estimated speed and Figure 4(c) illustrates the actual and estimated rotor position. It can be seen that the maximum speed error is 0.5% at transient and converges toward zero in steady state as shown in Figure 4(b). In conclusion, the SMO shows good performance at low speed.

5.3. Scenario 3: Sudden Speed

Figure (5) displays the response of the proposed sensorless sliding mode control based on SMO under a step change in order to verify the robustness of the proposed sensorless control. The reference speed is a step which reach 1500 rpm at t=0s and reverse to reach -1500 rpm at t=1s. Figure 5(a) presents the reference, real and estimated speed and Figure 5(c) shows the estimated and actual rotor position. The corresponding simulations results prove that the rotor speed track its reference value with good dynamics. Indeed, the rotor speed estimation is approximately zero at steady state and the maximum estimation error is approximately 0.02% for rotor speed a as shown in Figure 5(b). It can be inferred that the SMO works well under sudden speed.



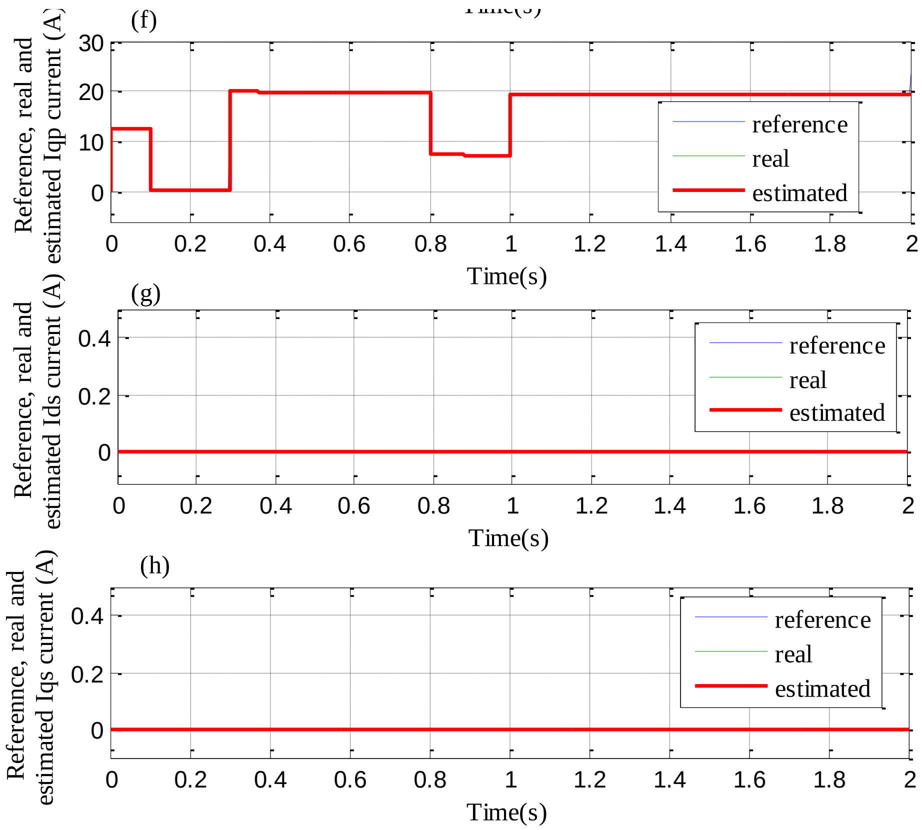
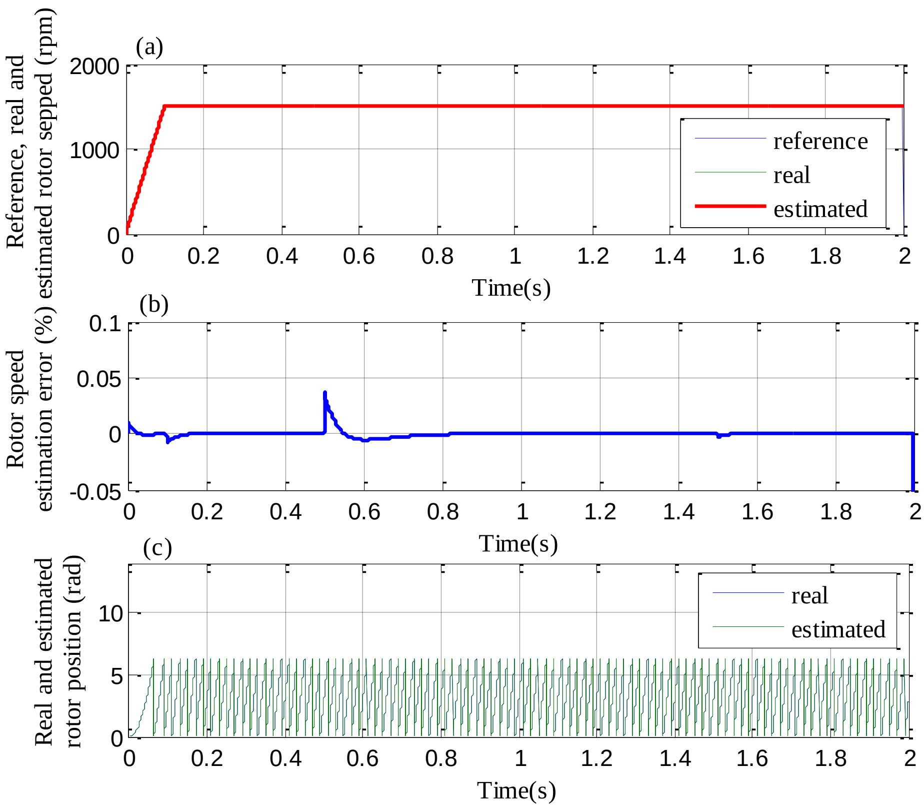


Figure 2. Sensitivity of the five phase PMSM performances at speed reverse: (a) reference, real and estimated rotor speed, (b) rotor speed estimation error, (c) real and estimated rotor position, (d) real and estimated electromagnetic torque, (e) real and estimated I_{dp} current, (f) real and estimated I_{qp} current, (g) real and estimated I_{ds} current , (h) real and estimated I_{qs} current



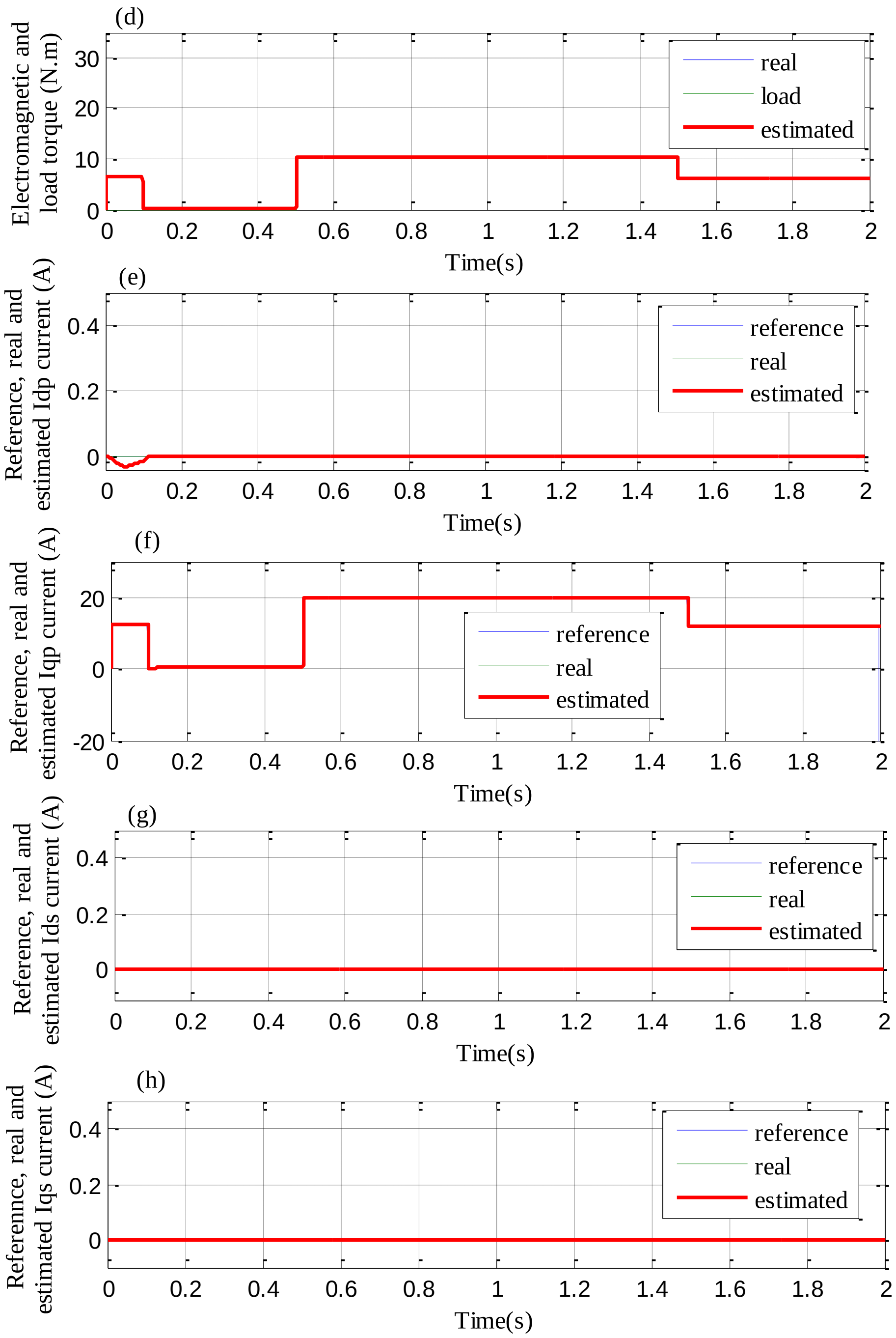
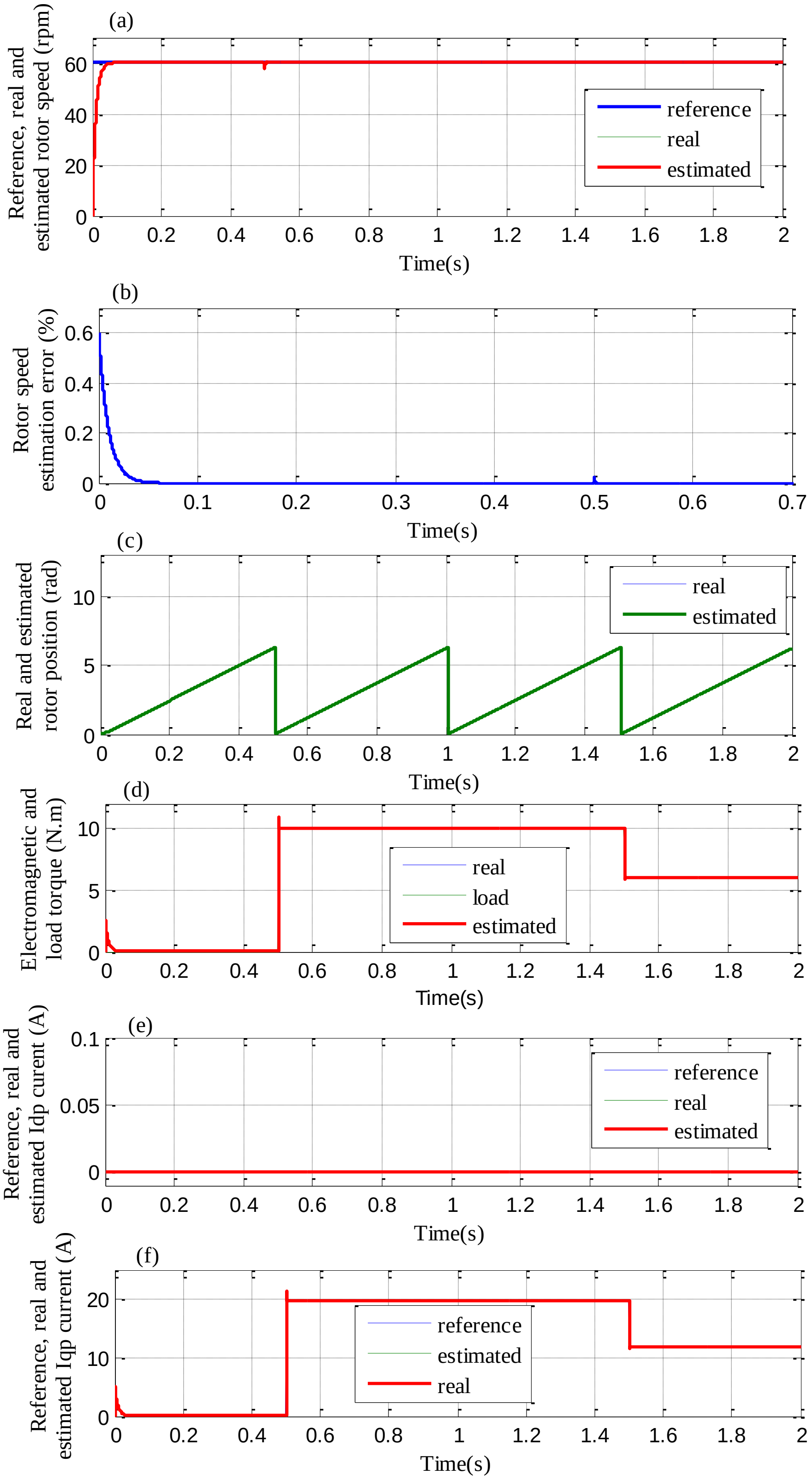


Figure 3. Sensitivity of the five phase PMSM performances at load disturbance: (a) reference, real and estimated rotor speed, (b) rotor speed estimation error, (c) real and estimated rotor position, (d) real and estimated electromagnetic torque, (e) real and estimated I_{dp} current, (f) real and estimated I_{qp} current, (g) real and estimated I_{ds} current , (h) real and estimated I_{qs} current


(a)
Reference, real and estimated rotor speed (rpm)
reference
real
estimated
Time(s)
(b)
Rotor speed estimation error (%)
Time(s)
(c)
Real and estimated rotor position (rad)
real
estimated
Time(s)
(d)
Electromagnetic and load torque (N.m)
real
load
estimated
Time(s)
(e)
Reference, real and estimated Idp curent (A)
reference
real
estimated
Time(s)
(f)
Reference, real and estimated Iqp current (A)
reference
estimated
real
Time(s)

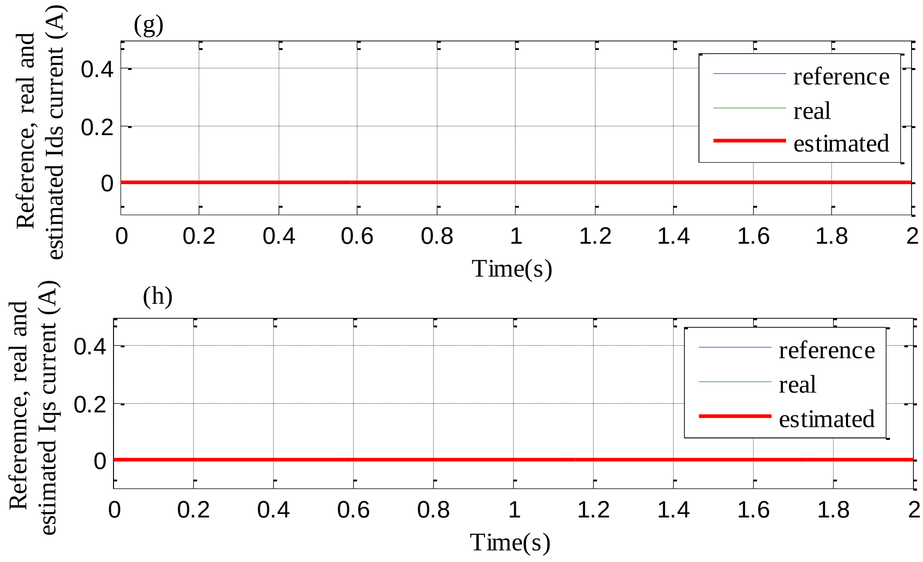
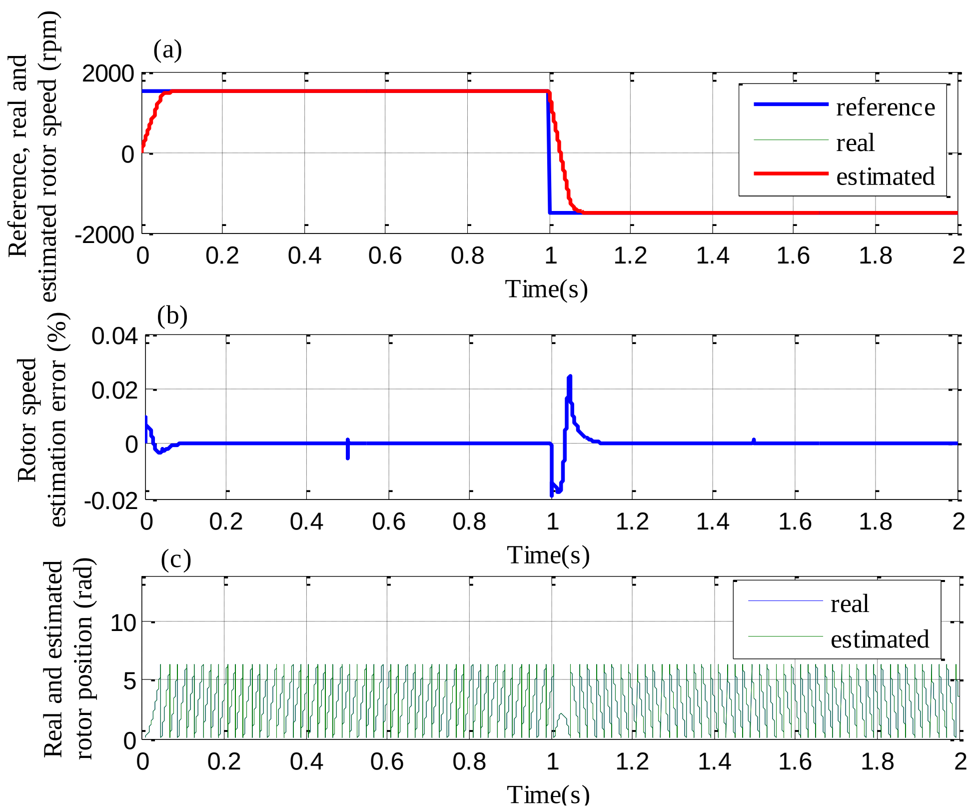


Figure 4. Sensitivity of the five phase PMSM performances at low speed:(a) reference, real and estimated rotor speed, (b) rotor speed estimation error, (c) real and estimated rotor position, (d) real and estimated electromagnetic torque, (e) real and estimated I_{dp} current, (f) real and estimated I_{qp} current, (g) real and estimated I_{ds} current , (h) real and estimated I_{qs} current



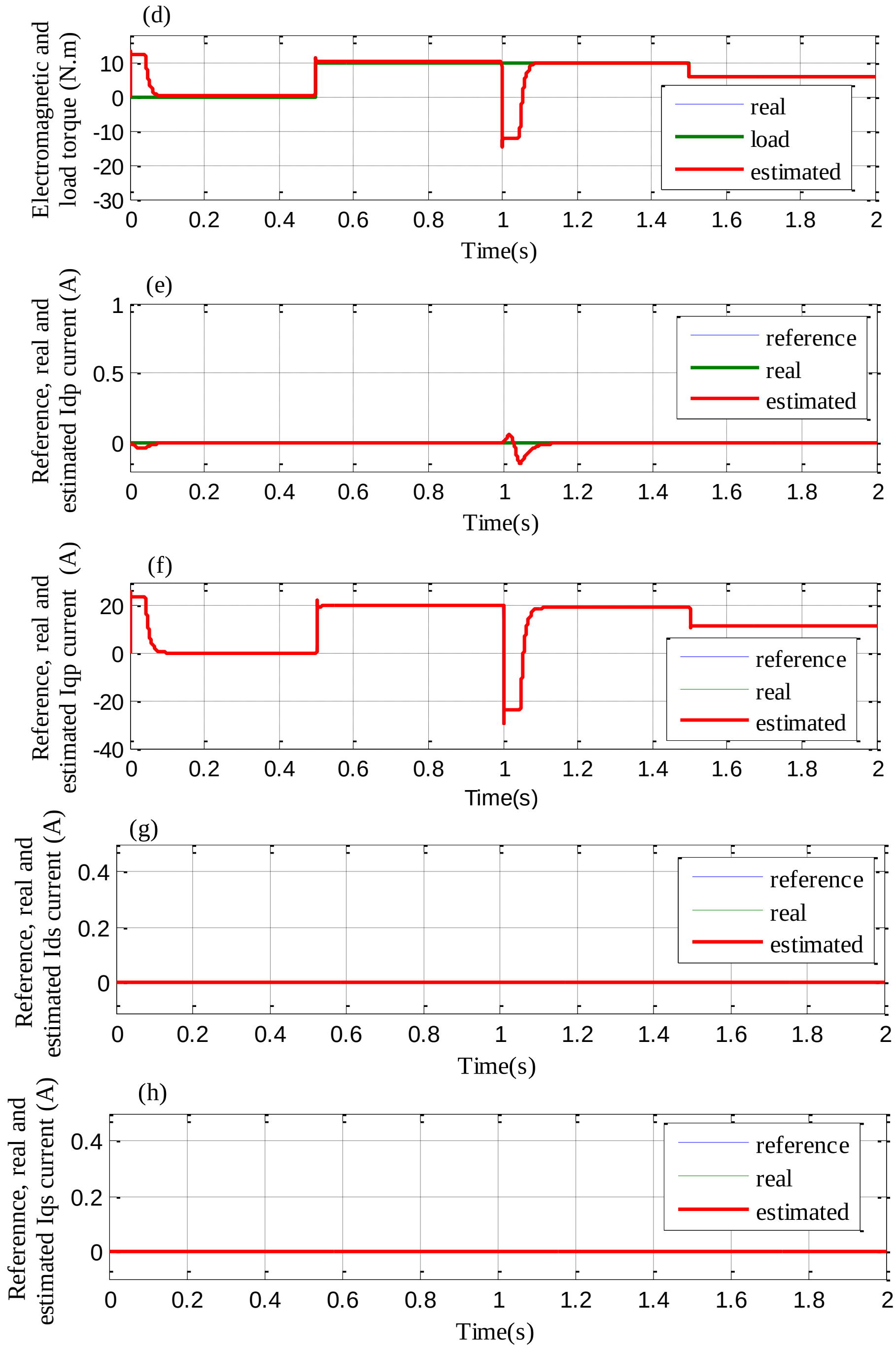


Figure 5. Sensitivity of the five phase PMSM performances at sudden speed: :(a) reference, real and estimated rotor speed, (b) rotor speed estimation error, (c) real and estimated rotor position, (d) real and estimated electromagnetic torque, (e) real and estimated I_{dp} current, (f) real and estimated I_{qp} current, (g) real and estimated I_{ds} current , (h) real and estimated I_{qs} current

5. CONCLUSIONS

This paper presented controlling aspect of five phase permanent magnet synchronous motor (PMSM) based on sensorless sliding mode control (SMC) with sliding mode observer (SMO) as speed estimator. A complete model of the AC drive system is numerically developed using simulation software and test for its transient and dynamic behaviors. It is verified from the investigation the proposed control technique offers good response to various perturbation conditions. Moreover, the proposed sliding mode observer improves the accuracy of speed estimation at very low speed variations without adaption of speed sensors. Developed sliding control algorithm satisfies the Lyapunov criterion under stability requirement for transient and dynamic behaviors. Proposed control scheme is applicable to AC drives that need to be adapted for sensitivity speed control in AC traction, Electric Vehicles and multi motor drives propulsion systems.

Table 1. Parameters of five-phase PMSM

L_p	L_s	P	Nominal speed
3.2 mH	0.93 mH	2	1500 rpm

Table 2. Parameters of sliding mode controller

k_1	k_2	k_3	k_4	k_5	q_1	q_2	q_3	q_4	q_5
0.1	250	200	40	30	50	600	680	60	60

Table 3. SMC Parameters of sliding mode observer

k_{o1}	k_{o2}	k_{o3}	k_{o4}	ϕ_1	ϕ_2	ϕ_3	ϕ_4
150	150	100	100	0.1	0.1	0.3	0.3

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