Adaptive Sliding Mode-MRAS Strategy for Sensorless Speed Control of SPIM Drives

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Abstract – *The objective of this paper is to provide sensorless control of Single Phase Induction Motor (SPIM) drives to improve the rotor speed and the stator currents. In order to achieve this high performance, adaptive Super Twisting Algorithm (STA) is used. The proposed method for the estimation of speed is based on Sliding Mode Model Reference Adaptive System (SM-MRAS) observer. An adaptive time varying switching gain is designed and adopted in order to cancel disturbance and uncertainties. To improve the estimator generated signal a discrete low pass filter is used. It represents a very simple design process compared to other chattering reduction methods as adding an observer. By using the Lyapunov approach the stability of the SM-MRAS speed estimation algorithm is proved. Simulation results prove the effectiveness of the proposed sensorless speed control algorithm.*

Key Words: Adaptive Sliding Mode-MRAS, Chattering, Super twisting algorithm, Single Phase Induction Motor (SPIM), Sensor-less sliding mode control.

1. INTRODUCTION

Variable Speed Drives (VSDs) applications are vastly used in the industry to control a wide range of speed and torque for machines, manufacturing process, pumps etc. VSDs have been integrated in several applications to accomplish one or more of the following objectives: energy saving, mechanical vibration reduction, power factor improvement, better coordination of motion on various shafts and production gains. In particular, the use of VSDs becomes recommended in many applications employing the Single Phase Induction Motors (SPIMs), such as blowers, washing machines, air conditioner, fans, compressors and pumps. Therefore, diverse control approaches have been proposed along the last years to drive the SPIM speed. Nowadays, Field-Oriented Controlled (FOC) induction motors are widely adopted to obtain high-dynamic performance in drive systems. Advanced controls such as Indirect Rotor Field- Oriented Control (IRFOC) need specific knowledge of the rotational speed information for feedback control. This information can be obtained via a mechanical sensor. Nowadays, many efforts are made to implement the sensorless control strategy to simplify the control structure and cut down the cost. In several applications different methods are applied to sensorless speed control of Induction Motor (IM) drive. The Extended Kalman Filter (EKF) has been extensively used for sensorless speed control because it has low-pass filter characteristics, and can extend the estimation of some parameters of the system and provides the best noise immunity. However, EKF is the most complicated to be implemented in DSP, instable due to linearization and erroneous parameters and its low speed performance is relatively poor. Compared to other methods, Sliding Mode Observer (SMO) method has attractive advantages of robustness to disturbances and low sensitivity to the system parameter variations. However, the chattering phenomenon limits the practical applications of conventional sliding mode observers. The Model Reference Adaptive System (MRAS) techniques are generally used for sensorless control of three-phase IM. In [7], the MRAS was compared with the EKF. It was found that the algorithm of MRAS is much simpler and faster. Due to their design simplicity, the speed estimator techniques based MRAS is widely used in the modernized industrial control. The drawback of MRAS is the large influence of parameter deviation at low- speed operation. In the other hand, this approach suffers from problems associated with pure integration, which limits the performance of the estimator at low values of stator frequency. In order to keep acceptable dynamic performances, simultaneous accurate knowledge or estimation of speed and at least one of motor parameters is required. As a result, classical MRAS algorithms become more complex and increase the installation cost which represents a severe constraint for real time implantation. Finally, in high speed, each observer can provide excellent performance but MRAS and SMO are more applicable in practical than EKF. Up to now, the speed estimation of field-oriented control SPIMs has been rarely presented so far. In fact, winding asymmetry in SPIM causes extra coupling between two stator windings and results in unbalanced machine operation. In the existing literature, some approaches have been suggested for speed sensorless SPIM, employing stator voltages and currents. In paper, the authors suggested to estimate the motor speed using rotor voltage vector which is defined in complex domain. The estimated speed is obtained from measurement only of q-axis stator current and that of reference generated by the control algorithm. Otherwise, in this method, both currents are measured and the machine parameters are required to implement the control system; therefore,

parameter variations can degrade speed control performances. Thus, some kind of parameter adaptation is required in order to obtain high-performance sensorless vector control drive. In paper [15], an observer based on two independent linear control systems provides the estimations of the rotor flux from the measurements of the stator currents and voltages is proposed. Consequently, this approach uses machine model equations that depend on some motor parameters; as a result an ordinary parameter variation can affect the estimator performances. However, in contrast to the above works; this paper adopts for the first time a Sliding Mode Model Reference Adaptive System (SM-MRAS) rotor speed observer. The proposed approach does not suffer from the problem of pure integration and low sensitivity to the system parameter variations. Consequently, the proposed SM-MRAS observer is capable of delivering high performances for a wide speed range, including very low speeds.

Conjointly, the SMC has been recognized as an appropriate technique for controlling nonlinear systems with uncertain dynamics, parameter variations and disturbances, thanks to its invariance and order reduction properties and simplicity design [17]-[18]. The main drawback of this technique is the well known chattering phenomenon caused by the excitation of un-modeled dynamics and sensor noise. Chattering may appear for on/off switches controlled systems if switching frequency is limited or the switching gain is inadequate. Particularly, in any SMC implementation, the chattering phenomenon causes harmful effects such as torque pulsation, current harmonics and acoustic noise [19], [20]. Mathematical analysis and numerical simulation have been performed to demonstrate the efficiency on chattering reduction of switching gain adaptation methods [19]-[22]. As well, it was noted that the gain adaptation techniques were not enough satisfactory when singly used in classical SMC algorithms. Hence, their experimental implementation associated with the use of a smoothing continuous function or LPF has been recommended in many advanced recent works [22], [23].

Therefore, this paper investigates the design and the DSP full implantation of an efficient robust sensorless speed vector control strategy devoted to drive a SPIM based SM technique and free of chattering. The synthesis of SOSM controllers based on an adaptive super twisting algorithm to control the speed and the stator currents, and a MRAS rotor speed estimation algorithm based sliding mode adaptation mechanism. Concerning the SM-MRAS rotor speed estimation, the adaptation mechanism is made by the association of an equivalent and discontinuous estimation laws. The equivalent law is synthesized from the desired sliding mode surface constraints; whereas a special care is given to the construction of switching gain of the discontinue term. Thus, a time varying switching

gain followed by a discrete Low Pass Filter (LPF) is considered for practical implementation. The proposed switching gain is defined to guaranties accurate rotor speed estimation and a good disturbances rejection. As it will be illustrated, this combination represents an excellent solution for chattering suppression. Finally, the SM-MRAS stability is verified via the Lyapunov approach.

2. SPIM State space model

The input–output SPIM state space model in a stationary reference frame (denoted by the superscript α, β) in term of the stator current and rotor flux can be given by

$$\frac{d}{dt}\begin{bmatrix} i_s \\ \varphi_r \end{bmatrix} = \begin{bmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{bmatrix} \begin{bmatrix} i_s \\ \varphi_r \end{bmatrix} + \begin{bmatrix} B \\ 0 \end{bmatrix} [v_s] \qquad (1)$$

Where $\quad i_s = [I_{S\alpha}, I_{S\beta}]T$, $\varphi_r = [\varphi_{r\alpha}, \varphi_{r\beta}]T$,

$$v_s = [v_{s\alpha}, v_{s\beta}]T$$

In addition, the matrix a_{ij} and B are

$$a_{11} = \begin{bmatrix} -\dfrac{1}{\sigma_d}\left(\dfrac{1}{\tau_{sd}} + \dfrac{1-\sigma_d}{\tau_r}\right) & 0 \\ 0 & -\dfrac{1}{\sigma_q}\left(\dfrac{1}{\tau_{sq}} + \dfrac{1-\sigma_q}{\tau_r}\right) \end{bmatrix}$$

$$a_{12} = \begin{bmatrix} \dfrac{1-\sigma_d}{\sigma_d \tau_r M_{srd}} & \dfrac{1-\sigma_d}{\sigma_d M_{srd}} w_r \\ \dfrac{1-\sigma_q}{\sigma_q M_{srq}} \omega_r & \dfrac{1-\sigma_q}{\sigma_q \tau_r M_{srq}} \end{bmatrix}$$

$$a_{21} = \begin{bmatrix} \dfrac{M_{srd}}{\tau_r} & 0 \\ 0 & \dfrac{M_{srq}}{\tau_r} \end{bmatrix}$$

$$a_{22} = \begin{bmatrix} -\dfrac{1}{\tau_r} & -\omega_r \\ \omega_r & -\dfrac{1}{\tau_r} \end{bmatrix}$$

$$B = \begin{bmatrix} \dfrac{1}{\sigma_d L_{sd}} & 0 \\ 0 & \dfrac{1}{\sigma_q L_{sq}} \end{bmatrix}$$

With $\quad \tau_{sd} = \dfrac{L_{sd}}{R_{sd}}, \qquad \tau_{sq} = \dfrac{L_{sq}}{R_{sq}}, \qquad \tau_r = \dfrac{L_r}{R_r},$

$$\sigma_d = 1 - \left(\frac{M_{srd}^2}{L_{sd}L_r}\right), \quad and \quad \sigma_q = 1 - \left(\frac{M_{srq}^2}{L_{sq}L_r}\right)$$

It is seen that there is an asymmetry in the model. Such an appropriate variable transformation of the stator variables is adopted in order to carry out a symmetric model for the FOC strategy use

$$\begin{cases} i_{s\alpha} = i_{s\alpha1} \\ i_{s\beta} = i_{s\beta1} \end{cases} \qquad \begin{cases} v_{s\alpha} = v_{s\alpha1} \\ v_{s\beta} = v_{s\beta1} \end{cases}$$

With : $K = \dfrac{M_{srd}}{M_{srq}}$.

While the mechanical and the electromagnetic torque expressions are, respectively, given by

$$J\mathring{\omega}_r + f\omega_r = n_p(T_e - T_l) \qquad (2)$$

$$T_e = \frac{n_p}{L_r}\left(M_{srq}i_{s\beta}\varphi_{r\alpha} - M_{srd}i_{s\alpha}\varphi_{r\beta}\right) \qquad (3)$$

Where, $v_{s\alpha}$, $v_{s\beta}$, $i_{s\alpha}$, $i_{s\beta}$, $\varphi_{r\alpha}$, and $\varphi_{r\beta}$ are the $\alpha - \beta$ axes voltages, currents, and fluxes of the stator and rotor in the stationary reference frame; L_{sd}, L_{sq}, L_r , M_{srd}, and M_{srq} denote the stator and rotor self and mutual inductances; R_{sd} , R_{sq}, and R_r denote $d - q$ axes stator and rotor resistances; ω_r, ω_{sl}, T_e , and T_l are the rotor angular frequency, the slip angular frequency, the electromagnetic torque, and the load torque; f, J , and n_p are the friction coefficient, the total inertia, and the pole pairs number.

2a. IRFOC strategy of SPIM

The block diagram of the proposed IRFOC control strategy associated to the SPIM is illustrated in Fig.2.1 The relationship between rotor-flux components and the stator currents in the synchronous reference frame (denoted by the superscript *d, q)* is expressed as follows:

$$\frac{d}{dt}\begin{bmatrix}\varphi_{rd} \\ \varphi_{rq}\end{bmatrix} = \begin{bmatrix} -\frac{1}{\tau_r} & \omega_{s1} \\ -\omega_{s1} & -\frac{1}{\tau_r} \end{bmatrix}\begin{bmatrix}\varphi_{rd} \\ \varphi_{rq}\end{bmatrix}\begin{bmatrix} \frac{M_{srd}}{\tau_r} & 0 \\ 0 & \frac{M_{srd}}{\tau_r} \end{bmatrix}\begin{bmatrix}i_{sd1} \\ i_{sq1}\end{bmatrix}$$

where

$$\begin{bmatrix}\varphi_{rd} \\ \varphi_{rq}\end{bmatrix} = T_s \left[\begin{bmatrix}\varphi_{r\alpha} \\ \varphi_{r\beta}\end{bmatrix}\right]$$

$$\begin{bmatrix}i_{sd} \\ i_{sq}\end{bmatrix} = T_s \begin{bmatrix}i_{s\alpha1} \\ i_{s\beta1}\end{bmatrix}$$

$$T_s = e^{j\theta_s} = \begin{bmatrix} cos\theta_s & sin\theta_s \\ -sin\theta_s & cos\theta_s \end{bmatrix}$$

And

$$\omega_{s1} = \omega_s - \omega_r$$

The $T_s = e^{j\theta_s}$ matrix performs the coordinate transformation from the reference frame aligned along with the stator-flux vector to the stationary reference frame. We choose a reference frame linked to the rotor flux, so that the *d*-axis coincides with the desired direction of the rotor flux ($\varphi_{rd} = \varphi_r$ and $\varphi_{rq} = 0$). The expression (2.10) can be decomposed into two equations

$$\rho\varphi_r^* = -\frac{1}{\tau_r}\varphi_r^* + \frac{M_{srd}}{\tau_r}i_{sd1}^* \qquad (4)$$

$$\omega_{s1}^* = \frac{M_{srd}}{\tau_r\varphi_r^*}i_{sq1}^* \qquad (5)$$

The expression for the torque can be calculated be

$$T_e = n_p \frac{M_{srd}}{L_r}\varphi_r^* i_{sq1}^* \qquad (6)$$

Fig.2.1 IRFOC block diagram associated to the SPIM via IGBTs inverter

3. Second order sliding mode speed and current controllers design

The proposed SOSMCs speed and current controllers design is based on the so-called STA. This algorithm ensures that the states can slide on the chosen sliding surface *S*. The control law is composed of two terms formulated around the sliding variables. The first one is the integral of a discontinue function, whereas the second term represents a continuous function to alleviate the chattering effect.

Let us consider the following sliding surfaces:

$$\begin{bmatrix} s_d \\ s_q \\ s_\omega \end{bmatrix} = \begin{bmatrix} c_d \\ c_q \\ c_\omega \end{bmatrix} \begin{bmatrix} \in_{sd} \\ \in_{sq} \\ \in_{s\omega} \end{bmatrix} = \begin{bmatrix} c_d \\ c_q \\ c_\omega \end{bmatrix} \begin{bmatrix} i_{sd} - i_{sd}^* \\ i_{sq} - i_{sq}^* \\ \omega_r - \omega_r^* \end{bmatrix} \qquad (7)$$

Taking in consideration that the electromagnetic torque reference represents the speed controller output in the IRFOC strategy, the first derivative forms of the errors are , as shown at the top of the next page. Adopting the following partial feedback linearization equation:

$$\begin{bmatrix} U_d \\ U_q \\ U_\omega \end{bmatrix} = \gamma \begin{bmatrix} v_{sd} - \omega_s L_{sd}\sigma_d i_{sq} \\ v_{sq} - \omega_s L_{sd}\sigma_d i_{sd} \\ L_{sd}\sigma_d \frac{n_p}{J} T_e \end{bmatrix} \text{ with } \gamma = \frac{1}{L_{sd}\sigma_d} \quad (8)$$

Therefore, the second errors derivative forms can be expressed as

$$\begin{bmatrix} \ddot{\in}_{sd} \\ \ddot{\in}_{sq} \\ \ddot{\in}_{s\omega} \end{bmatrix} = \gamma \begin{bmatrix} \dot{U_d} \\ \dot{U_q} \\ \dot{U_\omega} \end{bmatrix} + \begin{bmatrix} \dot{G_1} \\ \dot{G_2} \\ \dot{G_3} \end{bmatrix} \qquad (9)$$

With

$$\begin{bmatrix} G_1 \\ G_2 \\ G_3 \end{bmatrix} = \begin{bmatrix} \left(R_{sd} + \frac{M_{srd}^2}{L_r\tau_r}\right) i_{sd1} + \frac{M_{srd}}{L_r\tau_r} \varphi_r \\ \left(K^2 R_{sq} + \frac{M_{srd}^2}{L_r\tau_r}\right) i_{sq1} - \frac{M_{srd}}{L_r\tau_r} \omega\varphi_r \\ -L_{sd}\sigma_d \frac{f}{J} \omega_r \end{bmatrix} - \begin{bmatrix} i_{sd}^* \\ i_{sq}^* \\ \omega_r^* \end{bmatrix}$$

$$\begin{bmatrix} \dot{\in}_{sd} \\ \dot{\in}_{sq} \\ \dot{\in}_{\omega} \end{bmatrix} = \frac{1}{\sigma_d L_{sd}} \left(\begin{bmatrix} v_{sd} + \frac{M_{srd}}{L_r\tau_r} \varphi_r \\ v_{sq} - \frac{M_{srd}}{L_r} \omega\varphi_r \\ L_{sd}\sigma_d \frac{n_p}{J} T_e \end{bmatrix} - \begin{bmatrix} R_{sd} + \frac{M_{srd}^2}{L_r\tau_r} & \omega_s L_{sd}\sigma_d & 0 \\ -\omega_s L_{sd}\sigma_d & K^2 R_{sq} + \frac{M_{srd}^2}{L_r\tau_r} & 0 \\ 0 & 0 & -L_{sd}\sigma_d \frac{f}{J} \omega_r \end{bmatrix} \begin{bmatrix} i_{sd} \\ i_{sq} \\ \omega_r \end{bmatrix} \right) - \begin{bmatrix} i_{sd}^{*\cdot} \\ i_{sd}^{*\cdot} \\ \omega_r^* \end{bmatrix} \qquad (10)$$

We suppose that the terms γ , *G*1, *G*2, and *G*3 are bounded by known positive constants as

$$\begin{cases} \dot{G_i} \le \varphi_i; & i = 1,2,3 \\ 0 < m \le \gamma \le M \end{cases}$$

Adopting the following proposed control law:

$$\begin{cases} U_d = u_1 - \lambda_1 |\in_{sd}|^\rho Sign(\in_{sd}) \\ \quad \dot{u_1} = \alpha_1 Sign(\in_{sd}) \\ U_q = u_2 - \lambda_2 |\in_{sd}|^\rho Sign(\in_{sq}) \\ \quad \dot{u_2} = \alpha_2 Sign(\in_{sd}) \\ U_\omega = u_3 - \lambda_3 |\in_{sd}|^\rho Sign(\in_\omega) \\ \quad \dot{u_2} = \alpha_2 Sign(\in_{sd}) \end{cases}$$

Where ρ, α_i, and λ_i are the tuning controller parameters, which are positives constants respecting the following equations:

$$\begin{cases} \alpha_i > \frac{\varphi_i}{m} \\ \lambda_i^2 \ge \frac{4\varphi_i M(\alpha_i + \varphi_i)}{m^3(\alpha_i - \varphi_i)} \end{cases}$$

The parameter ρ is fixed at 0.5 to ensure that the maximal possible for two-sliding realization is achieved . Then, the proposed control law is capable of enforcing the occurrence of a SOSM on the designed sliding surface in finite time. It is known that the chattering level is proportional to the switching gain amplitude. Thus, the

used method for chattering reduction would be efficient if it may conduct the switching gain amplitude to converge to a small value and this without affecting the subsistence of sliding mode. To take care of probably parameter or load variations and the chattering reduction accentuation, a way can be considered resulting in the adjustment of the control gains according to the instantly sliding surface value and to its variation in time

$$\begin{cases} \alpha = \alpha(S, S^{.}, t) \\ \lambda = \lambda(S, S^{.}, t) \end{cases}$$

The adopted adaptive switching gains expressions are

$$\begin{cases} \dfrac{d}{dt}\alpha = \begin{cases} \omega_1\sqrt{\dfrac{\gamma_1}{2}}sign(\vdots s \vdots -\mu), & if\ \alpha > 0 \\ 0 \end{cases} \\ \qquad\qquad\qquad \lambda = 2 \in \alpha \end{cases}$$

Where $\omega_1, \gamma_1,\ \varphi,$ and $\in$ are arbitrary positive constants. Then, the adaptive STA will enforce the considered sliding surface and its first derivative to tend to zero in finite time, even at the presence of unknown disturbances and uncertainties with unknown boundary. To realize this time varying control gain, an integrator operator with external rest signal is performed. This procedure allows us keeping a small gain value when the state trajectory is at immediate proximity of the desired values. The designed structure has the advantage to derive the time varying gain to the fixed minimal value after disturbances cancellation. Otherwise, the incessant switching gain increasing may lead for chattering occurrence or estimator divergence.

4. SM-MRAS ROTOR SPEED ESTIMATOR DESIGN

The MRAS algorithm for speed estimation is based on the fact that there are two ways to estimate the rotor fluxes from the basic equations of the SPIM in the stationary reference frame.

From the stator equations

$$\emptyset_{r\alpha}^{.} = \frac{L_r}{M_{srd}}[v_{s\alpha} - (R_{sd} + \rho L_{sd}\sigma_d)i_{s\alpha}] \qquad (11)$$

$$\emptyset_{r\beta}^{.} = \frac{L_r}{M_{srq}}[v_{s\beta} - (R_{sq} + \rho L_{sq}\sigma_q)i_{s\beta}] \qquad (12)$$

From the rotor equations

$$\emptyset_{r\alpha}^{*} = -\frac{1}{\tau_r}\emptyset_{r\alpha}^{\wedge} - \omega_r^{\wedge}\emptyset_{r\beta} + \frac{M_{srd}}{\tau_r}i_{s\alpha} \qquad (13)$$

$$\emptyset_{r\beta}^{*} = -\frac{1}{\tau_r}\emptyset_{r\beta}^{\wedge} + \omega_r^{\wedge}\emptyset_{r\alpha} + \frac{M_{srq}}{\tau_r}i_{s\beta} \qquad (14)$$

Notice that the stator equations are independent of the rotor speed and are used as the reference model. The rotor equations are dependent on the rotor speed and thus can be used as the adjustable model.
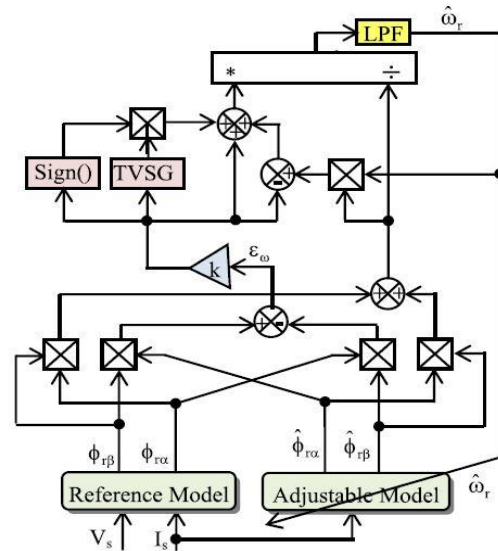

Fig 4 Speed estimator block diagram

The main idea of the MRAS is to compare the outputs of the two models and to adjust the value of rotor speed in order to minimize the resultant error. The adjusted value is the speed generated from the error between rotor fluxes observed. Fig. 4 gives the structure of the proposed SM-MRAS speed estimation. Where, the tuning signal $\in_\omega$ is the input of the designed sliding mode adaptation mechanism. This will delivers the value of the estimated rotor speed used to actualize the adjustable mode.

4.1 Design of SM-MRAS Algorithm

A classical speed tuning signal $\in_\omega$ is adopted and is expressed by the following equation:

$$\in_\omega = \emptyset_{r\alpha}^{\wedge}\emptyset_{r\beta} - \emptyset_{r\alpha}\emptyset_{r\beta}^{\wedge} \qquad (15)$$

Defining a sliding surface S proportional to the previous speed tuning signal as

$$S = k \in_\omega \qquad with\ k>0 \qquad (16)$$

When the desired state spaces reach the switching surface and stay there, this gives

$$S = S^\bullet = 0 \qquad (17)$$

Therefore, the error dynamic can be described by the following equation, meaning that it will be forced to exponentially tend to zero:

$$\dot{\epsilon}_\omega = -k\ \epsilon_\omega \qquad (18)$$

Differentiating **(18)** yields

$$\dot{\epsilon}_\omega = \widehat{\phi_{r\alpha}}\dot{\phi}_{r\beta} + \dot{\widehat{\phi_{r\alpha}}}\phi_{r\beta} - \widehat{\phi_{r\beta}}\dot{\phi}_{r\alpha} - \phi_{r\beta}\dot{\widehat{\phi_{r\beta}}}$$

Substituting the derivative estimated terms by their expressions, this equation can be arranged as

$$\dot{\epsilon}_\omega = k_1 - k_2\widehat{\omega_r} \qquad (19)$$

With

$$k_1 = \dot{\phi}_{r\beta}\widehat{\phi_{r\alpha}} - \dot{\phi}_{r\alpha}\widehat{\phi_{r\beta}} + \frac{1}{\tau_r}\left[(\widehat{\phi_{r\beta}} - M_{srq}i_{s\beta})\phi_{r\alpha} - (\widehat{\phi_{r\alpha}} - M_{srd}i_{s\alpha})\phi_{r\beta}\right] \qquad (20)$$

$$k_2 = \phi_{r\alpha}\phi_{r\alpha}^{\wedge} + \phi_{r\beta}\phi_{r\beta}^{\wedge} \qquad (21)$$

The rotor speed can be estimated using the following proposed adaptation control law made up equivalent term and nonlinear term:

$$\widehat{\omega_r} = U_{eq} + U_d \qquad (22)$$

Equalizing **(18)** and **(19)**, the equivalent term can be carried out and expressing the rotor estimated speed value when the state variable holds on the sliding surface

$$U_{eq} = \frac{k_1 + k\epsilon_\omega}{k_2} \qquad (23)$$

$$U_d = \frac{\Delta_1}{k_2}sign(s) \qquad (24)$$

4.2 Stability proof

Defining the following candidate Lyapunov function V as:

$$V = \frac{1}{2}S^2 \qquad (25)$$

The stability of the proposed sliding mode control law describing the adaptation mechanism can be easily established. Then, the first time derivative expression of the candidate Lyapunov function is:

$$V = \dot{S}(k\ \dot{\epsilon}_\omega) \qquad (26)$$

Taking into account expressions (19) and substituting the equivalent and nonlinear law terms by their expressions, we get:

$$\dot{V} = -k^2\ S^2 - k\Delta_1|S| \leq 0 \qquad (27)$$

Since k and G1 are positives then, the derivative form is always negative, therefore the proposed control law is stable and the switching function S is for all time attractive and is reached in finite time.

4.3 Estimator Chattering Reduction

The overall rotor speed estimator block diagram with a TVSG as designed is given by Fig. 4. The adaptive switching gain principle used in the above paragraph is the same used in the design of the speed estimator adaptive switching gain. For proper rotor speed estimation, the generated signal is followed by a discrete LPF. The use of an LPF is a recognized practical method to get a continuous signal from the generated signal.

The designed LPF recurrent expression is

$$u(k+1) = \zeta u(k+1) + (1-\zeta)\ u(k) \qquad (28)$$

Where ζ is a tuning positive constant filter parameter satisfying the following condition:

$$\zeta = \frac{T_s}{T_s + T_c} \qquad (29)$$

Where T_s is the sampling time and T_c is the considered time constant of the LPF. The speed is estimated very accurately by making T_c as small as possible, but larger than the sampling time. Then, the suppression chattering objective can be completely accomplished.
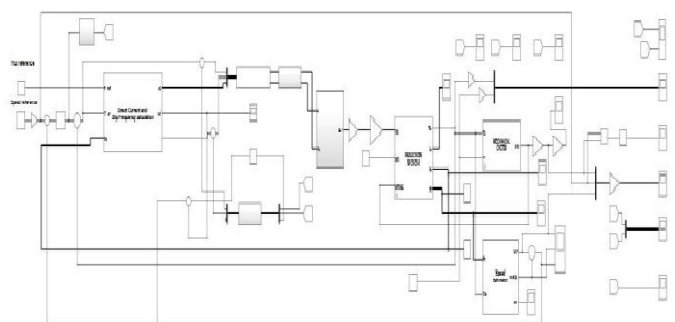
5. Simulation model and Results

Fig 5.1 The simulation block diagram

The simulation block diagram consists of mainly induction motor modeling block, voltage source inverter block and sliding mode model reference adaptive system block. The output of the inverter is given to the motor and when it receives below the rate speed then inverter currents $i_{s\alpha}$, $i_{s\beta}$ and motor voltages $v_{s\alpha}$, $v_{s\beta}$ are applied to sm-mras scheme. From this speed is estimated and it is compared with step reference speed and error is calculated. From that control signal is obtained which is given to the voltage source inverter. Based on control signal voltage source inverter adjust the firing pulses to obtained the reference speed.
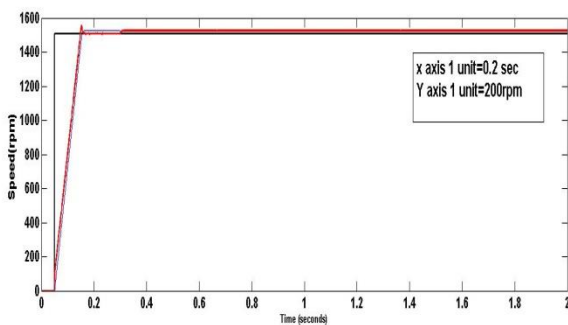
Fig. 5.2 rotor speed (rpm)
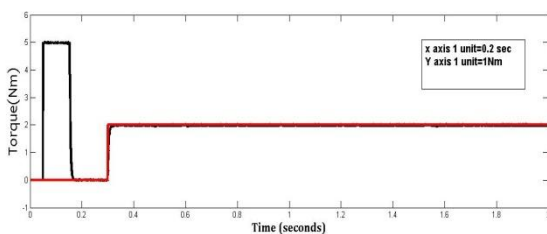
Fig. 5.3 Electromagnetic torque (Nm)
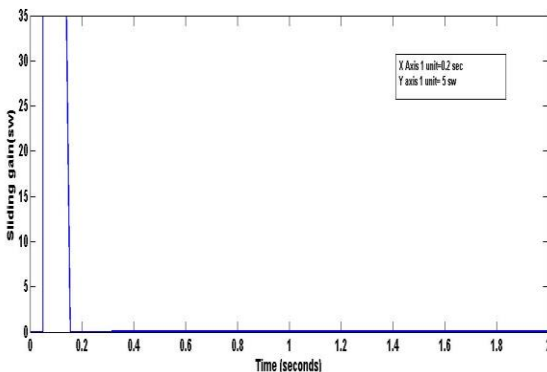
Fig. 5.4 Speed sliding surface (Sw)
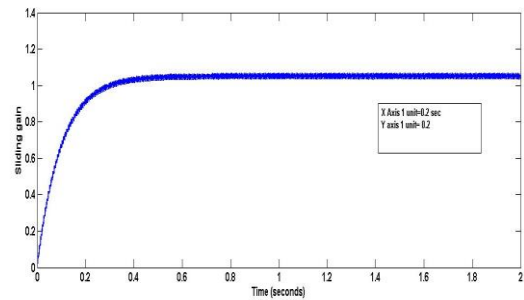
Fig. 5.5 Sliding gain
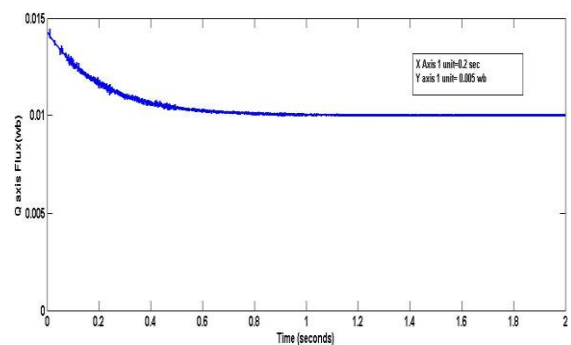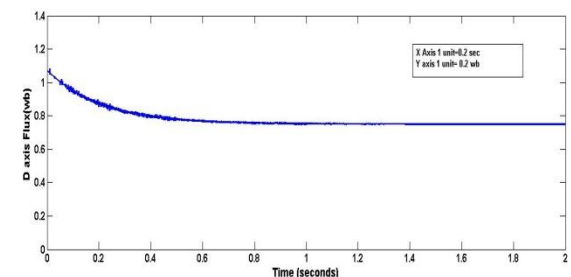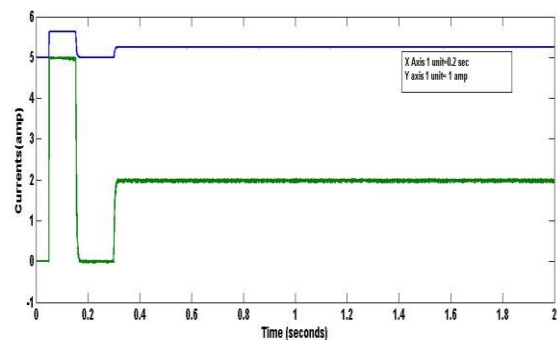

Fig. 5.6 Direct and quadratic flux (wb)

Fig. 5.7 Stator currents (A)

The SPIM is supplied by three leg voltage source inverter, throughout six IGBT transistors. The retained connection topology of the main and auxiliary windings to the three-leg inverter is that: 1) two legs are dedicated to the control of the main and auxiliary winding voltages and 2) the third one is used to control the offset voltage. The load torque is applied via a magnetic powder brake

coupled to the SPIM and piloted with a control unit. An isolation amplifier block is calibrated and then used to measure the main and auxiliary windings currents and the SPIM speed.

The simulation is achieved using MATLAB/SIMULINK software. It operates at a frequency of 250 MHz. To generate the PWM signals, we adopted a sinusoidal PWM technique with a switching frequency of 3 kHz and a dead time of 2 μs for the three-leg inverter. The sampling time of the control algorithm is fixed to 60 μs.

The SPIM started without load torque, and then a nominal load torque is applied at 6 s and removed at 16 s. In the proposed sysytem SPIM is operating at nominal speed. The proposed SOSM speed controller and the speed SM-MRAS estimator exhibit good tracking performances and fast response without steady error and with neglected overshoot (<3%).

6. CONCLUSIONS

An adaptive sliding-mode observer for speed sensorless FOC of SPIM drive has been presented. A robust SOSM speed controller, two SOSM controllers to drive the stator currents, and a SM-MRAS adaptation mechanism for SPIM speed estimation has been designed. In addition to its chattering elimination capability, the use of an adaptive STA in control law design guaranties fast responses and a short time convergence. Owing to a very simple online self adaptive switching gain, the proposed SOSM control law does not require the knowledge of uncertainties bounders, to guarantee the desired performances. The main advantage of the proposed SM-MRAS speed estimator is that it does not necessitate a sublimate algorithm for parameter estimation to have accurate speed estimation even at very low speed range. The SM-MRAS algorithm stability was proved by Lyapunov approach. Finally, and by means of simulation results, it has been shown that the designed sensorless speed SPIM drive, exhibits a very good tracking trajectory error and high performance speed estimation.

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