An Extended Kalman Filter For Sensorless Direct Torque and Field Controlled PMSM Speed Drive Using SVM Approach

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Abstract *— In this paper a speed sensorless Direct Torque and Flux Control (DTFC) method based on space vector modulation (SVM) and extended Kalman Filter (EKF) is proposed. This method is based on the PMSM models in the coordinate of stator flux linkage, flux and torque are controlled through stator voltage components in stator flux linkage coordinate axes and space vector modulation is used to control inverters. Simulation results verify that this proposed sensorless DTFC scheme could effectively decrease flux and torque ripples, and the speed estimation method could accurately track the motor speed and has good dynamic and static performance.*

Index Terms *—Permanent magnet synchronous motor, direct torque control, space vector pulse width modulation, Kalamn filter.*

I. Introduction

The first and most popular vector control (VC) method was field oriented control (FOC) [1]. This vector control technique is one of the most important closed loop techniques for AC machines in variable speed applications. Using this control technique, the torque and flux can be decoupled so each can be controlled separately. In recent years mid 80's a direct torque control (DTC) has become an alternative to the well known vector control. It was introduced in Japan by Takahashi (1984) and also in Germany by Depenbrock (1985)[2] [3]. The main feature of DTC is the high performance achieved with a simpler structure and control diagram. But it presents a problem of field linkage and torque ripple. In order to solve this problem the conventional DTC is combined with space vector pulse width modulation (SVPWM). This control theory has achieved great success in the control of PMSM. That has becoming a hotspot for resolving... Both methods, DTC, achieve decoupled control of torque and flux and it was first implemented in the control of induction motor drives [4][5][6].. It was recently also applied to different machines such as PMSM drives and Brushless DC motors [7-11].

In this paper, a DTC design for PMSM speed control is proposed. The EKF is adopted to estimate the speed and the torque. That is robust to the parameter uncertainties and load torque disturbance. The rest of this paper is organized as follows. Section 2 reviews the PMSM modeling. Section 3 shows the principle of the DTC and SVM-DTC. Section 4 resumes the EKF theory. Section 5gives some simulation results. Finally, the conclusion is drawn in section 6.

II. Mathematical Model of the PMSM

The PMSM stator field of is obtained from the following equation:

$$\overline{V}_s = R_s \overline{I}_s + \frac{d\overline{\phi}_s}{dt} \quad (1)$$

We obtain

$$\overline{\phi}_s = \overline{\phi}_{s0} + \int_0^t (\overline{V}_s - R_s \overline{I}_s)\, dt \quad (2)$$

The voltage drop due to the stator resistance can be neglected (for high speeds) we find then:

$$\overline{\phi}_s \approx \overline{\phi}_{s0} + \int_0^t \overline{V}_s \,\mathrm{dt} \quad (3)$$

For one sampling period, the voltage vector applied to the PMSM remains constant, we can write then:

$$\overline{\phi}_s(K+1) \approx \overline{\phi}_{s0}(K) + \overline{V}_s T_e \quad (4)$$

Where still:

$$\Delta\,\overline{\phi}_s \approx \overline{V}_s T_e \quad (5)$$

With:

- $\overline{\phi}_s(\mathrm{K})$: the field stator vector of the current sampling
- $\overline{\phi}_s(\mathrm{K}+1)$ the field stator vector of following sampling
- $\Delta\,\overline{\phi}_s$ The variation of the stator field vector ($\overline{\phi}_s(\mathrm{K}+1) - \overline{\phi}_s(\mathrm{K})$);
- T_e: The sampling period

For one constant sampling period $\Delta\,\overline{\phi}_s$ is proportional to stator applied voltage vector of the MSAP [12-13]. Figure 1 shows the evolution of the stator field vector in the (α, β) plan.

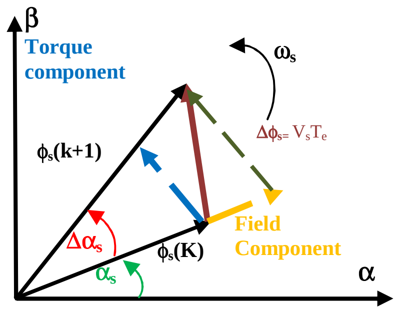

Fig.1 Stator field vector evolution in the (α,β) plan

The electromagnetic torque is proportional to the vector product between the vectors of stator field and rotor one according to the follow expression [7]:

$$T_e = k(\overline{\phi}_s + \overline{\phi}_f) = \frac{P}{L_q}\,|\overline{\phi}_s|\,|\overline{\phi}_f|\,sin\delta \quad (6)$$

Such as:

- $\overline{\phi}_s$ The stator field vector
- $\overline{\phi}_f$ The rotor field vector brought back to the stator;
- δ The angle between the stator field vectors and rotor one.

III. DTC AND DTC - SVPWM DEVELOPMENT

The basic concept of DTC is to control directly both the stator flux and electromagnetic torque of machine simultaneously by the selection of optimum inverter switching modes. The use of a switching table for voltage vector selection provides fast torque response, low inverter switching frequency and low harmonic losses without the complex field orientation by restricting the flux and torque errors within respective flux and torque hysteresis bands with the optimum selection being made[14]. The DTC controller consists of two hysteresis comparator (flux and torque) to select the switching voltage vector in order to maintain flux and torque between upper and lower limit as show in figure 2.

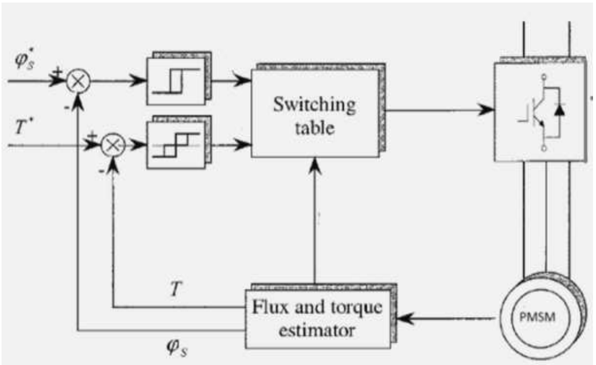

Fig.2 block diagramm of DTC for PMSM

The six different directions of *Vs* noted as V_i

i the combination of switches status of the inverter are given shown in figure 3 represents the choice of this vector selection.

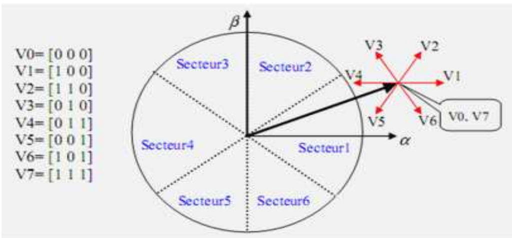

Fig. 3 Inverter output voltage vectors

The develop torque control of inverter fed PMSM is carried out by hysteresis control of magnitude stator flux and torque that selects one of the voltage vectors. The selection is made in order to maintain torque and flux error inside the hysteresis band in which the errors are indicated by ΔT_e and $\Delta\phi$ respectively.

ΔT_e: Torque error given by T_{eref} - T_e

$\Delta\phi$: Field error given by ϕ_{ref} - ϕ

Than we can write the different rules for $\Delta\emptyset$ and ΔT_e

$$\begin{array}{ll} if\ \Delta\emptyset > \varepsilon_\emptyset & \Rightarrow C_f = 1 \\ if\ \Delta\emptyset > \varepsilon_\emptyset\ \&\ \frac{d\,\Delta\emptyset}{dt} > 0 & \Rightarrow C_f = 0 \\ if\ \Delta\emptyset > \varepsilon_\emptyset\ \&\ \frac{d\,\Delta\emptyset}{dt} < 0 & \Rightarrow C_f = 1 \\ if\ \Delta\emptyset > \varepsilon_\emptyset & \Rightarrow C_f = 0 \end{array} \quad (7)$$

$$\begin{array}{ll} if\ \Delta T_e > \varepsilon_T & \Rightarrow C_T = 1 \\ if\ 0 \le \Delta T_e \le \varepsilon_T\ \&\ \frac{d\,\Delta T_e}{dt} > 0 & \Rightarrow C_T = 0 \\ if\ 0 \le \Delta T_e \le \varepsilon_T\ \&\ \frac{d\,\Delta T_e}{dt} < 0 & \Rightarrow C_T = 1 \\ if\ \Delta T_e < -\varepsilon_T & \Rightarrow C_T = -1 \\ if\ -\varepsilon_T \le \Delta T_e \le 0\ \&\ \frac{d\,\Delta T_e}{dt} > 0 & \Rightarrow C_T = 0 \\ if\ -\varepsilon_T \le \Delta T_e \le \varepsilon_T\ \&\ \frac{d\,\Delta T_e}{dt} > 0 & \Rightarrow C_T = -1 \end{array} \quad (8)$$

One selected switches state allowing determining the inverter operation sequences by using the table I, while basing on the field and torque errors, and according to the field vector position, it possible to determine the division of the complex plan in six angular sectors. For each sector given, the ordering sequence of the inverter switches which corresponds to the different states of $\Delta T$and $\Delta\emptyset$ according to the logic of the behavior of field and the torque with respect to the application of a stator voltage vector [6][8].

Table I

	Increase	***decrease***
$\emptyset$	V_{i-1}, V_i *and* V_{i+1}	V_{i-2}, V_{i+2} *and* V_{i+3}
T_e	V_{i+1} *and* V_{i+2}	V_{i-1} *and* V_{i-2}

While basing on the generalized table 1, one can draw up the table II of the sequences below summarizing the vector PWM, proposed by Takahashi, to control stator field and the torque of the PMSM.

TableII

	N	1	2	3	4	5	6
$C_\phi = 1$	$C_T = 1$	V_2	V_3	V_4	V_5	V_6	V_1
	$C_T = 0$	V_7	V_0	V_7	V_0	V_7	V_0
	$C_T = -1$	V_6	V_1	V_2	V_3	V_4	V_5
$C_\phi = 0$	$C_T = 1$	V_3	V_4	V_5	V_6	V_1	V_2
	$C_T = 0$	V_0	V_7	V_0	V_7	V_0	V_7
	$C_T = 1$	V_5	V_6	V_1	V_2	V_3	V_4

For such high power industrial applications, if in the DTC the hysteresis bands of the controllers become relatively wide, with the low inverter switching frequency, the resulting motor torque pulsations become high up to an undesired level. To overcome the above problems, some researchers have suggested, the DTC scheme using the space vector modulation (SVM) techniques [16-17]. In [5] and [18] control method has been discussed that allows constant switching frequency operation and uses two PI controller in order to generate the inverter reference voltage in the PMSM stator flux reference frame. In this control scheme, a PI speed controller is also used to produce the torque reference signal.

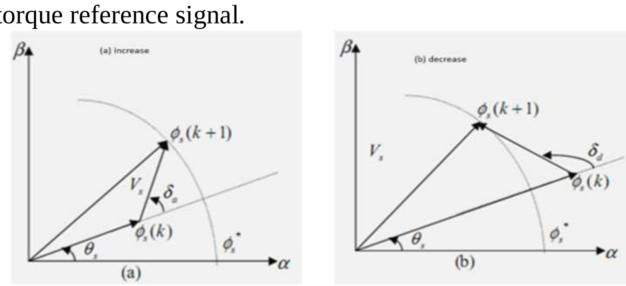

Fig.4 Angle and voltage vectors V_s determination

δ : Angle of the vector $\overline{V_s}$ in the (α,β) plan

θ_s: Field position

•If the field error is negative, one defines an angle δ_d between the vector $\overline{V_s}$ and the vector ϕ_s equal to 2 π/3. So the angle δ is given as: $\delta = \theta_s + \delta_d$

•If the field error is positive,δ_d between the vector $\overline{V}_s$ and the vector ϕ_s equal to $\pi/3$. So the angle δ is given as: $\delta = \theta_s + \delta_a$

The table III resume the different choice of δ

Table III

Field	0			1		
Torque	-1	0	1	-1	0	1
Angle δ	$\theta_s - \delta_a$	$\theta_s - \pi$	$\theta_s + \delta_a$	$\theta_s - \delta_a$	θ_s	$\theta_s + \delta_a$

IV. EXTENDED KALMAN FILTER

Now the extended Kalman filter observer will be applied to estimate the rotor speed which is feedback controlled by PI regulator. The EKF observer is based on the error of the stator currents generated from their measured and estimated values which must be converged toward zero via defined design.

The EKF has been described in many papers and is summarized in this section [19][20].

State equations for PMSM can be written as (31).

$$\dot{x} = g(x,u) + w$$

$$y = c.x + v$$

Here $x = \begin{bmatrix} I_d & I_q & \omega & \theta \end{bmatrix}, \; u = \begin{bmatrix} u_d & u_q \end{bmatrix}$

$$C = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix}, \; y = \begin{bmatrix} I_d \\ I_q \end{bmatrix}$$

w and v are random disturbances. In fact w is the process noise which stands for the errors of the parameters; v is the measurement noise which stands for the errors in the measurement and sample. The noise covariance matrixes are defined as follows:

$$Q = \mathrm{cov}(w) = E\left\{ww^T\right\}$$

$$R = \mathrm{cov}(v) = E\left\{vv^T\right\}$$

Kalman filter can be built by this follow derivation:

$$x(k+1) = f = \begin{cases} I_d(k) + \left(\frac{U_d}{L_d} - \frac{R.I_d}{L_d} + \omega \frac{L_q}{L_d} I_q \right) T_e \\ I_q(k) + \left(\frac{U_q}{L_q} - \frac{R.I_q}{L_q} - \omega \frac{L_d}{L_q} I_d - \omega \frac{\phi_f}{L_q} \right) T_e \\ \omega(k) \\ \theta(k) + \omega.T_e \end{cases}$$

Define matrix *F and H as*:

$$F = \frac{\partial f}{\partial x} = \begin{bmatrix} 1 - \frac{T_e}{\tau_d} & T_e \omega \frac{L_q}{L_d} & T_e \frac{L_q}{L_d} I_q & T_e \frac{u_d}{L_d} \\ -T_e \omega \frac{L_d}{L_q} & 1 - \frac{T_e}{\tau_q} & T_e \left(-\frac{L_d}{L_q} I_d - \frac{\phi_f}{L_q} \right) & T_e \frac{u_q}{L_q} \\ 0 & 0 & 1 & 0 \\ 0 & 0 & T_e & 1 \end{bmatrix}$$

$$h = \begin{bmatrix} i_d \\ i_q \end{bmatrix} ; \quad H = \frac{\partial h}{\partial x} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix}$$

$\tau_d = \frac{L_d}{R}, \tau_q = \frac{L_q}{R}$ are stator constants.

Extended Kalman filter can be realized by iteration as follows:

After deciding how to initialize the covariance matrices, the next step is prediction of the state vector at sampling time *(k+1)* from the input *u (k)*, state vector at previous sampling time, $x_{k|k}$:

$$\hat{x}_{k+1|k} = \hat{x}_{k|k} + \dot{x}.T_e$$

The notation $xk|k$ means that it is a predicted value at the *(k+1)th* instant, and it is based on measurements up to *k-th* instant. In the following step of the recursive EKF computation, covariance matrix of prediction is computed.

$$P_{k+1|k} = F_{k|k}.P_{k|k}.F_{k|k}^T + Q_k$$

In the second stage which is the filtering stage, the next estimated states $\hat{x}_{k+1}$, are obtained from the predicted estimates x_{k+1} by adding a correction term $K(y - \hat{y})$ to the predicted value. This correction term is a weighted difference between the actual output vector *(y)* and the predicted output vector $(\hat{y})$, where K is the Kalman gain.

Next step is the computation of the Kalman filter gain matrix as:

$$K_k = P_{k+1|k} C^T . \left(C.P_{k+1|k} C^T + R_k \right)^{-1}$$

The predicted state-vector is added to the innovation term multiplied by Kalman gain to compute state-estimation vector. The state-vector estimation (filtering) at time (k) is determined as:

$$\hat{x}_{k+1|k+1} = \hat{x}_{k+1|k} + K_{k+1}\left(y_{k+1} . C.\hat{x}_{k+1|k} \right)$$

Where $y_{k+1} = C.x_{k+1|k}$

Here we define the estimation covariance computation as:

$$P_{k+1|k+1} = [I - K_{k+1}.C].P_{k+1|k}$$

This observer can be constituted from the PMSM model.

The choice of the initial values for the matrices *P*, *Q* and *R* is very important for the EKF. Generally, $P_{0|0}$ determines the initial transient characteristics of the filter, but has little influence on the initial tuning procedure of the EKF. Since the algorithm does not require initial rotor position information and the motor is assumed to start from the standstill, the initial state vector $x_{0|0}$ is considered to be a null vector. Based on the discussion given by [21] after the trial-and-error procedure, initial values for the states and matrices *P*, *Q* and *R* were selected as follow:

$$\hat{x}_{0/0} = \begin{bmatrix} i_d \\ i_q \\ \omega \\ \theta \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} ; \; P_{0/0} = \begin{bmatrix} 0.1 & 0 & 0 & 0 \\ 0 & 0.1 & 0 & 0 \\ 0 & 0 & 100 & 0 \\ 0 & 0 & 0 & 10 \end{bmatrix}$$

$$Q = \begin{bmatrix} 100 & 0 & 0 & 0 \\ 0 & 100 & 0 & 0 \\ 0 & 0 & 1000 & 0 \\ 0 & 0 & 0 & 0.01 \end{bmatrix} ; \; R = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$$

Based on the EKF and the DTC, the proposed sensorless speed control of a PMSM is shown in Fig1. As depicted the proposed PI regulator compares the reference ω^* with speed ω_{est}, calculated with the estimated speed given by Kalman filter and it is delivers as output the torque reference T_{emref} . The flux is controlled to follow its reference .

V. SIMULATION RESULTS

In order to prove the effectiveness and the feasibility of the proposed SVM-DTC associated to EKF , the simulation module is built in MATLAB/SIMULINK®. The specifications for the used PMSM are listed in table IV.

Figure 5 show the torque dynamic performance, the locus of the stator flux and he stator current response.

Figure 6 gives the comparison between the two stator flux locus with classical DTC and SVM-DTC

Figure 7 shows the speed tracking controller operated in a critical situation of benchmark commands rapidly changes as 50 – 100 – 0 (rad/s) . The observed speed converges with reference speed in very short time with a negligible overshoot and no steady state. It can be observed with the Zoomed response

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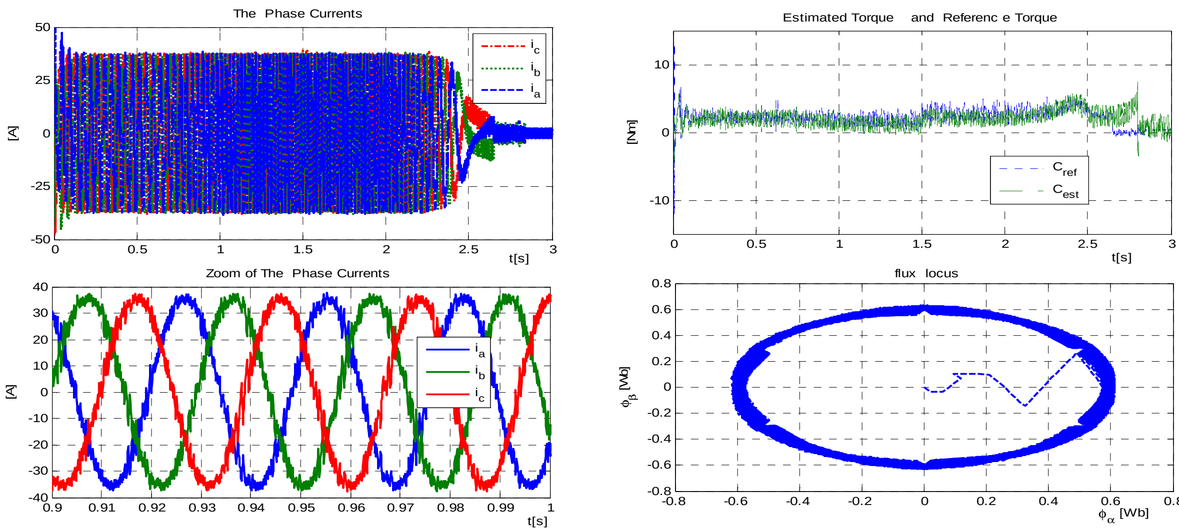

Fig.5 Simulation results without speed sensor

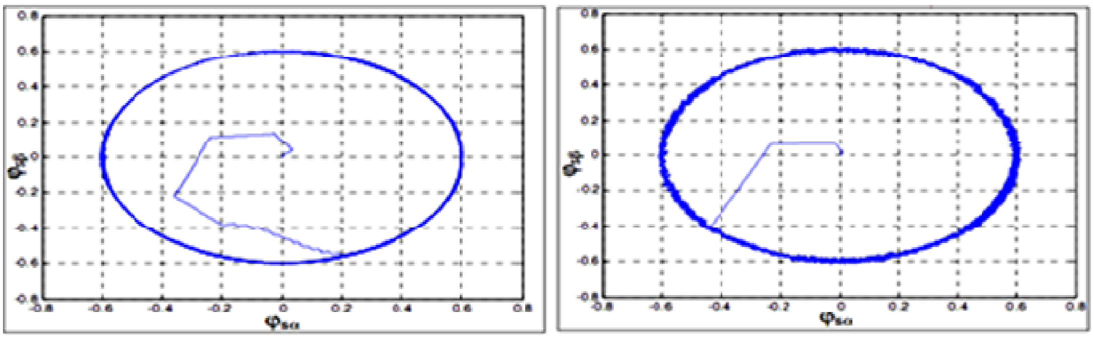

Fig.6 Ccomparison of the stator flux locus with DTC (right) and SVM-DTC (left)

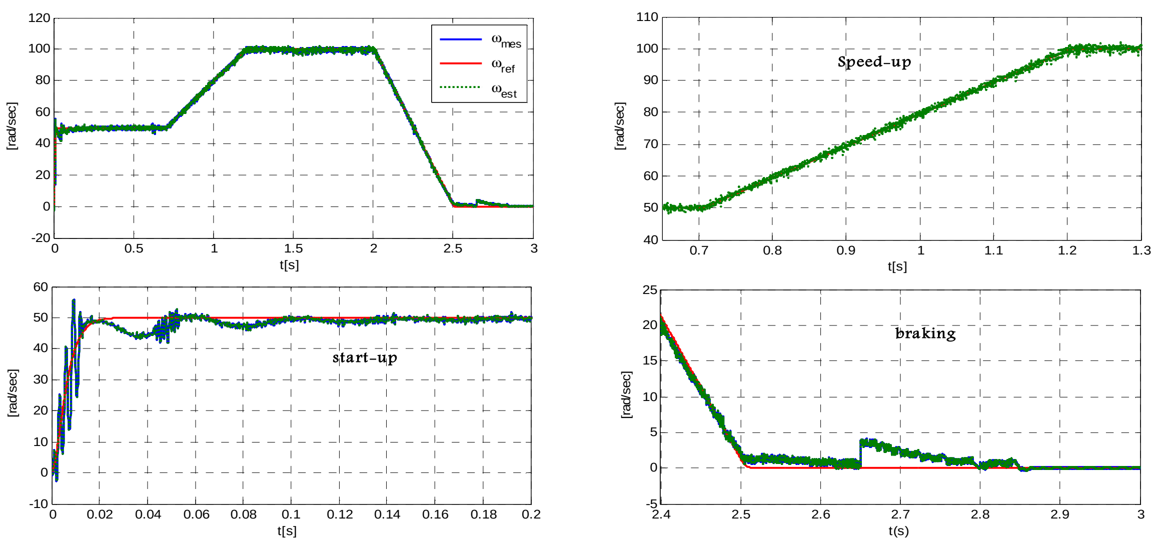

Fig.7 Speed simulation results under SVM-DTC -EKF

VI. CONCLUSION

To solve the problems of large flux and torque ripples and inconstant switching frequency in direct torque control for permanent magnet synchronous motors and to eliminate speed sensor since it deteriorates system reliability and increases the cost of system, a novel speed sensorless DTC method based on space vector modulation and EKF is proposed in this paper.

The torque ripple for this SVM-DTC is significantly improved and switching frequency is maintained constant. This study has successfully demonstrated the design of the EKF control for the speed control of a PMSM . The control laws were derived based on the motor model. Numerical simulations have been carried out showing the advantages of the SVM-DTC method with respect to the conventional DTC. The effectiveness and robustness at tracking a reference speed under critical situation of benchmark commands rapidly changes.

Table IV

Rated voltage	208V
Pole pair number *P*	3
d-axis inductance, L_d	66 *m*H
q-axis inductance, L_q	58 *m*H
Stator resistance, R_s	1.4Ω
Motor inertia, *J*	0.00176kgm^2
Friction coefficient, *B*	3.88. 10^{-4}*Nm/rad/s*
Magnetic flux constant ϕ_f	0. 1546 *Wb*

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