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# A New Approach Fuzzy-MRAS Speed Sensorless Sliding mode Control for Interior Permanent-Magnet Machine Drive

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***Abstract*: This paper presents a speed sensorless sliding mode control strategy of interior permanent magnet synchronous motor (IPMSM). The estimation of the rotor position and speed is achieved by using a model reference adaptive system (MRAS) to alleviate the need of physical sensors. Sliding mode control is able to increase the performance of the Fuzzy MRAS observer in terms of, speed reversion and load rejection. Simulation results with IPMSM shows that the designed sensorless sliding mode control based Fuzzy MRAS observer réalises a good dynamic behaviour of the motor with a rapid settling time, no overshoot and has better performance than sensorless vector control PI MRAS observer.**

***Keywords***: Sensorless control, Fuzzy MRAS, interior permanent magnet synchronous motor (IPMSM), sliding-mode control, vector control

## I. Introduction

Permanent Magnet Synchronous Motors (PMSM) are widely used in industrial drives due to their high power density, high torque-to- inertia ratio, small torque ripple and precise control at low speed range, possibility to torque control at zero speed, high efficiency and small size [1].

To exploit presented advantages, a sliding mode control should be used. The main advantage of the sliding mode technique is that the controlled system presents robustness to parameters variations and rejection to external disturbances. The sliding mode control of a permanent magnet synchronous motor is usually implemented through measuring the speed and position [2].

In order to obtain precise signals of the rotor position and speed, many detection devices were developed in recent years, such as Hall Effect sensor, photoelectric encoder and resolver etc [3].

Due to mechanical sensor increased the system cost, and the reliability of the system has many limitations, such as temperature, humidity and vibration, non-speed sensor becomes hot in recent years [4].

Several approaches to this problem are reported in the literature.

In this paper, a novel nonlinear adaptation scheme based on fuzzy logic is proposed to improve the speed estimation performance. The proposed speed controller is evaluated by simulation using MATLAB/SIMULINK under different operation point.

## II. Mathematic Model of IPMSM

Let we develop the state space model of the IPMSM in a synchronous reference frame. Fig. 1 shows a general purpose of three-phase inverter fed IPMSM drive.

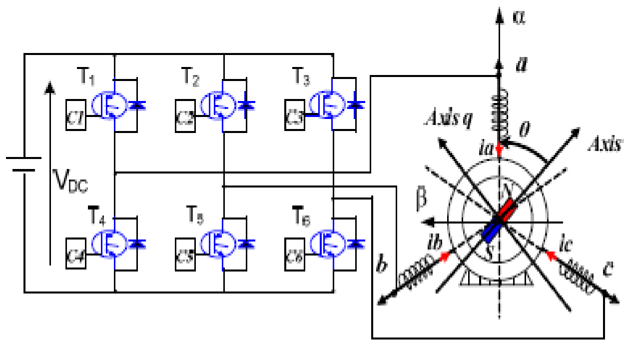



**Figure 1.** Voltage source inverter fed IPMSM drive

A simplified schematic three phase IPMSM drive fed by voltage source inverter (VSI) controlled by SVPWM technique. The power circuit of a three-phase VSI is shown in Fig. 1, where $v_a$, $v_b$ and $v_c$ are the output voltages applied to the star-connected IPMSM windings and $V_{DC}$ is the $DC$ voltage input inverter [6].

The d-q model in the rotating reference frame is used to analyze the IPMSM and the proposed method of speed and position estimation. The stator voltage and flux equations of the IPMSM in the rotating d-q reference frame are given by:

$$\begin{cases} v_d = R_s i_d + L_d \dfrac{di_d}{dt} - \omega_r L_q i_q \\ v_q = R_s i_q + L_q \dfrac{di_q}{d_t} + \omega_r L_d i_d + \omega \psi_r \end{cases} \quad (1)$$

$$\begin{cases} \psi_d = L_d i_d + \psi_r \\ \psi_q = L_q i_q \end{cases} \quad (2)$$

where $v_d, v_q, i_d, i_q$ are the stator voltages and currents, respectively, $R_s$ is the stator resistance, $L_d, L_q$ are the d-q axis stator inductances, respectively, $\psi_d$ and $\psi_q$ are the d-q axis stator magnetic flux, respectively, $\omega$ is the electrical rotor speed, $\psi_r$ is the rotor flux, $\Omega_r$ is the mechanical rotor speed, as $\omega_r = P.\Omega_r$.

The electro-mechanical torque is

$$T_e = P\left[\psi_r i_q + (L_d - L_q) i_d i_q\right] \quad (3)$$

The mechanical equation of the PMSM drive is described as:

$$J \frac{d\omega_r}{dt} = P\left(T_e - T_L - \beta \frac{\omega_r}{P}\right) \quad (4)$$

Where $J$ and $\beta$ are the inertia and the friction coefficient of the rotor speed and position. The primary methods the motor, respectively, $T_e$ and $T_L$ are the electromagnetic torque and the load torque, respectively, $P$ is the pole pairs [7].

## III. SVPWM Based Three Phase Inverter

The principle of SVM is based on the fact that there are only eight possible switching combinations for a three phase inverter to calculate inverter transistor on times. The transition from one switching state to the next involves only two switches in the same inverter leg, when one being switched ON, the other being switched OFF, in order to minimize the device switching loss [8]. Six non-zero vectors ($V_1$ to $V_6$), called the active vectors, shape the axis of hexagonal and the angle between any adjacent two non-zero vectors is 60 °[9].The tips of these vectors form a regular hexagon. Two of these states ($V_0$ and $V_7$) correspond to a short circuit on the output called the zero vectors, while the other six is considered to form stationary vectors in the $\alpha - \beta$ complex plane [10] as shown in Fig. 2. The reference voltage is generated by two adjacent non-zero vectors and two zero vectors. In linear operation range on times is calculated as follows,

$$T_1 = \frac{V_{s\_ref}}{V_{DC}} \cdot \frac{\sin(\Pi/3 - \theta)}{\sin(\Pi/6)} . T_s \quad (5)$$

$$T_1 = \frac{V_{s\_ref}}{V_{DC}} \cdot \frac{\sin\theta}{\sin(\Pi/6)} . T_s \quad (6)$$

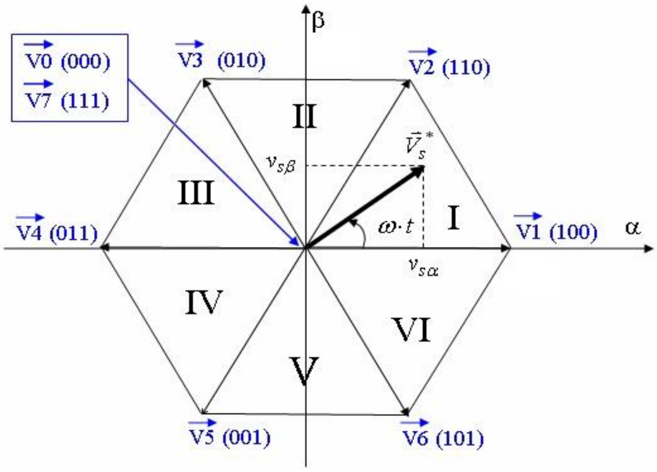



**Figure 2.** Space vector hexagon

## IV. Direct Field Oriented Control

Similarly to the IM, in the IPMSM a decoupled control of the torque and flux magnitudes can be achieved, emulating a DC motor, by means of the FOC strategy [11]. This is done using the d-q transformation that separates the components d and q of the stator current responsible for flux and torque production respectively. Due to the presence of the constant flux of the permanent magnet, there is no need to generate flux by means of the $i_{ds}$ current, and this current can be kept to a zero value, which in turns decreases the stator current and increases the efficiency of the drive. The control scheme of the FOC strategy is shown in Fig. 3.

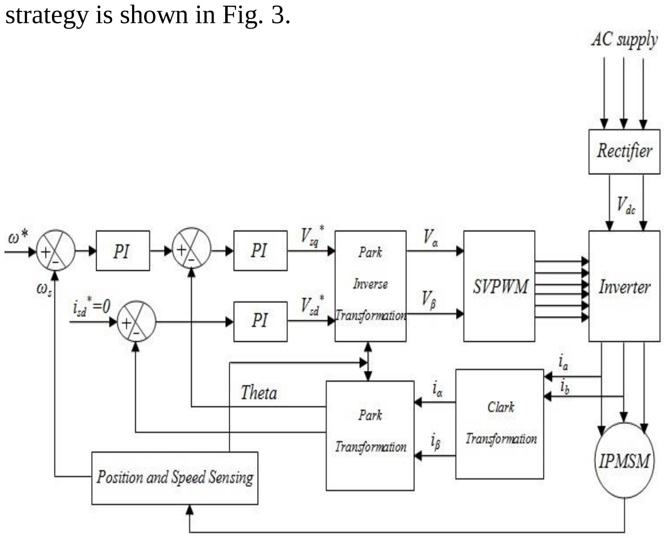



**Figure 3.** IPMSM Vector Control Algorithm Overview

The control system is divided into three different loops: the d loop, which controls the flux; and the q loops, which control the speed and torque. The d loop performs the control of $i_{sd}$ with a current PI regulator. The reference value for this loop can be set to 0. The q loops are connected in cascade. The inner loop controls the torque by means of controlling $i_{sq}$ with a current PI regulator. The fact that the torque can be controlled by means of $i_{sq}$ comes from the following simplification of (3), valid for Surface Mounted (SM) IPMSM:

The reference for this inner loop is given by the speed PI regulator of the outer loop. From the voltage equations of the IPMSM model (1) it can be seen that d and q axis are not completely independent and there are coupling terms which depend on the current from the other axis. To achieve completely independent regulation it is necessary to cancel the effect of these coupling terms at the output of the current PI regulator (see Fig. 3). The use of decoupling achieves the linearization of the control system as well as higher dynamics.

## V. Sliding Mode Control

### *A. Design of Sliding Mode Controller*

The sliding mode control can be justified and designed using the notion of lyapunov stability. Whatever the application, the design of sliding mode control can summarized in 3 steps [2]:

*1) The choice of the number of the sliding surfaces:*
Generally the number of the sliding surfaces is equal to the dimension of the input control vector.

*2) The choice of the sliding surface equation form:*
It must satisfy the convergence of the control and the stability of the system. This goal can be reached if the control variable $u_c$ permits to satisfy the Lyapunov function:

$$\dot{s}s\langle 0 \quad (7)$$

Based on this condition Slotine [2] propose a general form of sliding surface:

$$s(x)=\left(\frac{\partial}{dt}+\lambda_x\right)^{n-1}\left(x_{ref}-x\right) \quad (8)$$

*3) The control law design:*
The control variable is decomposed in two parts: $u_{equ}$ and $u_n$ :

$$u_c = u_{equ} + u_n \quad (9)$$

The dynamic while in sliding mode can be written as

$$\dot{s}(x) = 0 \quad (10)$$

By solving this equation ( $\dot{s} = 0$ ), the equivalent control $u_{equ}$ can be obtained. The $u_n$ component satisfies $\dot{s}.s < 0$ and is given by:

$$u_n = -k.sign(s(x)) \quad (11)$$

Where $k$ is the control gain.

The controller described by the equation (11) presents high robustness, insensitive to parameter fluctuations and disturbances [12], but it will have high-frequency switching (chattering phenomena) near the sliding surface due to sign function involved. These drastic changes of input can be avoided by introducing a boundary layer with width ε [13]. Thus replacing $sign(s(x))$ by in (11), we have

$$u_c = u_{equ} - k.sat(s(x)) \quad (12)$$

Where $\varepsilon\rangle 0$, $\varphi = s(x)/\varepsilon$

$$sat(\varphi)=\begin{cases} sign(\varphi) & if\ |\varphi|\geq 1 \\ \varphi & if\ |\varphi|\langle 1 \end{cases} \quad (13)$$

### *B. Speed control*

The speed error is defined by [14]:

$$e_\Omega = \Omega_{rref} - \Omega_r \quad (14)$$

For $n = 1$, the position control manifold equation can be obtained from equation (8) as follow:

$$s(\Omega_r) = \Omega_{rref} - \Omega_r \quad (15)$$

The derivative of this surface is given by the expression:

$$\dot{s}(\Omega) = -c_1\Omega_r + \frac{T_L}{J} + \dot{\Omega}_{rref} - (c_2 i_d + c_3) i_q \quad (16)$$

During the sliding mode and in permanent regime, we have

$$s(\Omega_r) = 0, \dot{s}(\Omega_r) = 0, i_{qn} = 0 \quad (17)$$

The current control $i_q$ is defined by:

$$i_q = i_{qequ} - i_{qn} \quad (18)$$

The control voltage $i_{qref}$ is defined by:

$$i_{qref} = \frac{-c_1\Omega_r + \frac{T_L}{j} + \Omega_{rref} + k_\Omega \Omega_r sign(s(\Omega_r))}{c_2 i_d + c_3} \quad (19)$$

### *C. Direct current controller*

The direct current error is defined by:

$$e_d = i_{dref} - i_d \quad (20)$$

For $n = 1$, the direct current control manifold equation can be obtained from equation (8) as follow:

$$s(i_d) = i_{dref} - i_d \quad (21)$$

Substituting the expression of $i_d$ given by equation (1) and (2) in equation (20) we obtain:

$$\dot{s}(i_d) = \frac{d}{dt} i_{dref} - a_1 i_d - a_2 i_q \Omega_r - \frac{1}{L_d} v_d \quad (22)$$

During the sliding mode and in permanent regime, we have

$$s(i_d) = 0, \dot{s}(i_d) = 0, v_{dn} = 0 \quad (23)$$

The control voltage $v_{dref}$ is defined by:

$$v_{dref} = \frac{\left[\frac{d}{dt} i_{dref} - a_1 i_d - a_2 i_q \Omega_r\right] + k_d sign(s(i_d))}{L_d} \quad (24)$$

### *D. Quadrature current control*

The quadrature current error is defined by:

$$e(i_q) = i_{qref} - i_q \quad (25)$$

For $n = 1$, the quadrature current control manifold equation can be obtained from equation (8) as follow:

$$s(i_q) = i_{qref} - i_q \quad (26)$$

Substituting the expression of $i_q$ given by equation (1) and (2) in equation (25) we obtain:

$$\dot{s}(i_q) = \frac{d}{dt} i_{qref} - b_1 i_q - b_2 i_d \Omega_r - b_3 \Omega_r - \frac{1}{L_q} v_q \quad (27)$$

During the sliding mode and in permanent regime, we have

$$s(i_q) = 0, \dot{s}(i_q) = 0, v_{qn} = 0 \quad (28)$$

The control voltage $v_{qref}$ is defined by:

$$v_{qref} = \frac{\left[\frac{d}{dt} i_{qref} - b_1 i_q - b_2 i_d \Omega_r + b_3 \Omega_r + k_q sign(s(i_q))\right]}{L_d} \quad (29)$$

With:

$$a_1 = \frac{-R_s}{L_d}, a_2 = \frac{PL_q}{L_d}, b_1 = \frac{-R_s}{L_q}, b_2 = \frac{-PL_d}{L_q},$$

$$b_3 = \frac{-P\phi_f}{L_q}, c_1 = \frac{-B}{J}, c_2 = \frac{P(L_d - L_q)}{J}, c_3 = \frac{P\phi_f}{J}$$

The suitable choice of the parameters $k_d$ $k_q$ and $k_\Omega$ [2]:

- Ensures the rapidity of the reaching mode.
- Imposes the dynamic of the convergence and sliding mode.
- Allow to the drive to work with maximum energy during transient state.

## VI. Fuzzy Logic MRAS Speed Observer

As the rotor speed $\Omega_r$ is included in current equations, we can choose the current model of the IPMSM as the adjustable model, and the motor itself as the reference model. These two models both have the output $i_d$ and $i_q$, According to the difference between the outputs of the two models, through a certain adaptive mechanism, we can get the estimated value of the rotor speed. Then the position can be obtained by integrating the speed [15].

The equation (1) can be written as below form:

$$\begin{cases} L_d \dfrac{di_d}{dt} = -R_s i_d + \omega_r L_q i_q + v_d \\ L_q \dfrac{di_q}{dt} = -R_s i_q - \omega_r L_d i_d - \omega \psi_r + v_q \end{cases} \quad (30)$$

The d-q axis currents $i_d$, $i_q$ are the state variables of the current model of the IPMSM, which is described by (30). Rewrite (30) into matrix form as below

$$\frac{d}{dt}\begin{bmatrix} i_d + \dfrac{\psi_r}{L_d} \\ i_q \end{bmatrix} = \begin{bmatrix} -\dfrac{R_s}{L_d} & \omega_r \dfrac{L_q}{L_d} \\ -\omega_r \dfrac{L_d}{L_q} & -\dfrac{R_s}{L_q} \end{bmatrix} \begin{bmatrix} i_d + \dfrac{\psi_r}{L_d} \\ i_q \end{bmatrix} + \begin{bmatrix} \dfrac{v_d}{L_d} + \dfrac{R_s \psi_r}{L_d} \\ \dfrac{v_q}{L_q} \end{bmatrix} \quad (31)$$

For the convenience of stability analysis, the speed $\Omega_r$ has been confined to the system matrix

$$A = \begin{bmatrix} -\dfrac{R_s}{L_d} & \omega_r \dfrac{L_q}{L_d} \\ -\omega_r \dfrac{L_d}{L_q} & -\dfrac{R_s}{L_q} \end{bmatrix} \quad (32)$$

To be simplified, define:

$$x = \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} = \begin{bmatrix} i_d + \dfrac{\psi_r}{L_d} \\ i_q \end{bmatrix} \quad (33)$$

$$u = \begin{bmatrix} u_1 \\ u_2 \end{bmatrix} = \begin{bmatrix} \dfrac{v_d}{L_d} + \dfrac{R_s \psi_r}{L_d} \\ \dfrac{v_q}{L_q} \end{bmatrix} \quad (34)$$

Then the reference model can be rewritten as:

$$\frac{d}{dt} x = Ax + u \quad (35)$$

The adaptation mechanism uses the rotor speed as corrective information to obtain the adjustable parameter current error between two models in order to drive the current error to zero, when we can take the estimation value as a correct speed. The process of speed estimation can be described as follows:

$$\frac{d}{dt}\begin{bmatrix} \hat{x}_1 \\ \hat{x}_2 \end{bmatrix} = \begin{bmatrix} -\dfrac{R_s}{L_d} & \hat{\omega}_r \dfrac{L_q}{L_d} \\ -\hat{\omega}_r \dfrac{L_d}{L_q} & -\dfrac{R_s}{L_q} \end{bmatrix} \begin{bmatrix} \hat{x}_1 \\ \hat{x}_2 \end{bmatrix} + \begin{bmatrix} \hat{u}_1 \\ \hat{u}_2 \end{bmatrix} \quad (36)$$

Where $\hat{\omega}$ is to be estimated, (36) can be simplified as below:

$$\frac{d}{dt} \hat{x} = \hat{A}\hat{x} + \hat{u} \quad (37)$$

The error of the state variables is

$$e = x - \hat{x} \quad (38)$$

According to equation (35) and (36), estimation equation can be written as:

$$\begin{cases} \dfrac{d}{dt} e = Ae - Iw \\ v = De \end{cases} \quad (39)$$

Where $w = (\hat{A} - A)\hat{x}$, choose $D = I$, the $v = Ie = e$

According to Popov Super Stability Theorem, if the following conditions are met [16]:

(1) $H(s) = D(sI - A)^{-1}$ is a strictly positive matrix,

(2) $\eta(0, t_0) = \int_0^{t_0} v^T w dt \geq -\gamma_0^2, \forall t_0 \geq 0$, where $\gamma_0^2$ is a limited positive number, the $\lim_{t \to \infty} e(t) = 0$, in other words, the model reference adaptive system is asymptotically stable.

Finally, the error between the two models is:

$$e = (x_1 \hat{x}_2 - x_2 \hat{x}_1) \quad (40)$$

Replacing $x$ with $i$ :

$$e=\left[i_d\hat{i}_q-i_q\hat{i}_d-\frac{\psi_r}{L_d}(i_q-\hat{i}_q)\right] \quad (41)$$

The rotor position can be obtained by integrating the estimated speed:

$$\hat{\theta}_r=\int_0^t \hat{\omega}_r d\tau \quad (42)$$

The error between two models is the input of a fuzzy logic controller whose output is the estimated rotor speed; this estimated speed is used to adjust the adaptive model [6].

Fuzzy control provides a formal methodology for representing, manipulating, and implementing a human's heuristic knowledge about how to control a system and for constructing nonlinear controllers via the use of heuristic information. This kind of fuzzy controller is also known as "PI fuzzy controller". The plant control u is inferred from the two state variables, error (E) and change in error Δe. The adequate choice of Gεe and Gεce, Gεu permits to adapt the estimation.

Fuzzy control uses a set of rules to represent how to control the plant. Table I shows one of possible control rule base. The rows represent the rate of the error change DE and the Columns represent the error E. Each pair (E, DE) determines the output level NB to PB corresponding to U.

Here NB is negative big, NM is negative medium, ZR is zero, PM is positive medium and PB is positive big, are labels of fuzzy set sand their corresponding membership functions are depicted in Fig. 5, 6 and 7, respectively.

The continuity of input membership functions, reasoning method, and defuzzification method for the continuity of the mapping U (E, DE) is necessary. In this paper, the Triangular membership function, the max-min reasoning method, and the center of gravity defuzzification method are used, as those methods are most frequently used in many literatures [1].

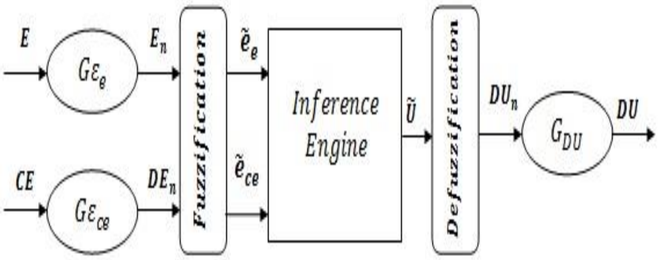



**Figure 4.** Fuzzy logic controller for adaptation mechanism

*Table 1.* Rule base for fuzzy controller.

| *DU* | | *DE<sub></sub>n* | | | | |
|---|---|---|---|---|---|---|
| | | NB | NM | ZR | PM | PB |
| $E_n$ | NB | NB | NB | NM | NM | ZR |
| | NM | NB | NM | NM | ZR | PM |
| | ZR | NM | NM | ZR | PM | PM |
| | PM | NM | ZR | PM | PM | PB |
| | PB | ZR | PM | PM | PB | PB |

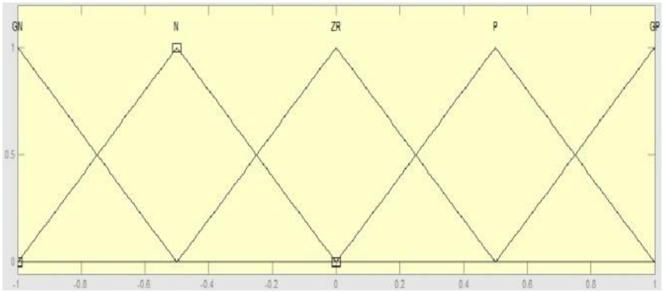



**Figure 5.** Membership Functions for Input E

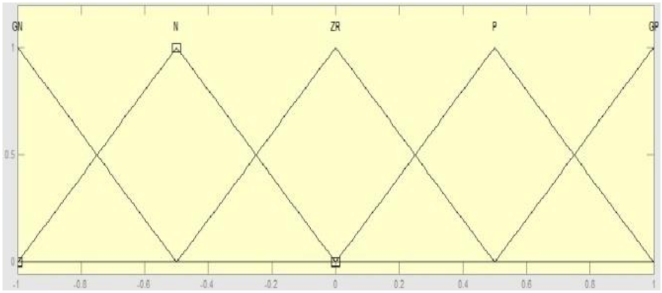



**Figure 6.** Membership Functions for Input DE

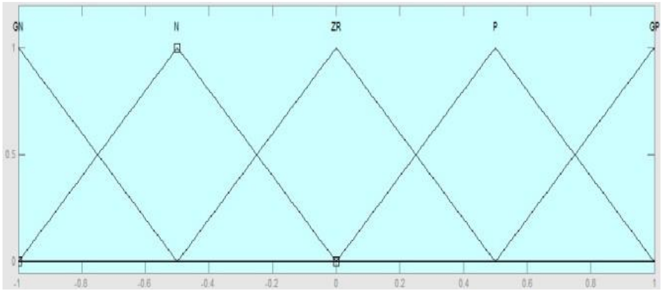



**Figure 7.** Membership Functions for Output DU

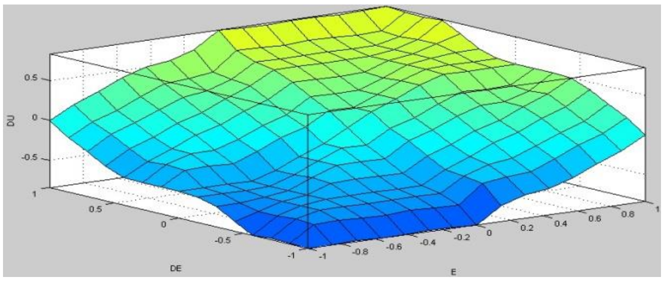



**Figure 8.** Control surface of fuzzy controller

The overall MRAS speed observer with FL speed estimation mechanism (MRAS-FL) is shown in Fig. 9. The inputs of the speed and position estimation block are the voltages and currents of the real motor, and the outputs are the estimated speed and position.

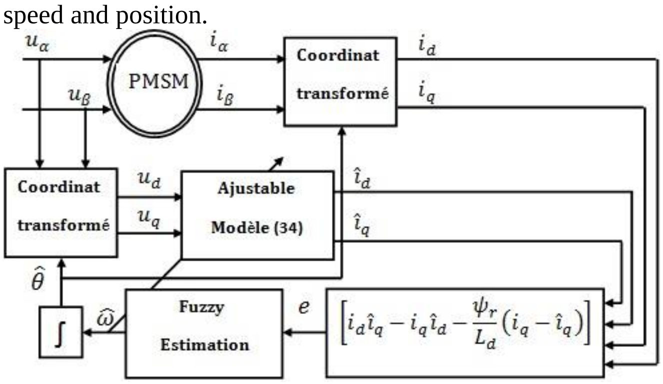



**Figure 9.** Control block scheme of MRAS

## VII. Simulation Results, Discussion And Comparison

In order to verify the proposed control strategies as discussed above, digital simulation studies were made the system described in Fig.10.

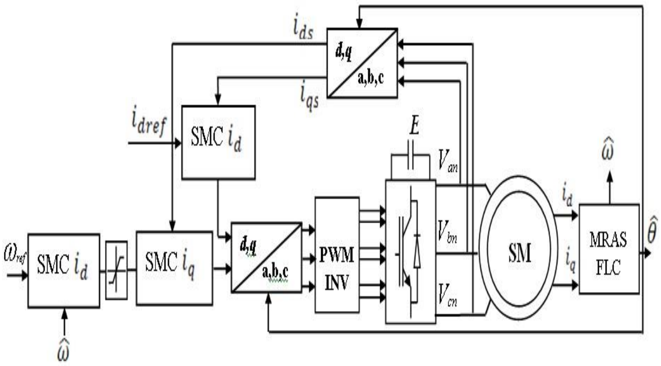
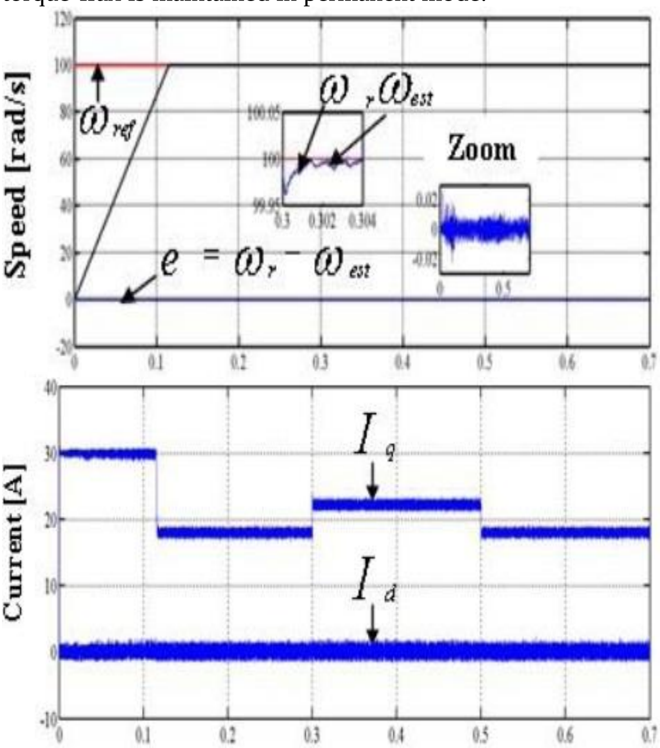



**Figure 10.** Block diagram SSMC of IPMSM based MRAS

The speed and currents loops of the drive were also designed and simulated respectively with sliding mode control.
The parameters of IPMSM used in the simulation are listed in Table II. The simulation is realized using the SIMULINK software in MATLAB environment.
Fig.11, shows the settling performance of the controller and its disturbance rejection capability.
Initially the machine is started up with a load of 8.5 N.m. At 0.3 s, a 2 N.m load is applied during a period of 0.2 s.
The transient speed, the speed error, the current and the torque responses also have a good dynamic performance when the load torques change. Especially, the speed of the motor and estimated speed quickly approach to the reference speed in case of two suddenly Load variation. The speed error is always kept at zero during the steady state. The decoupling of torque-flux is maintained in permanent mode.



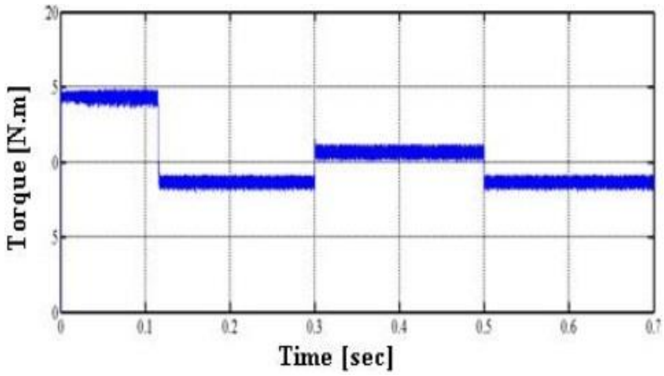




**Figure 11.** Controllers' Load Disturbance Rejection Performance

Fig.12 show the speed tracking performance of the control proposed under no load. The reference speed change from 100rpm to -100rpm at t=0.3sec. We remark that the estimated speed was in accord with the real speed and speed reference, thus, the sensorless sliding mode control using Fuzzy MRAS was found to perform well in high speed regions, including a forward–reverse rotation.
The Fig.13 shows the performance of MRAS technique in low speed. The fuzzy logic controller provides high control in this region.

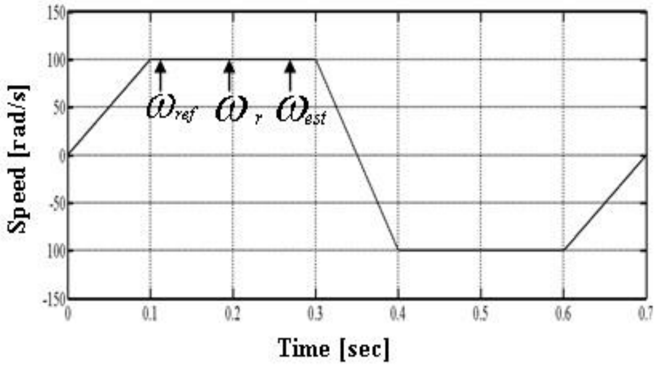



**Figure 12.** Controllers' Speed Tracking Performance in high speed

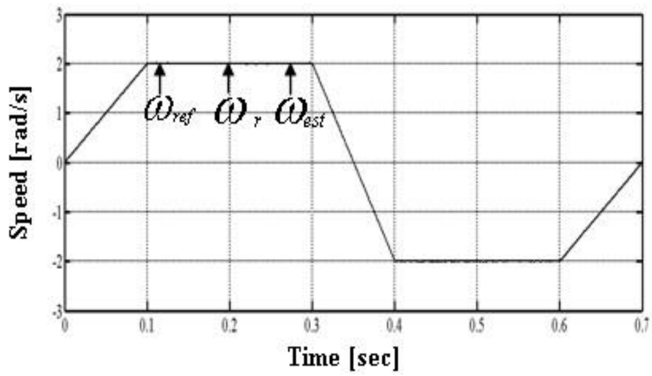



**Figure 13.** Controllers' Speed Tracking Performance in low speed

In order to test the robustness of the used method we have studied the effect of the parameters uncertainties on the performances of the speed control.
Three cases are considered:

- The moment of inertie (+50%) at 0.3sec.
- The stator resistances (+50%) at 0.3sec.

Fig.14 shows the reaction of the proposed control method to moment of inertia variation.

Fig.15 shows the reaction of the proposed control method to stator resistance variation.

For the robustness of control, an increase of the moment of inertia J, doesn't have any effects on the performances of the technique used (Fig.14 and 15), but in stator resistance variation, the rotor flux is deviated clockwise from the direct axis. Finally estimation of the stator resistance is required.

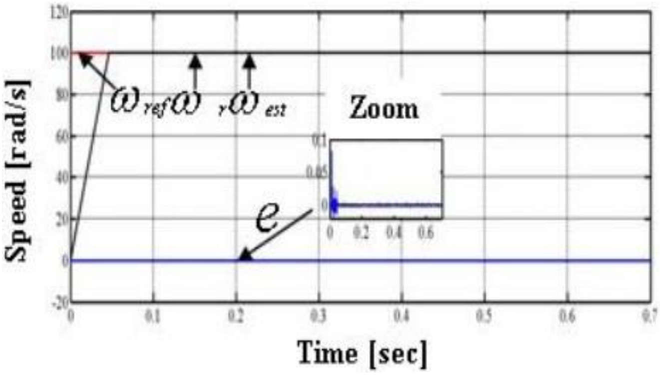



**Figure 14.** Test of robustness for moment of inertia variation

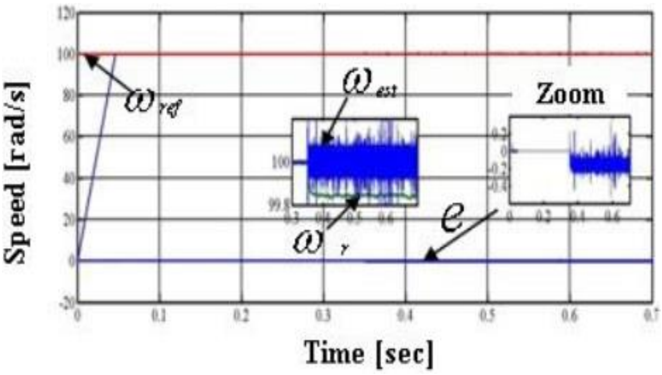



**Figure 15.** Test of robustness for stator resistance variation and zoom

## VIII. Conclusion

A fuzzy-MRAS based synchronous speed estimation method is proposed for the sensorless sliding mode controlled IPMSM drive. The proposed fuzzy-MRAS on-line synchronous speed estimation skill is based on stator current observer, in which the traditional PI adaptive mechanism is replaced by adaptive fuzzy ones in MRAS configuration.

The system was designed and the performance of this new approach was simulated and analyzed, such as dynamic/static and robust property. Simulation results shows that the designed sensorless sliding mode control using fuzzy-MRAS realizes a good dynamic behaviour of the motor with a rapid settling time, no overshoot and has better performance than sensorless vector control using MRAS observer.

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## Appendix

*Table 2.* Interior permanent magnet synchronous motor.

| ***Parameters*** | ***Specification*** | | |
|---|---|---|---|
| Rs | 0.18 | rated power 3Kw | |
| Ld | 0.0021H | J | 6.6*10-3Kg.m2 |
| Lq | 0.0042mH | B | 0.0014N.m/rad-1 |
| P | 4 | Te | 8.5N.m |
| ψr | 0.12wb | Rated Quadrature current | 20A |

## References

[1] N. Vasilios, Katsikis, *MATLAB – A Fundamental Tool for Scientific Computing and Engineering Applications*, Volume 1, September, 2012.

[2] F. Benchabane, A. Titaouine, O. Bennis, A. Guettaf, K. Yahia, D. Taibi, "Robust Position and Speed Estimation Algorithms for Permanent Magnet Synchronous Drives,'' European Journal of Scientific Research, vol.57, no.1, pp.6-14, 2011.

[3] W. Yan, H. Lin, H. Li, Hu. Li and J. Lu, "A MRAS based Speed Identification Scheme for a PM Synchronous Motor Drive Using the Sliding Mode Technique,'' *International Conference on Mechatronics and Automation*, IEEE, 2009.

[4] W. Wenjie, Z. Min, W. Qinghai, "Application of Reduced-Order extended Kalman Filter in permanent magnet synchronous motor sensorless regulating system'', *IEEE International Conference on Digital Manufacturing & Automation*, 2010.

[5] Y. Liang and Y. Li, "Sensorless control of PM synchronous motors *based on MRAS method and initial position estimation", In Proc. Int. Conf. Elect.* Mach. Syst., 2003, vol. 1, pp. 96–99.

[6] A. Khlaief, M. Bendjedia, M. Boussak, A. Chaâri, "Nonlinear Observer for Sensorless Speed Control of IPMSM Drive with Stator Resistance Adaptation'', *2nd International Conference on Communications, Computing and Control Applications (CCCA)*, pp 1-7, 6-8 Dec 2012.

[7] Xi Xiao, Yongdong Li, "A sensorless control based on MRAS method in interior permanent-magnet machine drive," *Proceedings of the International Conference on Power Electronics and Drive Systems*, vol.1, pp.734-736, 2005.

[8] G. K. Nisha, S. Ushakumari and Z. V. Lakaparampil, "Harmonic Elimination of Space Vector Modulated Three Phase Inverter", In *Proc. IMECS*, pp. 1109-1115, Mar. 2012.

[9] H. W. Van der Broeck, H. C. Skudelny and G.V. Stanke," Analysis and realisation of a pulse width modulator based on voltage space vectors", IEEE Trans. Ind. Appl., vol. 24, pp. 142-150, Jan./Feb. 1988.

[10] D. G. Holmes and T. A. Lipo, *Pulse Width Modulation for Power Converters: Principles and Practice*, New Jersey: Wiley IEEE Press, 2003.

[11] Krishnan R. *Electronic Motor Drives: Modeling, Analysis and Control*, Upper Saddle River, New Jersey, USA: Prentice Hall, Feb.2001.

[12] V. I. Utkin, "Variable structure system with sliding modes,'' IEEE Trans. on Automatic Control, vol. AC-22, April 1977, 210–222.

[13] J. J. E. Slotine, W. Li, *Applied nonlinear control*, Prentice Hall, USA, 1998.

[14] A. G. Aissaoui, M. Abid, H. Abid, A. Tahour and A. Tilmatine, "Sliding Mode Application for Synchronous Motor Drive,'' Acta Elecrotehnica, Vol. 47, no. 3, 2006.
[15] Yuchao Shi, Kai Sun, Hongyan Ma, Lipei Huang, " Permanent Magnet Flux Identification of IPMSM based on EKF with Speed Sensorless Control'', *-36th Annual Conference on IEEE Industrial Electronics Society*, pp. 2252- 2257, 7-10 Nov 2010.
[16] W. Huau, Z. Y. Chen, W. P. Cao, "Simulation research on optimization of permanent magnet synchronous motor sensorless vector control based on MRAS". *IEEE 2012.*

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