

AN IMPROVED SLIDING MODE OBSERVER FOR SPEED SENSORLESS DIRECT TORQUE CONTROL OF PMSM DRIVE WITH A THREE-LEVEL NPC INVERTER BASED SPEED AND STATOR RESISTANCE ESTIMATOR.

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Abstract: This paper presents a sensorless Direct Torque Control (DTC) methods for permanent magnet synchronous motors (PMSM) supplied by three-level Neutral-Point-Clamped (NPC) inverter based on the sliding mode observe (SMO). The stability is verified by Lyapunov theory. The SMO is utilized to compensate the effects of parameter variation on the stator resistance, which makes flux and torque estimation more accurate and insensitive to parameter variation.

In classical DTC, due to the hysteresis based scheme, the switching frequency is variable, current distortion and torque ripple are more important. In order to obtain a constant switching frequency and hence torque ripple minimization, we introduce a new control technique for PMSM using a three-level NPC inverter.

Simulation results confirm the effectiveness of the proposed method.

Key words: PSPM, DTC, three-level NPC inverter, PMW, SMO.

1. Introduction.

Permanent magnet synchronous motors (PMSM) have been widely used as servo-machines over the last two decades. In recent years, they are used more in the variable speed applications due to some advantages like: more simplicity, low dependency on the motor parameters, good dynamic torque response, high rate torque/inertia [1], [2]. Direct Torque Control (DTC) was introduced in 1985 by Takahashi and Depenbrock especially for the asynchronous and synchronous machines [2], [3]. The main advantages of DTC are the simple control scheme, a very good torque dynamic response, as well as the fact that it does not need the rotor speed or position to realize the torque and flux control (for this reason DTC is considered a "sensorless" control strategy), moreover DTC is not sensitive to parameters variations (except stator resistor),[2].

In the classic DTC the employment of the hysteresis controllers to regulate the stator magnetic flux and torque is natural to have high torque ripples and variable switching frequency, which is varying with speed, load torque, selected hysteresis bands and difficulty to control torque and flux at very low speed [4]-[11]. To overcome the above drawbacks, some researchers have been trying to propose solution to solve these problems by substitute hysteresis control by fuzzy control [4], where torque and flux ripple is not solved. An effective modality for reducing the torque ripple without using a high sampling frequency is to calculate a proper reference voltage vector that can produce the desired torque and flux values, and then applied to the inverter using space vector modulation (SVM) [5]-[9]. This approach is known in the literature as DTC-SVM. Even though this control method provides fast torque response and small torque ripples, its PI-based torque-load angle and speed control scheme does not provide satisfactory control performance in the presence of parameter variations and disturbances under a wide range of driving conditions.

In [7] a modified DTC scheme that utilizes (SVM) was reported for IM with fixed switching frequency and low torque and flux linkage ripples was reported. This system requires two proportional-integral (PI) controllers properly tuned at the same time for the best performance. In [8],[9], SVM was introduced into sensorless DTC for PMSM and the control system was analyzed.

Other researchers use multilevel inverters [10], the resolution of the voltage vectors can be improved and hence, more smooth torque and flux responses.

In [11] it was recently proposed a novel DTC algorithm for three-phase induction motor which employs a three-level inverter showing the same

dynamic performance as those obtained with a two-level inverter with lower torque and flux ripples results as well as a lower harmonic content in the stator current.

In almost all the implementations, the rotor position angle is measured by a shaft position sensor such as an optical encoder or a resolver. However, the presence of this sensor (expensive and fragile and require special treatment of captured signals), causes several disadvantages from the standpoint of drive cost, encumbrance, reliability and noise problem [1],[12],[13].

This is the reason why, in the last few years, many authors proposed sensorless *DTC* control *PMSM* drives using different methods for the estimation of the rotor position. Two kinds of those approaches seem to be preferable, depending on the speed operating range required by the application: signal injection techniques and state observers. Signal injection techniques take advantage of the magnetic saliency of the machine, due to saturation or geometric construction, to detect the rotor position through the injection of proper test signals [8],[9]. These methods offer a proper solution both for standstill and low speed operation. And the methods of the state observer are preferred for medium/high speed operation. They require the measurement of the stator currents and the information of the stator input voltages to estimate the back *EMF* which the rotor position and speed are estimated from. But the problems of the state observer are sensitive to the variation of the motor parameters [8]. Sliding mode observers (*SMO*) provide an effective way of improving the robustness of the control system against parameter variation, load disturbance and measurement errors due to its simple structure and fast response [12],[14].

This paper presents a speed and position sensorless *DTC* of *PMSM* with a sliding mode control based speed and stator resistance estimator, so that it can overcome the problem of sensitivity in the face of motor parameter variation.

A three-level *NPC* inverter reduces the torque ripple of permanent magnets synchronous motor performance compared to that obtained with a two-level inverter.

2. PSPM model.

The stator and rotor flux equation of *PMSM* can be written in the reference frame of Park in the following form:

$$\begin{bmatrix} \phi_d \\ \phi_q \end{bmatrix} = \begin{bmatrix} L_d & 0 \\ 0 & L_q \end{bmatrix} \begin{bmatrix} I_{sd} \\ I_{sq} \end{bmatrix} + \begin{bmatrix} \phi_e \\ 0 \end{bmatrix} \quad (1)$$

While the equations of the stator voltages are written in this same reference frame in the following form:

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = r_s \begin{bmatrix} i_{ds} \\ i_{qs} \end{bmatrix} + \begin{bmatrix} L_d & 0 \\ 0 & L_q \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{ds} \\ i_{qs} \end{bmatrix} + p \Omega_r \begin{bmatrix} 0 & -L_q \\ L_d & 0 \end{bmatrix} \begin{bmatrix} i_{ds} \\ i_{qs} \end{bmatrix} + p \Omega_r \begin{bmatrix} \phi_e \\ 0 \end{bmatrix} \quad (2)$$

In addition the electromagnetic torque can be expressed:

$$T_e = \frac{3}{2} p \left((L_d - L_q) I_{sd} I_{sq} + \phi_e I_{sq} \right) \quad (3) \quad \text{The}$$

mechanical equation of the motor can be expressed as flows:

$$J \dot{\Omega} = T_e - T_l - f_r \Omega \quad (4)$$

3. Conventional DTC

The methods of direct torque control (*DTC*) as shown in figure 1 consist of directly controlling the turn off or turn on of the inverter switches on calculated values of stator flux and torque from relation (6). The changes of state of the switches are linked to the changes in electromagnetic state motor. They are no longer controlled based on voltage and frequency references given to the commutation control of a pulse width voltage modulation inverter [2],[3]. The reference frame related to the stator makes it possible to estimate flux and the torque, and the position of flux stator. The aim of the switches control is to give the vector representing the stator flux the direction determined by the reference value

$$\begin{cases} \phi_{s\alpha} = \int_0^t (v_{s\alpha} - r_s I_{s\alpha}) dt \\ \phi_{s\beta} = \int_0^t (v_{s\beta} - r_s I_{s\beta}) dt \end{cases} \quad (5)$$

The *DTC* is deduced based on the two approximations described by the formulas (6) and (7) [2]:

$$\bar{\phi}_s(k+1) \approx \bar{\phi}_s(k) + \bar{V}_s T_E \rightarrow \Delta \bar{\phi}_s \approx \bar{V}_s T_E \quad (6)$$

$$T_e = k(\bar{\phi}_s \times \bar{\phi}_r') = k |\bar{\phi}_s| |\bar{\phi}_r'| \sin(\delta) \quad (7)$$

More over:

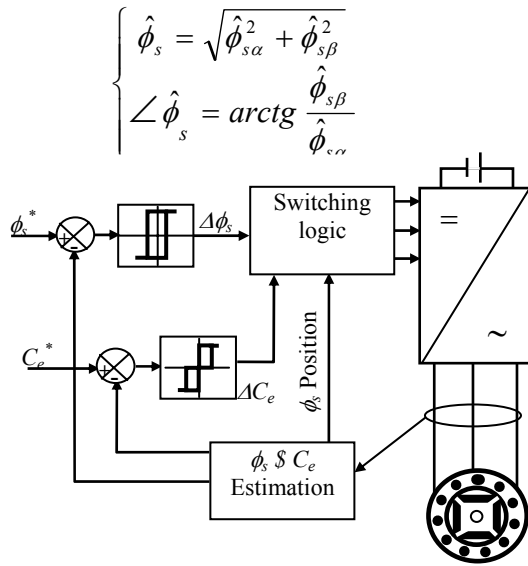


Fig. 1 Diagram of *DTC* control applied for *MSAP* supplied with a three-phase inverter with *PMW*.

A two levels classical voltage inverter can achieve seven separate positions in the phase corresponding to the eight sequences of the voltage inverter.

These positions are illustrated in Fig .2. In addition, Table II shows the sequences for each position, such as: $S_i = 1, \dots, 6$, S_i are the areas of localization of stator flux vector. Three levels *NPC* inverter release 19 vector voltages and 16 different sequences switches. These positions are illustrated in Figures 4. Furthermore, Tables I and II have the sequences corresponding to the position of the stator flux vector in different sectors (see Figures 1 and 2). The flux and torque are controlled by two comparators with hysteresis illustrated in Figure 3. The dynamics torque are generally faster than the flux then using a comparator hysteresis of several levels, is then justified to adjust the torque and minimize the switching frequency average [2].

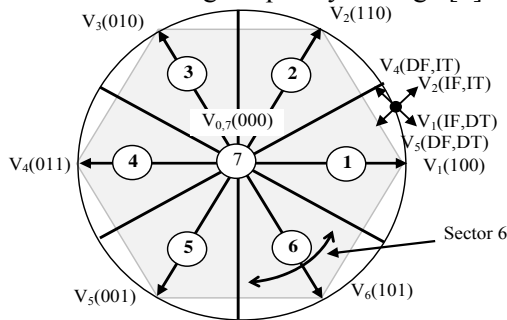


Fig. 2. Different vectors of stator voltages provided by a two levels inverter.

Where:

I(D)F : Increase (Decrease) of Flux amplitude.

I(D)T : Increase (Decrease) of Torque.

Tab. I vectors Voltage localization Table (two levels inverter)

$\Delta\phi_s$	ΔC_e	S_1	S_2	S_3	S_4	S_5	S_6
1	1	110	010	011	001	101	100
	0	000	000	000	000	000	000
	-1	101	100	110	010	011	001
0	1	010	011	001	101	100	110
	0	000	000	000	000	000	000
	-1	001	101	100	110	010	011

Where, $S_i=1\dots,6$ are localization sectors of the stator vector flux.

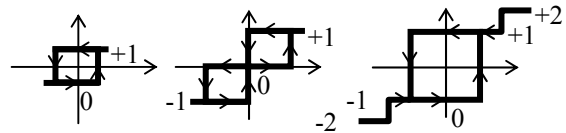


Fig.3 Comparators with hysteresis used to regulate flux and torque.

The same principle of *DTC* command of *PMSM* supplied by an inverter two levels applied for three *NPC* inverter unless the comparators hysteresis are 3 and 5 levels, which increases the number of sequences to apply. In the case of 3-level inverter (see Figure 4), sequence voltage 211 is similar to the sequence 100 so 221 is like 110, etc.... So we have 19 combinations of vectors achievable voltages including zero sequence.

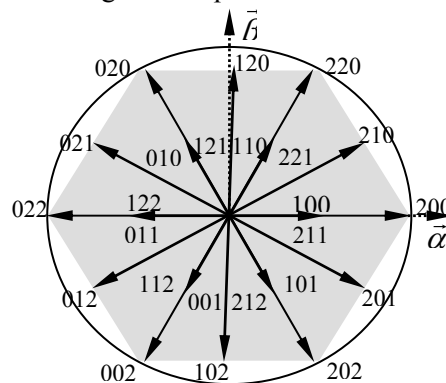


Fig. 4. Different vectors of stator voltages provided by a three levels inverter.

Tab. II vectors Voltage localization Table (tree levels inverter)

$\Delta\phi_s$	ΔC_e	S_1	S_2	S_3	S_4	S_5	S_6
1	2	220	020	022	002	202	200
	1	210	120	021	012	102	201
	0	200	220	020	022	002	202
	-1	201	210	120	021	012	102
	-2	202	200	220	020	022	002
	2	120	021	012	102	201	210
	1	120	021	012	102	201	210
0	0	000	000	000	000	000	000
	-1	102	201	210	120	021	012
	-2	102	201	210	120	021	012
	2	020	022	002	202	200	220
	1	121	122	112	212	211	221
-1	0	211	221	121	122	112	212
	-1	112	212	211	221	121	122
	-2	002	202	200	220	020	022

4. Full State Sliding Mode Observer Synthesis

The sliding mode control (SMC) is a powerful method to control nonlinear dynamic systems operating under uncertainty conditions [15]-[20]. The technique consists of two stages. First, a sliding surface, to which the controlled system trajectories must belong, is designed with accordance to some performance criterion. Then, a discontinuous control is designed to force the system state to reach the sliding surface such that a sliding mode occurs on this manifold. When sliding mode is realized, the system exhibits robustness properties with respect to parameter perturbations and external matched disturbances. In spite of claimed robustness properties, high frequency oscillations of the state trajectories around the sliding manifold known as chattering phenomenon [18],[19] are the major obstacles for the implementation of SMC in a wide range of applications. A number of methods have been proposed to reliably prevent chattering: among them, observer-based solution [19].

The sliding mode observer for estimating rotor position angle and speed is based on a stator current estimator using discontinuous control. Due to the fact that only stator currents are directly measurable in a PMSM drive. In this way, when the estimated currents, i.e., state, reach the manifold and then the

sliding mode happens and has been enforced, the current estimation error becomes zero and the estimated currents track the real ones regardless of certain disturbances and uncertainties of the drive system[20].

In pole smooth (PMSM), we can assume that ($L_d \approx L_q$), it allows us to simplify the calculation [21]. The model of PMSM linked to the stator can be written [16], [22]:

$$\dot{X} = AX + BU + K_E \zeta_s + \varsigma \quad (9)$$

Where

$$X = [i_{as} \ i_{\beta s}]^T, \ U = [v_{as} \ v_{\beta s}]^T, \ \zeta_s = [\zeta_{as} \ \zeta_{\beta s}]^T$$

$$A = \frac{r}{L_d} \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, \ B = \frac{1}{L_d} \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix},$$

$$C = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$$

$$\zeta_{as} = K_e \omega_r \sin(\psi), \ \zeta_{\beta s} = -K_e \omega_r \cos(\psi)$$

Then the model of the observer can be written as follows [22]:

$$\dot{\hat{X}} = A\hat{X} + BU + B\hat{\zeta}_s - K \quad (10)$$

Where

$$K = K_1 S + K_2,$$

$$K_1 = \begin{bmatrix} K_{11} & 0 \\ 0 & K_{12} \end{bmatrix} \text{ and } K_2 = \begin{bmatrix} K_{21} \\ K_{22} \end{bmatrix}$$

The sliding surface S is defined by:

$$S = [s_1 \ s_2]^T = [i_{as} - \hat{i}_{as} \ i_{\beta s} - \hat{i}_{\beta s}]^T = e_{is} = 0 \quad (11)$$

Moreover, the terms of errors in current estimates are given by the following equations:

$$\dot{S} = \frac{d}{dt}(\hat{i}_s - i_s) = (\hat{A}\hat{i}_s - A_s) + B(\hat{\zeta}_s - \zeta_s) + K \quad (12)$$

To estimate the speed and stator resistance, using the Lyapunov function V chosen as [22]:

$$V = \frac{1}{2} S^T S + \lambda_1 \left(\frac{\hat{\omega}_r - \omega_r}{2} \right)^2 + \lambda_2 \left(\frac{\hat{r}_s - r_s}{2} \right)^2 \quad (13)$$

Where λ_1 and λ_2 are positive constants.

The derivative of this Lyapunov function V is:

$$\dot{V} = \dot{S}^T S + \lambda_1 (\hat{\omega}_r - \omega_r) \dot{\hat{\omega}}_r + \lambda_2 (\hat{r}_s - r_s) \dot{\hat{r}}_s \quad (14)$$

Substituting (8) into (10), we find:

$$\dot{V} = \dot{S}^T [(A - A) \hat{i}_s + A(\hat{i}_s - i_s) + B(\hat{\zeta}_s - \zeta_s) - K] + \lambda_1 \Delta \omega_r \dot{\hat{\omega}}_r + \lambda_2 \Delta r_s \dot{\hat{r}}_s \quad (15)$$

Where $\Delta \omega_r = \hat{\omega}_r - \omega_r$ and $\Delta r_s = \hat{r}_s - r_s$ are obtained checking the stability criterion of Lyapunov ($\dot{V} < 0$). Indeed, it is obligatory to guarantee the stability of the *SMO*, this choosing for example [22]:

$$S^T [(A - A) \hat{i}_s] + S^T [B(\hat{\zeta}_s - \zeta_s)] + \lambda_1 \Delta \omega_r \dot{\hat{\omega}}_r + \lambda_2 \Delta r_s \dot{\hat{r}}_s = 0 \quad (16)$$

and

$$S^T [A(\hat{i}_s - i_s) - K] < 0 \quad (17)$$

We can thus derive an algorithm for estimating speed and stator resistance starting from the relation (16). Just after rearrangement to choose:

$$S^T [(A - A) \hat{i}_s] + \lambda_2 \Delta r_s \dot{\hat{r}}_s = 0 \quad (18)$$

and

$$S^T [B(\hat{\zeta}_s - \zeta_s)] + \lambda_1 \Delta \omega_r \dot{\hat{\omega}}_r = 0 \quad (19)$$

Therefore, the derivative of the estimated stator resistance is obtained as:

$$\dot{\hat{r}} = \frac{1}{\lambda_2 l_d} [(s_1 \hat{i}_{\alpha s} + s_2 \hat{i}_{\beta s})] \quad (20)$$

And the derivative of the estimated speed is obtained from the flows equation:

$$\dot{\hat{\omega}}_r = \frac{K_e}{\lambda_1 l_d} [(s_1 \sin \hat{\psi} - s_2 \cos \hat{\psi})] \quad (21)$$

The other by the estimated rotor position $\hat{\psi}$ is obtained by integrating the estimated speed from (21).

Now, to ensure the stability condition of *Lyapunov* ($\dot{V} < 0$) [22], we choose the gains of the observers K_1 and K_2 to satisfy the following equation:

$$S^T [A(\hat{i}_s - i_s) - K_1 S - K_2] < 0 \quad (22)$$

hence this condition achieved as :

$$K_1 < A \text{ and } S^T K_2 < 0 \quad (23)$$

To determine K_1 and K_2 , we must then verify the condition (17), it is obtained as :

$$K_{11}, K_{12} > \frac{\hat{r}}{l_d} \quad (24)$$

$$\text{and } K_{2i} = \begin{cases} \alpha_i & \text{if } s_i > 0 \\ -\beta_i & \text{if } s_i < 0 \end{cases} \quad (25)$$

Where α_i and β_i ($i = 1, 2$) are positive constants.

5. Results of Simulation

Table (III), summarizes the *PMSM* parameters used in this simulation[23].

Pole pairs	3
Rated power KW (at 50 Hz)	1.5
Rated voltage (V)	220/380
Rated Flux (Wb)	0.30
Rated torque (Nm)	5
R_s (Ω)	1.4
$L_d ; L_q$ (H)	0.0066; 0.0058
Flux magnet (Wb)	0.15
J (Kg.m ²)	0.00176
f_r (Nm/(rad/s))	0.0038

5.1. Sensorless control results at nominal speed.

We simulated the system drive for a reference speed of 100 (rd / s) load at startup. At $t = 0.1$ (s), the *PMSM* is tracking load equal to 5 (Nm), then from $t = 0.02$ (s), we assumed a variation of the stator resistance (see Fig. 5). Finally, we reversed the direction of rotation from time $t = 0.22$ (s), changing the speed set point from 100 to -100 (rd / s).

The results are obtained using a *PI* speed controller and gains of the *SMO* are obtained after several trials of simulations to achieve the best results. The following values are then adopted:

$$K_{11}=999, K_{12}=499, K_{21}=8, K_{22}=8, \lambda_1=0.02, \lambda_2=1.$$

Figure 5 illustrates the evolution of stator resistance, actual and estimated (delivered by the propose *SMO*). The two quantities are combined in practice, in steady state. As in Figure 6, it illustrates estimated speed (rad / s) issued by *SMO*, the speed response is achieved without dip and with a shorter recovery time which is almost similar with the actual speed motor.

Figure 7, shows the stator flux estimation using the *SMO* with two and three level inverter. We notice that it is not affected by these changes. However, the variation of stator resistance slightly disturbs the electromagnetic torque transient. In steady state, the

electromagnetic torque follows his rate as shown in Figure 9.

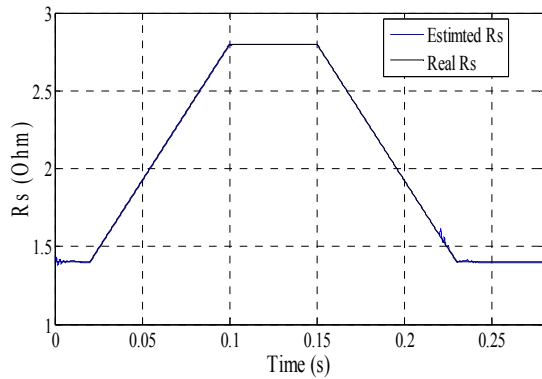


Fig. 5. Estimated and real stator resistance.

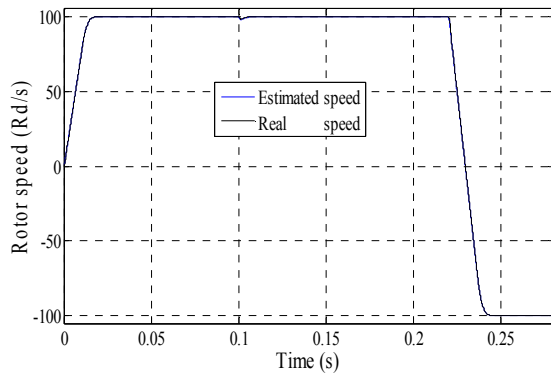


Fig.6. Estimated and real speed with reversed direction

The use of three-level *NPC* inverter has improved the band Electromagnetic Torque are shown in Figure 10. Indeed this reduction is worth approximately 31.43% of the torque value which is an important advantage.

The error between the estimated and actual current does not exceed 0,005 (A) transient and disappears almost in steady state as shown in Figure 11.

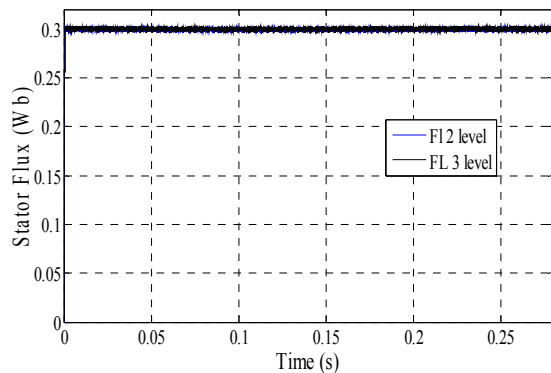


Fig. 7. Evolution of motor's stator Flux.

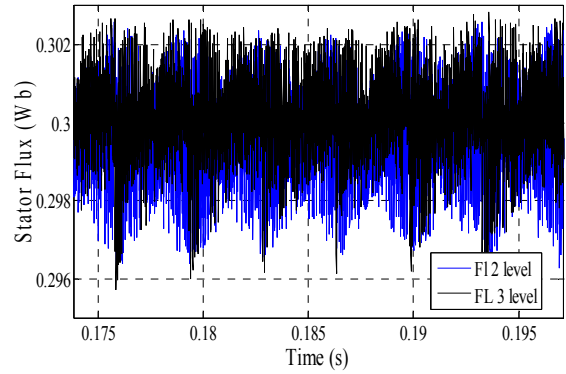


Fig. 8. Scale of the estimated Stator Flux.

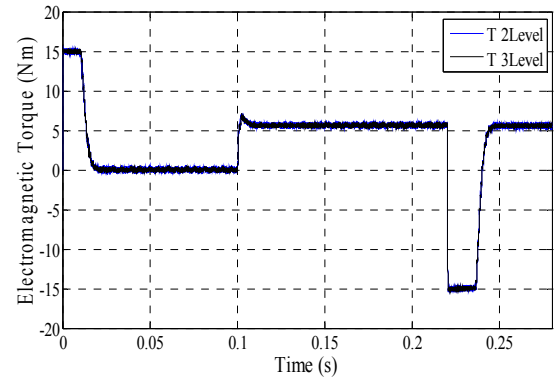


Fig. 9. Evolution of motor's Electromagnetic torque.

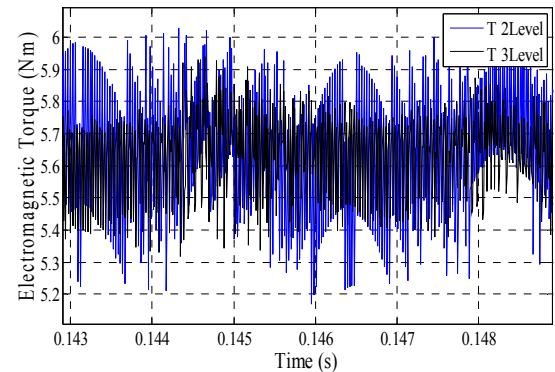


Fig.10. Scale of the estimated Electromagnetic Torque.

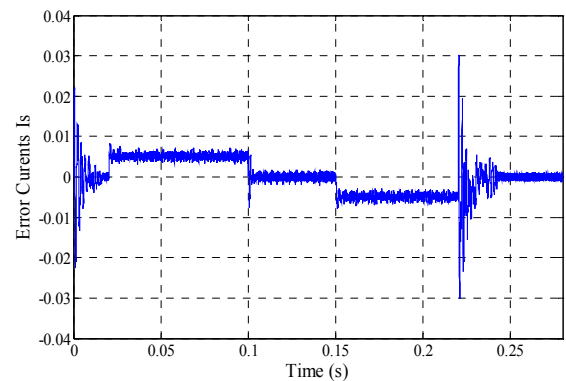


Fig. 11. Evolution error stator currents.

5.2. Sensorless control results at very low speeds.

In this section, some simulation results were obtained using the control laws for very low speed sensorless *DTC* proposed. The results correspond to the three studied cases; we tried to evaluate the robustness of the control. The first case was when the control law was based on under *PMSM* flux weakening operating (- 12% at $t=0.12(s)$) with considering some variations of motor parameters ($\Delta R_s = 100\%$ at $t=0,10(s)$, $\Delta L_d = +10\%$ at $t=0,14(s)$, $\Delta J_n = -25\%; +50\%; +100\%$ at $t=0(s)$). A second case was when the control was based on scaling torque load and speed with no machine parameters variation (stator inductance and resistance). A third and last case was when the stator inductance and resistance and inertia values were deviated with respect to the nominal values ($\Delta R_s = 100\%$ at $t=0.1(s)$, $\Delta L_d = +10\%$ at $t=0.14(s)$, $\Delta J_n = +50\%; +100\%$ at $t=0.16(s)$) with torque load and speed variation.

From these tests, it can be noted that during transient and with modify load, the sensorless *DTC* performance in this case is more improved.

We can see in Figure (12) peaks of Torque that appeared right at the time of the change speed motor (40rd/s —10rd/s —3 rd/s—1rd/s) and disappears almost steady state. Figure 13 shows the good performance response of flux . Figure 14 shows speed motor performance in the low speed region considering some key parameters variation. We can notice that a quick and a stable response is obtained at rated operating conditions. In fact, a negligible steady-state error is obtained either at high or low speed and good tracking is achieved during transient. There is no sensible difference on the speed trajectories.

Figures 14, 20 shows the Evolution of motor's Electromagnetic Torque when the motor is operated at different speed with and without parameters variation,

As the speed is decreased, however, the steady state values of the estimated speed tend to deviate from those of the actual speed as shown in figures 17. Sliding mode control of *PMSM* is very robust with respect to parametric uncertainties. Figure 21 shows the drive dynamic under different values of inertia with scaling speed reference. It is clear that the speed tracking is little affected by those changes.

Thus, the position error at medium and low speeds is not important at all. The speed estimation error really counts in the speed control loop. Note that, if

the rotor position error is rather constant, then the speed is correctly estimated.

Case 1: control results at very low speeds

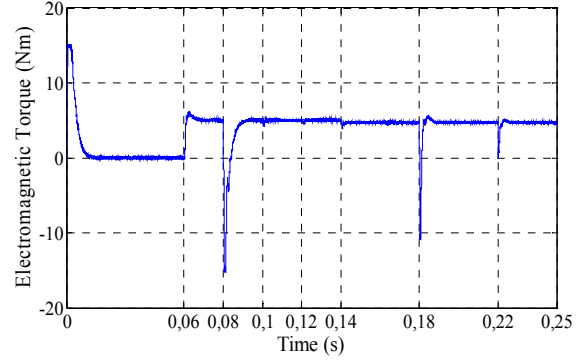


Fig.12. Evolution of motor's Electromagnetic torque.

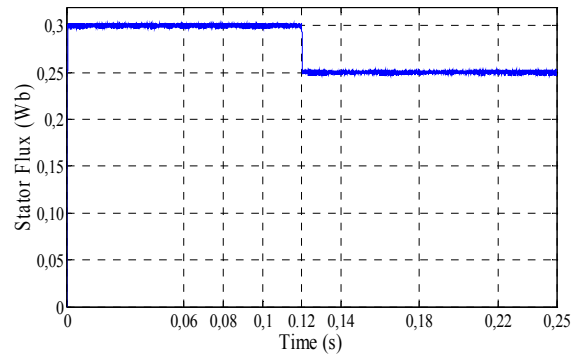


Fig. 13. Evolution of motor's stator Flux.

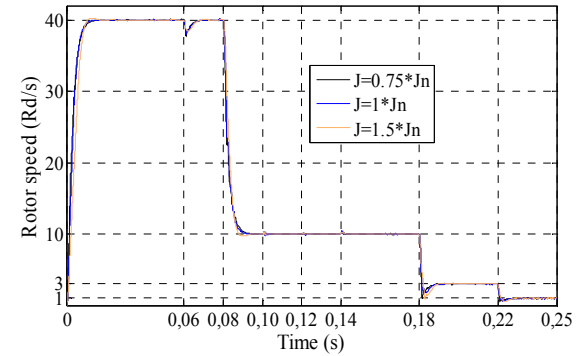


Fig.14. Estimated and real speed

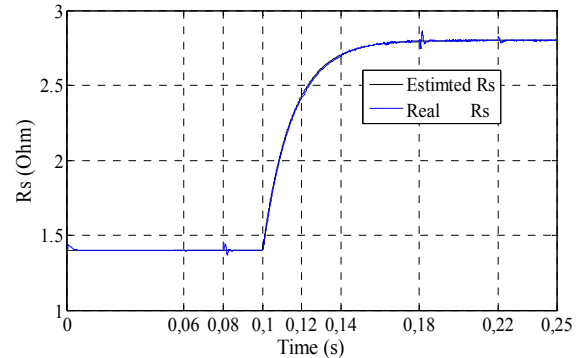


Fig.15. Estimated and real resistance

Case 2: Control with no parameters variation

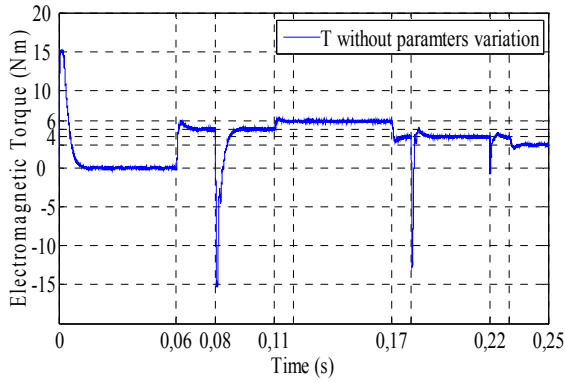


Fig. 16. Evolution of motor's Electromagnetic torque.

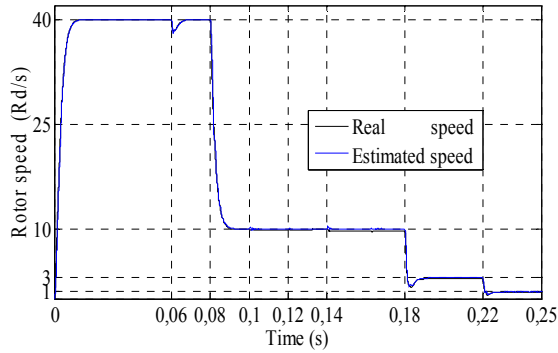


Fig.17. Estimated and real speed motor.

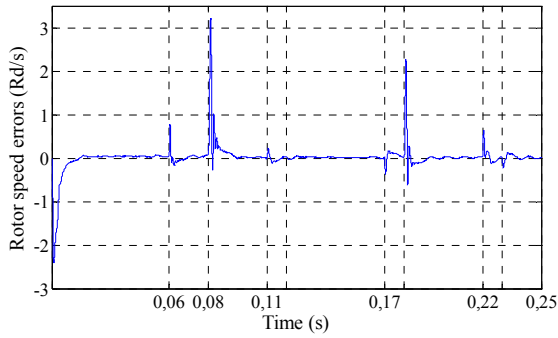


Fig. 18. Evolution error speed without machine parameters variation

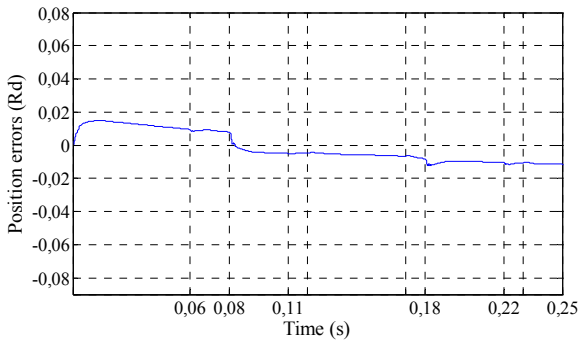


Fig. 19. Evolution error position without machine parameters variation.

Case 3 : Control with parameters variation

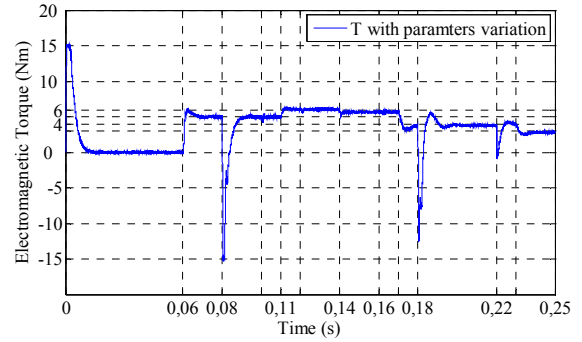


Fig. 20. Evolution of motor's Electromagnetic torque with machine parameters variation.

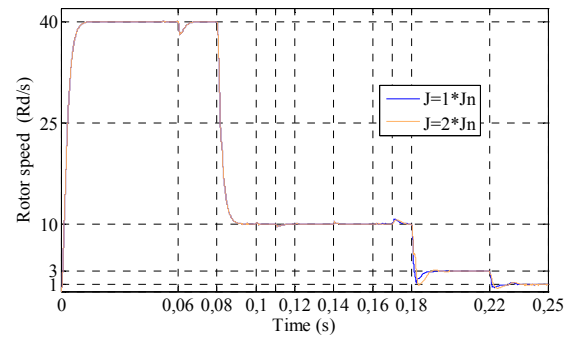


Fig. 21. Evolution error speed with machine parameters variation.

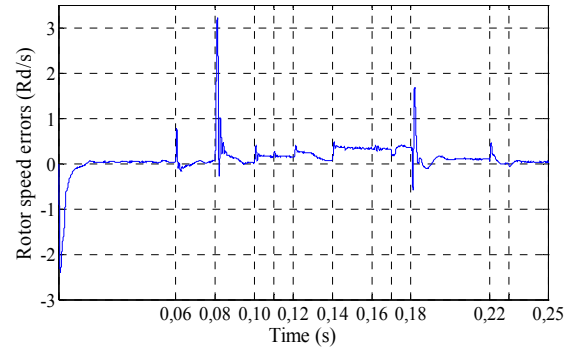


Fig. 22. Evolution error speed with machine parameters variation.

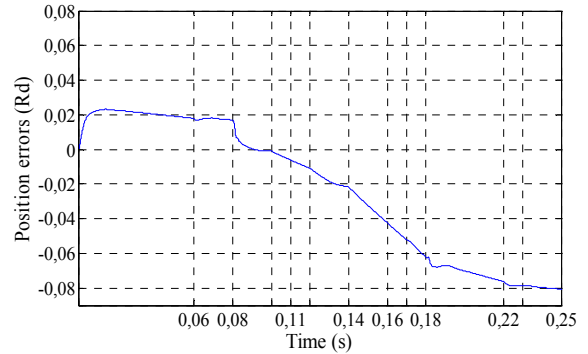


Fig. 23. Evolution error position with machine parameters variation.

7 Conclusions

In this paper, we proposed a sliding mode observer (SMO) for estimating the stator resistance and speed of a PMSM, and to achieve speed sensorless DTC. The effect of the variations of motor parameters such as torque constant, stator resistance and stator inductance on speed estimations, at very low range, have been studied. Using three-level NPC inverter reduces the torque ripple of PMSM performance compared to obtain with a two-level inverter. The simulation results obtained were satisfactory, and system stability has been insured.

8 References

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