ROBUST DTC OF AN ADJUSTABLE SPEED SENSORLESS SWITCHED RELUCTANCE MOTORS BASED ON SVM USING A PI PREDICTIVE CONTROLLER

M. BIRAME L.MOKRANI
Materials Laboratory, Electrical Engineering Department, Laghouat University, BP 37G, Ghardaia Street, Laghouat (03000), Algeria, E-mail : bir_moh@yahoo.fr, Mokrani_lakhdar@hotmail.com

B. AZOUI
LEB Laboratory, Electrical Engineering Department, Batna University, Chahid M.E.H. Boukhlouf Street, Batna (05000), Algeria, Azoui_b@yahoo.com

A. Naamane N.K. Msird
LSIS - Laboratory of the System and Information sciences,marseille, Paul Czanne University , FRANCE
aziz.naamane@lsis.org, nacer msirdi@lsis org

Abstract: In This study presents a new sensorless direct torque control method for voltage inverter fed RSM. The control method is used a modified Direct Torque Control scheme with constant inverter switching frequency using Space Vector Modulation (SVPWM). The variation of stator resistance due to changes in temperature or frequency deteriorates the performance of DTC-SVM controller by introducing errors in the estimated flux linkage and the electromagnetic torque. As a result, this approach will not be suitable for high power drives such as those used in tractions, as they require good torque control performance at considerably lower frequency. A novel stator resistance estimator is proposed. The estimation method is implemented using the Extended Kalman Filter. Finally, extensive simulation results are presented to validate the proposed technique. The system is tested to change in value of the stator resistance and a very satisfactory efficiency was obtained..

Key words: *RSM, DTC-SVPWM, observerless and sensorless speed control, Extended Kalman Filter (EKF).DTC, SVM,.*

1. Introduction.

Reluctance Synchronous Motors (RSM) has attracted significant interest of industry due to their main advantages are [1-3]:
- Simplicity and robustness
- High torque overloads capacity
- High efficiency over wide speed-range
- Low machine inertia
- decreased maintenance requirements

The absence of windings and magnets on the rotor enables SR motors to run high speed and temperature. An SR motor can produce large torque in a wide speed range. However, it has some drawbacks such as torque ripple and acoustic noise. High torque ripple is inevitable unless ripple reduction strategy is applied. Firstly DTC was proposed for IM [3], however now is applied also for PMSM and RSM[4]. Direct Torque Control (DTC) seems to be a good performance alternative to the classical vector control drives. After its implementation on induction motor drives, this control method in recent years has been proposed for synchronous reluctance motor with good results. DTC is able to produce fast torque and stator flux response with a well designed flux, torque and speed estimator. In order to reduce the torque and current pul- sations, in steady state a mixed DTC- SVM control method seems more suitable [1, 6].

SVM techniques [5 6] offer better DC link utilization and they lower the torque ripple. The emphasis of research on RSM has been on sensorless drive [7, 8, 9, 10], which eliminates flux and speed sensors mounted on the motor. In addition, the development of effective speed and flux estimators has allowed good rotor flux-oriented performance at all speeds except those close to zero. Sensorless control has improved the motor performance, compared to the Volts/Hertz (or constant flux) controls.

The EKF is considered to be suitable for use in high-performance RSM drives, and it can provide accurate speed estimates in a wide speed-range, including very low speed [5] [10, 11].

The variation of stator resistance due to changes in tempera- ture or frequency deteriorates the performance of DTC controller by introducing errors in the estimated flux linkage and the electro- magnetic torque [13-16]. A novel stator resistance estima- tor during the operation of the motor is proposed.

This paper describes a novel DTC-SVM method for a speed sensorless control of RSM drive. According to this method, a conventional PI predictive controller is used to determine the po- lar components of the voltage command vector. The results show that a satisfactory control performance is obtained.

2. Modeling of the RSM.

The electrical and mechanical equations of the RSM in the rotor reference (d,q) frame as follows

$$\left\{\begin{array}{l}\dfrac{dI_d}{dt}=-\dfrac{R_s}{L_d}+\dfrac{L_q}{L_d}P\mu_r I_q+\dfrac{1}{L_d}U_d\\ \dfrac{dI_q}{dt}=-\dfrac{R_s}{L_q}-\dfrac{L_d}{L_q}P\mu_r I_d+\dfrac{1}{L_q}U_q\\ \dfrac{d\mu_r}{dt}=\dfrac{3P}{2J}\left(L_d-L_q\right)I_d I_q-\dfrac{1}{J}T_L-\dfrac{B}{J}\mu_r\end{array}\right\} \quad (1)$$

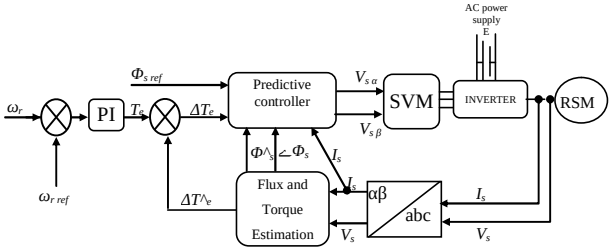


Fig. 1. RSM Drive System Control.

where I_d, I_q are the d-q axis currents, U_d, U_q are the d-q axis voltages, ωr denotes the rotor speed, R_s is the stator resistance, L_d, L_q are the stator inductances, P is the pole pairs, J is the rotor moment of inertia, B is the viscous friction coefficient, T_L is the load torque.

The block scheme of the investigated direct torque control with space vector modulation (DTC-SVM) for a voltage source PWM inverter fed RSM is presented in Fig 1(a). [5, 6]

The internal structure of the predictive torque and flux controller is shown in Fig 1(b).

The objective of the DTC-SVM scheme, and the main difference between the classic DTC, is to estimate a reference stator voltage vector V_{sref} in order to drive the power gates of the inerter with a constant switching frequency. Although, the basic principle of the DTC is that the electromagnetic torque of the motor can be adjusted by controlling the angle Δd between the stator and rotor flux vectors, the torque of a RSM can be calculated by the following equation [5, 6]

$$T_e=\frac{3}{2}P\frac{\Phi_{s\_ref}}{L_d L_q}\left(\frac{1}{2}\Phi_{s\_ref}\left(L_d-L_q\right)\sin 2¢\right) \quad (2)$$

Where the change in the stator flux vector, if we neglect the voltage drop in the stator resistance, can be given by the following equation,

$$\Delta\Phi_s=V_s T_s \quad (3)$$

Where ΔΦs is the deviation from |Φs| which are defined by:

$$\Delta\Phi_s=\left|\Phi_{sref}\right|-\left|\Phi_s\right| \quad (4)$$

The predictive controller determinates the stator voltage command vector in polar coordinates Vs ref [V_{sref}, δ] for space vector modulator; witch finally generates the pulses S_a, S_b, S_c.

Sampled torque error ΔT_e and reference stator flux amplitude Φ_s ref are delivered to the predictive controller. The relation between error of torque and increment of load and angel Δδ is nonlinear. Therefore PI controller, witch generates the load angle increment required to minimize the instantaneous error between reference T_{eref} and actual T_e torque, has been applied. The reference values of the stator voltage V_{s_ref},, γs ref is calculated based on stator resist- ance Rs, Δδ signal, actual stator current vector Is, actual stator flux amplitude Φ_s and position γ_s as: The α, β axes components of the stator reference voltage V_{s_ref},, are calculated according to the following equation[5,13]::

$$V_{s\sim ref}=\frac{\Phi_{sref}\cos(¥_s+\Delta¢)-\Phi_{sref}\cos¥_s)}{T_s}+\hat{R}_S I_{s\sim} \quad (5)$$

$$V_{s\ ref}=\frac{\Phi_{sref}\sin(¥_s+\Delta¢)-\Phi_{sref}\sin¥_s)}{T_s}+\hat{R}_S I_s \quad (6)$$

$$V_{sref}=\sqrt{V_{s\sim ref}{}^2+V_{s\ ref}{}^2} \quad (7)$$

$$¥_{sref}==\arctan\left(\frac{V_{s\ ref}}{V_{s\sim ref}}\right) \quad (8)$$

Where, T_s is sampling time.

3. Voltage Space Vector Modulation.

In the space vector modulation techniques, the reference voltages are given by space voltage vector and the output voltages of the inverter are considered as space vectors. There are eight possible output voltage vectors, six active vectors V_1- V_6, and two zero vectors V_0, V_7 (Figure 2). The reference voltage vector is realized by the sequential switching of active and zero vectors. In the Figure.3 there are shown reference voltage vector V_{ref} and eight voltage vectors, which correspond to the possible states of inverter. The six active vectors divide a plane for the six sectors 1- 6. In the each sector the reference voltage vector V_{ref} is obtained by switching on, for a proper time, two adjacent vectors

Presented in Figure.1 the reference vector V_{ref} can be implemented by the switching vectors of V_1, V_2 and zero vectors V_0, V_7

The figure shows the case when the reference vector is in the sector 1.

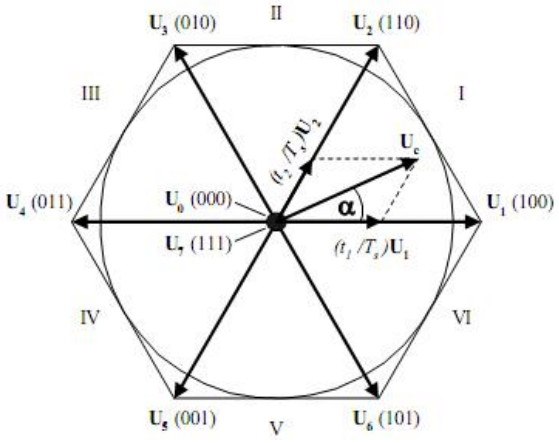


Fig. 3. Principle of the space vector modulation

For example, in the first sector, Vsref is a synthesized voltage space vector and expressed by:

$$V_{sref} = V_0T_0 + V_1T_1 + V_2T_2 \tag{9}$$

The times T_1 and T_2 are obtained by simple trigonometrical relationships and can be expressed in the following equations

$$T_s = T_0 + T_1 + T_2 \tag{10}$$

Where T_0, T_1 and T_2 is the work time of basic space voltage vectors V_0, V_1 and V_2 respectively. [5-7] [9] The determination of the amount of times T_1 and T_2 is given by simple projections

$$T_1 = \frac{T_s}{2E}\left(\sqrt{6}\ V_{s\ ref} - \sqrt{2}V_{s\ ref}\right) \tag{11}$$

$$T_2 = \sqrt{2}\frac{T_s}{E}V_{s\ ref} \tag{12}$$

The rest of the period spent in applying the null-vector. For every sector, commutation duration is calculated. The amount of times of vector application can all be related to the following variables:

$$X = \frac{T_s}{E}\sqrt{2}V_{s\ ref} \tag{13}$$

$$Y = \frac{T_s}{E}\left(\frac{\sqrt{2}}{2}V_{s\ ref} + \frac{\sqrt{6}}{2}V_{s\sim ref}\right) \tag{14}$$

$$Y = \frac{T_s}{E}\left(\frac{\sqrt{2}}{2}V_{s\ ref} + \frac{\sqrt{6}}{2}V_{s\sim ref}\right) \tag{15}$$

The application durations of the sector boundary vectors are tabulated as follows:

Table 1. Durations of the sector boundary vectors.

SECTOR	1	2	3	4	5	6
T_i	z	Y	-z	-x	x	-Y
T_2	Y	-x	x	z	-Y	-z

The goal of this step is to compute the three necessary duty cycles as;

$$T_{aon} = \frac{T_s - T_1 - T_2}{2} \tag{16}$$

$$T_{bon} = T_{aon} + T_1 \tag{17}$$

$$T_{con} = T_{bon} + T_2 \tag{18}$$

The last step is to assign the right duty cycle (T_{aon}) to the right motor phase according to the sector.

Table 2. Assigned duty cycles to the PWM outputs.

SECTOR	1	2	3	4	5	6
S	T_{bon}	T_{aon}	T_{aon}	T_{con}	T_{bon}	T_{con}
S	T_{aon}	T_{con}	T_{bon}	T_{bon}	T_{con}	T_{aon}
S	T_{con}	T_{bon}	T_{con}	T_{aon}	T_{aon}	T_{bon}

4. EKF state estimation.

The Kalman filter is in principle a state observer that gives an estimation of the state variables of a system, by the minimization a square error, when both the system inputs and outputs are subject to random disturbances. The EKF can also be used for unknown parameter estimation [8]. Nonlinear discrete systems model with white noise is given as follows [5] [10,11]:

$$\begin{cases} X(k+1) = f\left(X(k),u(k)\right) + W(k) = A_d X(k) + B_d U(k) + W(k) \\ Y(k) = h\left(X(k)\right) + V(k) = C_d X(k) + V(k) \end{cases} \tag{19}$$

With: w(k) is the measurement noise and v(k): is the process noise, Ad, Bd and Cd matrix of discrete system.

$$\begin{cases} A_d = e^{AT_s} \approx I - AT_s \\ B_d = \int_0^{T_s} e^{AT_s} Bd^2 \approx BT_s \\ C_d = C \end{cases} \tag{20}$$

I: identity matrix of system depending on the size of the state vector[5,6].

Where:, $U = \begin{bmatrix} V_d & V_q \end{bmatrix}^T$ and $Y = \begin{bmatrix} i_d & i_q \end{bmatrix}^T$

$$f_1 = (1 - a_1 R_s(k)) I_d(k) + a_2 \mu_r(k) I_q(k) + a_3 U_d(k)$$

$$f_2 = a_4 \mu_r(k) I_d(k) + (1 - a_5 R_s(k)) I_q(k) + a_7 U_q(k)$$

$$f_3 = a_8 I_d(k) I_q(k) + (1 - a_{10}) \mu_r(k) + a_{11} T_l(k)$$

$$f_4 = a_{12} \mu_r(k) + \phi(k)$$

$$f_5 = T_l(k) \qquad f_6 = R_s(k)$$

$$A_d = \begin{bmatrix} (1 - a_1 R_s(k)) & a_2 \mu_r(k) & 0 & 0 & 0 & a_1 \\ -a_4 \mu_r(k) & (1 - a_5 R_s(k)) & a_6 & 0 & 0 & a_1 \\ a_8 I_q(k) & a_9 & 1 - a_{10} & 0 & a_{11} & 0 \\ 0 & 0 & a_{12} & 1 & 0 & 0 \\ 0 & 0 & 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 0 & 0 & 1 \end{bmatrix} \quad (21)$$

$$B_d = \begin{bmatrix} a_3 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & a_7 & 0 & 0 & 0 & 0 & 0 \end{bmatrix}^T$$

$$C_d = \begin{bmatrix} 1 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 & 0 & 0 & 0 \end{bmatrix}^T$$

$$a_1 = \frac{R_S}{L_d} T_s \; ; a_2 = \frac{L_q}{L_d} T_s \; ; \; a_3 = \frac{1}{L_d} T_s$$

$$a_4 = \frac{L_d}{L_q} T_s \; ; a_5 = \frac{R_s}{L_q} T_s \; ; a_7 = \frac{1}{L_q} T_s$$

$$a_8 = p \frac{L_d - L_q}{J} T_s \; ; a_{10} = \frac{B}{J} T_s \; ;$$

$$a_{11} = \frac{1}{\cdot} T_s \; ; a_{12} = p T_s$$

Where: f (x(k), u(k)): Nonlinear function vector of the states. x(k): extended state vector. A_d: system matrix. ue (k) is the control input vector, B_d: input matrix. h (x(k), v(k)): Function vector of the outputs. C_d: Measurement matrix. $w(k)$ and $v(k)$: process and measurement noise respectively. [5] [10,11]

5. Application of the Extended Kalman Filter Algotithm.

The extended kalman filter algotithm should de calculated by using matlab-simulink.

5.1 Estimation of Error Covariance Matrix.

The prediction covariance is updated by:

$$P(k+1/k) = F(k) P(k/k) F^T(k) + Q(k) \quad (22)$$

Where: Q: covariance matrix of the system noise,

$$F(k) = \left. \frac{\partial f}{\partial X} \right|_{X(k) = \hat{X}(k/k)} \quad (23)$$

5.2 Computation of Kalman Filter Gain.

The Kalman gain is given by (24);

$$L(k+1) = P(k+1/k) C^T(k) \left(C(k) P(k+1/k) C^T(k) + R(k) \right)^{-1}$$

Where: $C(k) = \left. \frac{\partial c(x(k))}{\partial x(k)} \right|_{X(k) = \hat{X}(k)}$ (24)

5.3 State Estimation.

The extended Kalman correction equation is described by (25).

$$\hat{X}(k+1/k+1) = \hat{X}(k+1/k) + L(k+1)(y(k+1) - C \hat{X}(k+1/k)) \quad (25)$$

6. Proposed Sensorless RSM Drive.

The proposed sensorless RSM drive is depicted in Figure 5. The stator flux is estimated by the EKF and used in the DTC control the parameters are listed in Table 3

Table 3. Durations of the sector boundary vectors.

Parameters of the RSM	
Rated torque	T_L=6 Nm
d-Axis inductance	L_d = 0.0072 H
q-Axis inductance	L_q = 0.0028 H
Stator resistance	R_s = 1
Moment of rotor inertia	J = 0.003 kgm^2
Viscous friction coefficient	F = 0.000035Kg.m^2
Numbers of pole pairs	P = 3

7. Sensorless Drive imulation Results.

In this section, the effectiveness of the proposed algorithm is verified by computer simulations. During the simulations, the torque set value is limited to 6 N.m (rated torque), In order to show the performances and the robustness of the combined DTC-SVM-EKF algorithm, a series of tests were conducted to check the performance of the proposed DTC-SVM. In all sketched figures, the time axis is scaled in seconds.

The block scheme of the investigated direct torque control with space vector modulation (DTC-SVM) for a voltage source inverter fed RSM is presented in (Figure 5). The specifications for the used RSM are listed in table (3).. Figure 6 shows the actual and

estimated responses of the proposed sensorless scheme The machine is started from rest and assumed to follow a certain speed trajectory. A load torque of 6 N.m. is assumed to be applied at time 0.15s, with a change in value of the stator resistance of 1Ω to 2Ω ohm and 2Ω to 1Ω. Current ripple has also a notable reduction in DTC-SVM compared to classic DTC.

DTC-SVM has a significantly lower ripple level both in torque, flux and stator current, a lower current ripple advantageous because the machine will have less EMI noise. Figure 6 show the trajectory of the estimated stator flux components DTC-SVM has as good dynamic response as the classical DTC

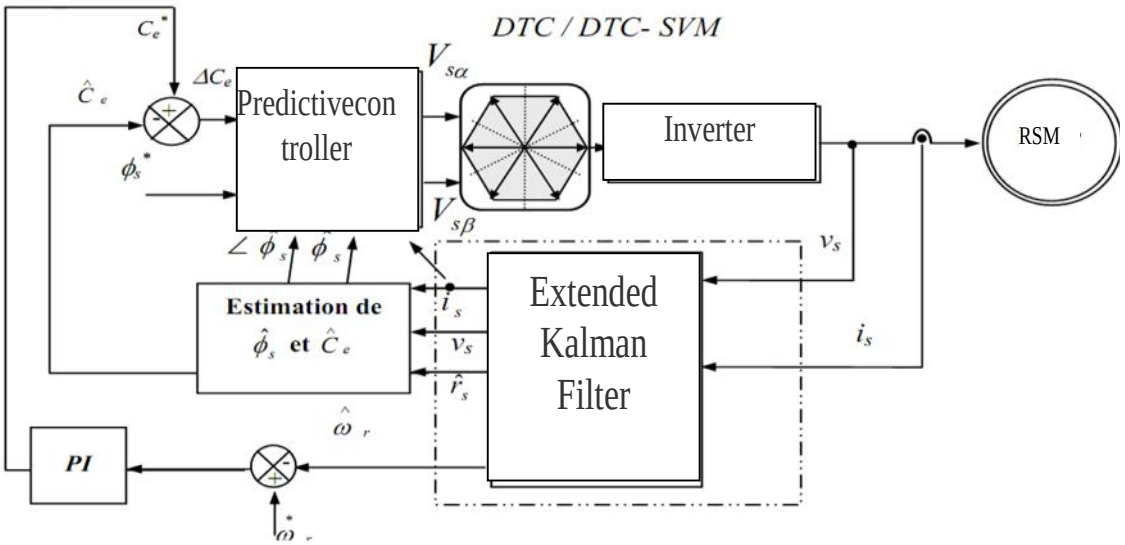
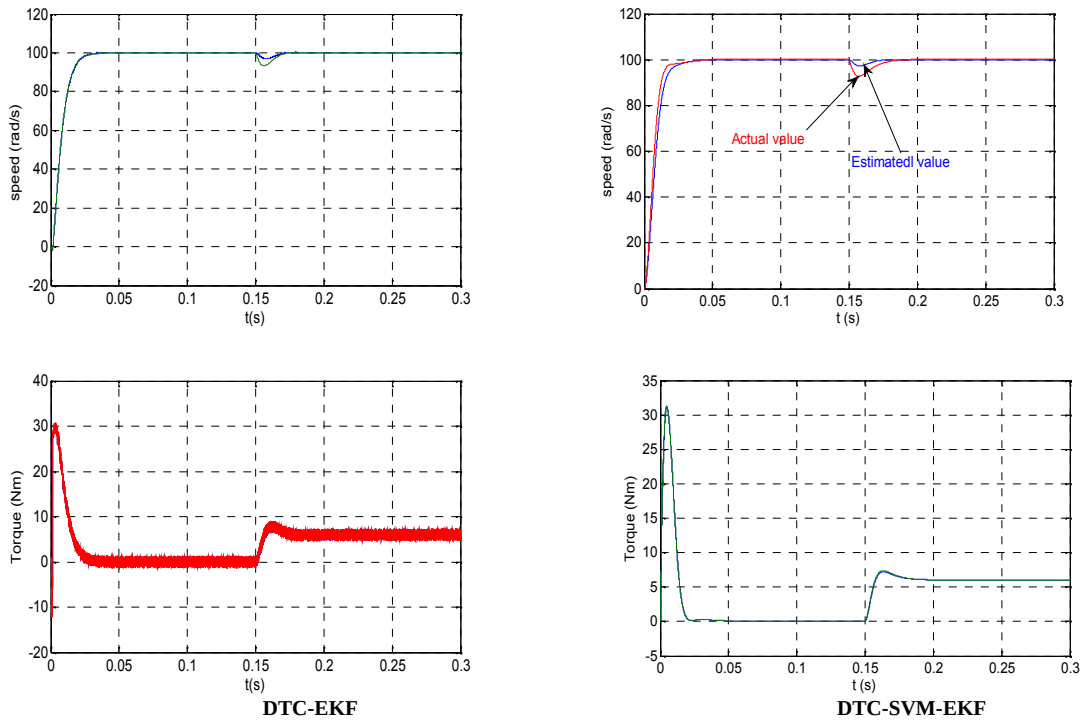


Fig. 5. Block diagram of the sensorless system



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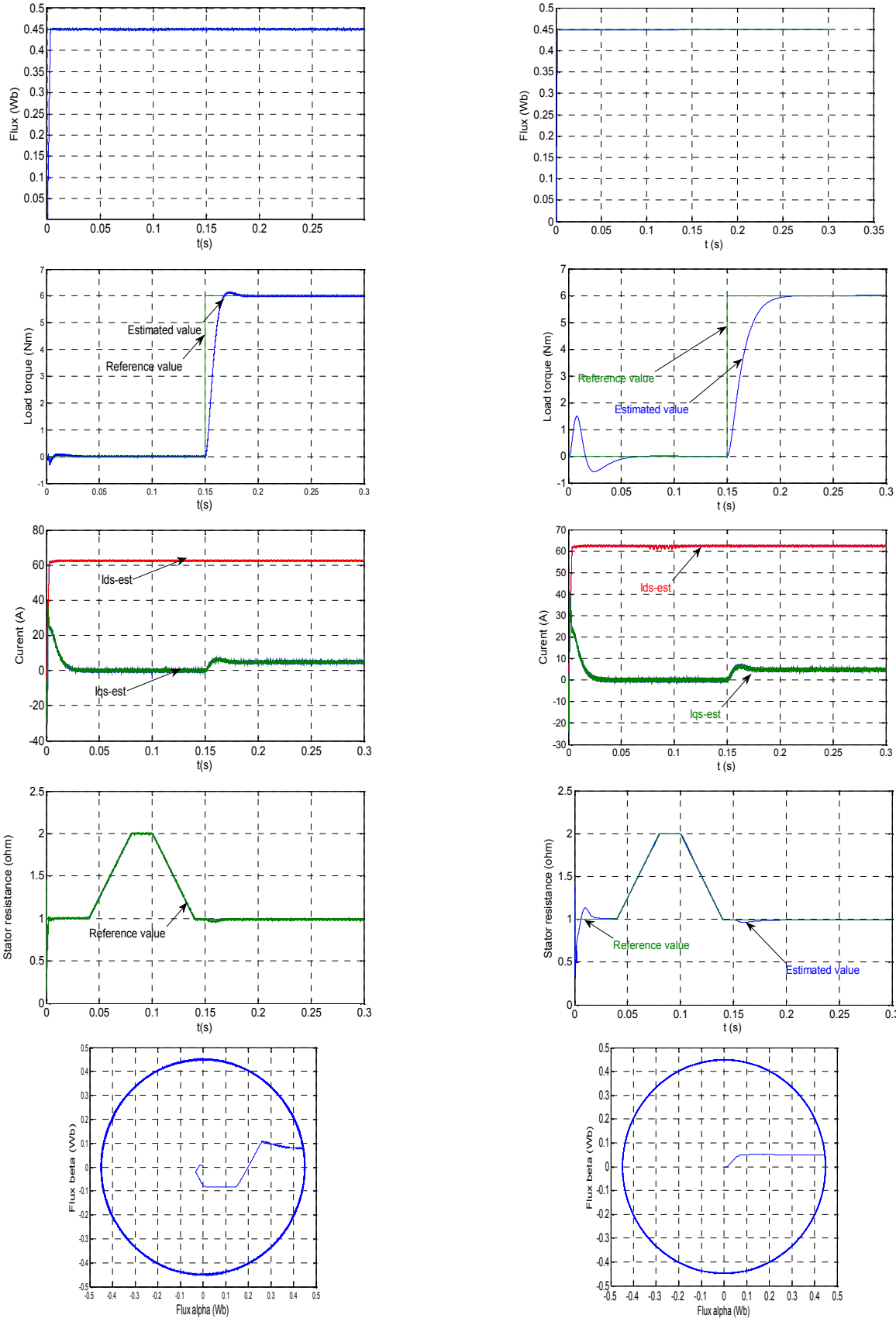

DTC-EKF **Classical DTC-EKF**

Fig. 6**.** Simulation results: A load torque of 6 N.m) is applied at t = 0.15 sec.

8. Conclusion.

As an estimator, the Extended Kalman filtering technique is developed for the speed sensorless direct torque control strategy combined with space vector modulation. The complete sensorless solution is presented with the combined DTC-SVM-EKF strategy; low torque ripple operation has been obtained with PMSM In spite of lower switching frequency, the DTC-SVM scheme has lower harmonic current, and consequently lower ripple than conventional hysteresis based DTC. Simulation results obtained clearly demonstrate the effectiveness of the estimator in estimator in estimating the stator resistance and improving performance of DTC. Additionally, the application of SVM guarantee:
Good dynamic response; High robustness; Low sampling frequency is required; Distortion caused by sector changes is delimited; Inverter switching frequency is constant.

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