Space Vector Double Frame Field Oriented Control of Six Phase Induction Motors

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Abstract: – In this paper a new rotor field orientation control of six phase induction machine is proposed. The machine has two three phase windings spatially shifted by 30 electrical degrees. The dynamic model of the machine is based on the three two dimensional orthogonal subspaces. The proposed method uses a simple estimator for identifying rotor flux and its orientation. To eliminate the extra harmonic currents of stator windings that produce losses, space vector PWM technique and double synchronous frame current controller is employed. The simulation results illustrate the validity and efficiency of the proposed method.

Key-Words: - Six-phase induction machine; Vector control; Space voltage modulation.

Nomenclature

R_s	stator resistance matrice
R_r	rotor resistance matrice
r_s	stator resistance
r_r	rotor resistance
L_{ss}	stator inductance matrice
L_{rr}	rotor inductance matrice
L_{sr}	*mutual ind*uctance matrice
L_m	mutual inductance
L_s	stator inductance
L_r	rotor inductance
τ_r	rotor time constant
τ_s	stator time constant
λ_r	rotor flux linkage
i_s	stator phase current
v_s	stator phase voltage
i_r	rotor phase current
v_r	rotor phase voltage
I_s	stator current matrice
V_s	stator voltage matrice
I_r	rotor current matrice
V_r	rotor voltage matrice
e	back-emf
P	number of pole pairs
J	inertia coefficient
Ω	rotor angular speed
Ω_s	angular speed of δ–γ frame
ω_s	electric angular frequency
θ	rotor position
ϑ	position of δ–γ frame
φ	rotor position error ($\varphi = \vartheta - \theta$)
T_m	electromagnetic torque
T_L	load torque
T_s	sampling period
T_k	*k*th interval time

Subscripts

a, b ,c	phases
s	stator
r	rotor
α, β	stationary coordinates
d, *q*	synchronous coordinates
δ, γ	control coordinates

Superscripts

^	estimated value
~	error value

1 Introduction

During past several years, multiphase motor drives have been proposed by different authors. In the multi phase drive systems, the electric machine has more than three phases in the stator and the same numbers of inverter legs are in the inverter side. So the current per phase in the machine and inverter is reduced. The most common multi phase machine drive structure is the six phase induction machine (SPIM), which has two sets of three-phase windings, spatially phase shifted by 30 electrical degrees with double neutral (Fig. 1). In this structure, each set of three phase windings is excited by a three leg (three phase) inverter, so the total rating of system is doubled, the torque pulsations will be smoothed, the rotor harmonic losses as well as the harmonics content of the dc link current will be reduced and the loss of one machine phase does not prevent the machine working, so improving the system reliability.

When a SPIM is fed by a six step voltage source inverter, large harmonic currents that do not contribute to the air gap flux, have been observed [1]. These harmonics generate additional losses in the machine and increase the size and cost of the machine drive system [2], [3]. The researchers are focused on the suppression of these stator current harmonics. In this way, some approaches concern the machine structure design [4]-[6]. The other approaches concern applying a proper PWM technique for reducing the harmonic voltages of the motor. In this case, without any modification on the machine and its structure, and only by proper control strategy of inverter, they try to reduce undesired current components.

Some researches concerning control of unbalanced current sharing between the two three-phase windings sets. This unbalanced current is due to the small system asymmetries. This problem has been discussed in [7] and [8] and has been solved in [7] by closing the control loops in phase coordinates and in [8] by introducing a double synchronous frame current controller.

In this paper a new configuration of rotor field orientation control of SPIM with space vector PWM technique and double d-q synchronous frame current controller is developed. Before in [9] we have proposed an indirect field orientation control of SPIM. In this paper we have developed double d-q synchronous frame current controller based on the technique proposed in [8] in the indirect field orientation control of SPIM with space vector PWM.

This paper is organized in seven sections. The model of SPIM is described in the next section. The third section presents the vector control of SPIM. In section four, PWM of SPIM based on the space vector decomposition method is discussed. The double d-q synchronous frame current controller is described in the fifth section. The application of the proposed control method to the SPIM is presented in the sixth section. Finally some conclusions and perspectives will be discussed in the seventh section.

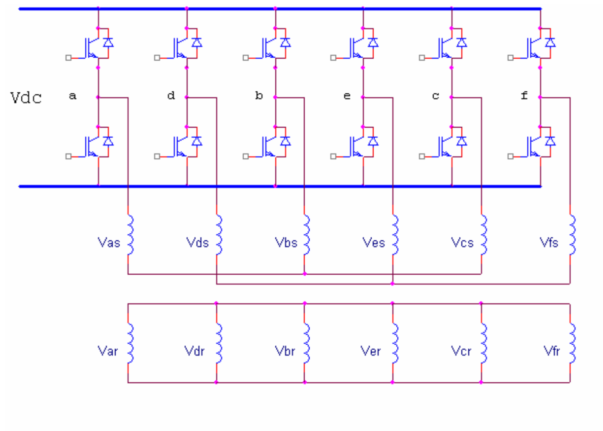


Fig. 1. Circuit diagram of SPIM and its inverter.

2 Model of SPIM

Model of SPIM is derived based on the following assumptions:

- the machine windings are distributed sinusoidaly;
- the mutual leakage inductances are negligible;
- the magnetic saturation and the core losses are neglected.

The voltage equations of the SPIM are as follow. For the stator circuit we can write:

$$[V_s]=[R_s][I_s]+p([L_{ss}][I_s]+[L_{sr}][I_r]) \tag{1}$$

and for the rotor circuit we have:

$$[V_r]=[R_r][I_r]+p([L_{rr}][I_r]+[L_{rs}][I_s]) \tag{2}$$

in which the current and voltage vectors are:

$$[V_s]=\begin{bmatrix} v_{as}\\ v_{bs}\\ v_{cs}\\ v_{ds}\\ v_{es}\\ v_{fs}\end{bmatrix},\ [I_s]=\begin{bmatrix} i_{as}\\ i_{bs}\\ i_{cs}\\ i_{ds}\\ i_{es}\\ i_{fs}\end{bmatrix},\ [V_r]=\begin{bmatrix}0\\0\\0\\0\\0\\0\end{bmatrix},[I_r]=\begin{bmatrix} i_{ar}\\ i_{br}\\ i_{cr}\\ i_{dr}\\ i_{er}\\ i_{fr}\end{bmatrix}$$

and: $p=\frac{d}{dt}$

As can be seen, the SPIM is a six dimensional system. It is shown in [8]-[11] that it can be decomposed into three two-dimensional orthogonal subspaces, (α, β), (z1, z2), (z3, z4), by the following transformation:

$$[T]=\frac{1}{\sqrt{3}}\begin{bmatrix} 1 & \cos(\gamma) & -\frac{1}{2} & \cos(2\pi/3+\gamma) & -\frac{1}{2} & \cos(4\pi/3+\gamma) \\ 0 & \sin(\gamma) & \frac{\sqrt{3}}{2} & \sin(2\pi/3+\gamma) & -\frac{\sqrt{3}}{2} & \sin(4\pi/3+\gamma) \\ 1 & \cos(\pi-\gamma) & -\frac{1}{2} & \cos(\pi/3-\gamma) & -\frac{1}{2} & \cos(5\pi/3-\gamma) \\ 0 & \sin(\pi-\gamma) & -\frac{\sqrt{3}}{2} & \sin(\pi/3-\gamma) & \frac{\sqrt{3}}{2} & \sin(5\pi/3-\gamma) \\ 1 & 0 & 1 & 0 & 1 & 0 \\ 0 & 1 & 0 & 1 & 0 & 1 \end{bmatrix} \tag{3}$$

In this matrix, γ is the electrical angle between two three-phase windings; that is 30 electrical degrees in this paper. By applying the transformation (3) to the voltage equations (1) and (2), we can obtain decoupled model for SPIM:

$$[T].[V_s]=[T].[R_s].[T^{-1}].[T].[I_s]+p([T].[L_{ss}].[T^{-1}].[T].[I_s]+[T].[L_{sr}].[T^{-1}].[T].[I_r]) \tag{4}$$

$$[T].[V_r]=[T].[R_r].[T^{-1}].[T].[I_r]+p([T].[L_{rr}].[T^{-1}].[T].[I_r]+[T].[L_{rs}].[T^{-1}].[T].[I_s]) \tag{5}$$

where:

$$[T].[v_s]=\begin{bmatrix} v_{ds} \\ v_{qs} \\ v_{z1s} \\ v_{z2s} \\ v_{z3s} \\ v_{z4s} \end{bmatrix}, \qquad [T].[I_s]=\begin{bmatrix} i_{ds} \\ i_{qs} \\ i_{z1s} \\ i_{z2s} \\ i_{z3s} \\ i_{z4s} \end{bmatrix}$$

$$[T].[v_r]=\begin{bmatrix} v_{dr} \\ v_{qr} \\ v_{z1r} \\ v_{z2r} \\ v_{z3r} \\ v_{z4r} \end{bmatrix}=\begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix}, \quad [T].[I_r]=\begin{bmatrix} i_{dr} \\ i_{qr} \\ i_{z1r} \\ i_{z2r} \\ i_{z3r} \\ i_{z4r} \end{bmatrix}$$

We can obtain new system equation of SPIM as follows:

Machine model in α-β subspace

The stator and rotor voltage equations are:

$$\begin{bmatrix} v_{s\alpha} \\ v_{s\beta} \\ 0 \\ 0 \end{bmatrix}=\begin{bmatrix} r_s+L_s p & 0 & Mp & 0 \\ 0 & r_s+L_s p & 0 & Mp \\ Mp & \omega_r M & r_r+L_r p & \omega_r L_r \\ -\omega_r M & Mp & -\omega_r L_r & r_r+L_r p \end{bmatrix}.\begin{bmatrix} i_{s\alpha} \\ i_{s\beta} \\ i_{r\alpha} \\ i_{r\beta} \end{bmatrix} \tag{6}$$

with:
$$L_s=L_{ls}+M$$
$$L_r=L_{lr}+M$$
$$M=3L_{ms}$$

This model is similar to the three phase machine model in the stationary reference frame.

Machine model in z1-z2 subspace

$$\begin{bmatrix} v_{sz1} \\ v_{sz2} \end{bmatrix}=\begin{bmatrix} r_s+L_{ls}p & 0 \\ 0 & r_s+L_{ls}p \end{bmatrix}.\begin{bmatrix} i_{sz1} \\ i_{sz2} \end{bmatrix} \tag{7}$$

$$\begin{bmatrix} 0 \\ 0 \end{bmatrix}=\begin{bmatrix} r_r+L_{lr}p & 0 \\ 0 & r_r+L_{lr}p \end{bmatrix}.\begin{bmatrix} i_{rz1} \\ i_{rz2} \end{bmatrix} \tag{8}$$

Machine model in z3-z4 subspace

$$\begin{bmatrix} v_{sz3} \\ v_{sz4} \end{bmatrix}=\begin{bmatrix} r_s+L_{ls}p & 0 \\ 0 & r_s+L_{ls}p \end{bmatrix}.\begin{bmatrix} i_{sz3} \\ i_{sz4} \end{bmatrix} \tag{9}$$

$$\begin{bmatrix} 0 \\ 0 \end{bmatrix}=\begin{bmatrix} r_r+L_{lr}p & 0 \\ 0 & r_r+L_{lr}p \end{bmatrix}.\begin{bmatrix} i_{rz3} \\ i_{rz4} \end{bmatrix} \tag{10}$$

As it can be seen from these three subsystems, the electromechanical energy conversion takes place in the α-β subsystem, and the other subsystems do not contribute in the energy conversion. The z1-z2 and z3-z4 subsystems are only producing losses, so they should be controlled to be as small as possible. It can be concluded that analyzing the SPIM is performed by help of the α-β subspace that is similar to the dynamic model of a three-phase induction machine.

3 Vector Control of SPIM

By using park's transformation, *d-q* reference frame model for portion of the SPIM that contributes in the energy conversion can be obtained as follows:

$$\begin{cases} \frac{d}{dt} i_{sd} = \frac{-1}{\sigma \tau'_s} i_{sd} + \omega_s i_{sq} + \frac{\mu}{\sigma \tau_r} \lambda_{rd} + \frac{\mu}{\sigma} P\Omega \lambda_{rq} + \frac{1}{\sigma L_s} v_{sd} \\ \frac{d}{dt} i_{sq} = \frac{-1}{\sigma \tau'_s} i_{sq} - \omega_s i_{sd} + \frac{\mu}{\sigma \tau_r} \lambda_{rq} - \frac{\mu}{\sigma} P\Omega \lambda_{rd} + \frac{1}{\sigma L_s} v_{sq} \\ \frac{d}{dt} \lambda_{rd} = \frac{-1}{\tau_r} \lambda_{rd} + \frac{M}{\tau_r} i_{sd} + (\omega_s - P\Omega) \lambda_{rq} \\ \frac{d}{dt} \lambda_{rq} = \frac{-1}{\tau_r} \lambda_{rq} + \frac{M}{\tau_r} i_{sq} - (\omega_s - P\Omega) \lambda_{rd} \end{cases} \tag{11}$$

where v_{sd}, v_{sq}, i_{sd} and i_{sq} are *d-q* components of stator voltage and current vectors respectively that are obtained from park's transformation of α-β components; λ_{rd} and λ_{rq} are the rotor flux *d-q* components and Ω is the rotor angular speed. The machine parameters are r_s, r_r, L_r, M, L_s and P, with:

$$\sigma = 1 - \frac{M^2}{L_s L_r}, \quad \mu = \frac{M}{L_s L_r},$$

$$\tau'_s = \frac{L_s}{r_{s\,eq}}, \quad \tau_r = \frac{L_r}{r_r}, \quad r_{s\,eq} = r_s + \frac{M^2}{L_r^2} r_r$$

The mechanical equation is the following:

$$J \frac{d}{dt} \Omega = T_m - T_L \tag{12}$$

and:

$$T_{\mathrm{m}} = \frac{PM}{L_r} \left(\lambda_{rd} i_{sq} - \lambda_{rq} i_{sd} \right) \tag{13}$$

is the torque generated by the motor. T_L is the load torque supposed to be unknown.

According to model (11), if λ_{rq} tends to be zero, the rotor flux becomes independent from i_{sq} while the motor torque will be proportional to i_{sq}. It is the objective of the vector control. The only degree of freedom is the angular speed of *d-q* frame ω_s which must be used to regulate λ_{rq} to zero. In order to do this and according to (11), the stator voltage angular frequency ω_s is determined by the following vector control law:

$$\omega_s = P\Omega_s = P\Omega + \frac{M}{\tau_r} \frac{i_{sq}}{\lambda_r} \tag{14}$$

It can be easily shown that this vector control law guarantees the regulation of λ_{rq} to zero if the motor parameters are well known. Replacing (14) to (11), we can write the following equations that describe the dynamic behavior of vector-controlled SPIM:

$$\frac{d}{dt} i_{sd} = \frac{-1}{\sigma \tau'_s} i_{sd} + \omega_s i_{sq} + \frac{M}{\sigma L_s L_r \tau_r} \lambda_r + \frac{1}{\sigma L_s} v_{sd} \tag{15a}$$

$$\frac{d}{dt} i_{sq} = \frac{-1}{\sigma \tau'_s} i_{sq} - \omega_s i_{sd} - \frac{PM}{\sigma L_s L_r} \lambda_r \Omega + \frac{1}{\sigma L_s} v_{sq} \tag{15b}$$

$$\frac{d}{dt} \lambda_r = \frac{-1}{\tau_r} \lambda_r + \frac{M_s}{\tau_r} i_{sd} \tag{15c}$$

$$\frac{d}{dt} \Omega = \frac{PM}{J L_r} \lambda_r i_{sq} - \frac{1}{J} T_L \tag{15d}$$

As can be seen from (15), the rotor flux is independent from the load torque and i_{sq}, and the load torque is proportional to i_{sq} and λ_r.

4 Space Vector PWM

This method of modulation is used in modern control of IM. Inverter control by PWM permits to apply near sinusoidal voltages to the motor and also reduces torque pulsations. By combinational analysis of all states of 12 inverter switches, a total of 64 switching modes can be obtained. By applying the transformation (3), 64 voltage vectors are projected on the α-β, z1-z2 and z3-z4 subspaces. As mentioned before, we have used double neutral topology of SPIM as illustrated in Fig 1. In this case, the projection of the current vectors on the z3-z4 subspace is zero and the space vector PWM is performed only on the α-β and z1-z2 subspaces.

Fig 2 shows voltage vectors in the α-β and z1-z2 subspaces for SPIM. The decimal numbers in the figure, show switching states of the inverter. By converting each decimal number to a six digit binary number, the 1's indicate, on state of the upper switch in the corresponding arm of the inverter. The most significant bit (MSB) of the number represents the switching state of phase *a*, the second MSB for phase *d*, the third for phase *b*, and so on. We selected the 12 biggest vectors in the α-β subspace to obtain the highest amplitude of torque (Fig 3). The respective voltages in the z1-z2 subspace will be the smallest amplitude vectors. Thus we obtain maximum voltage output on the α-β subspace and at the same time, the smallest voltages on the z1-z2 subspace, so the harmonics and losses due to this subspace will be minimized.

The space vector PWM method of this research, according to the position of the reference voltage vector $V^*_{\alpha\beta}$, uses four adjacent voltage vectors and a zero voltage vector, all on the α-β subspace [10].

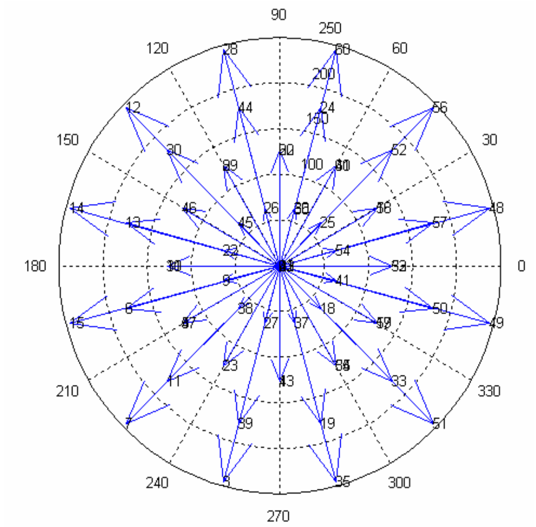


Fig. 2-a. Voltage vectors in the α-β subspace.

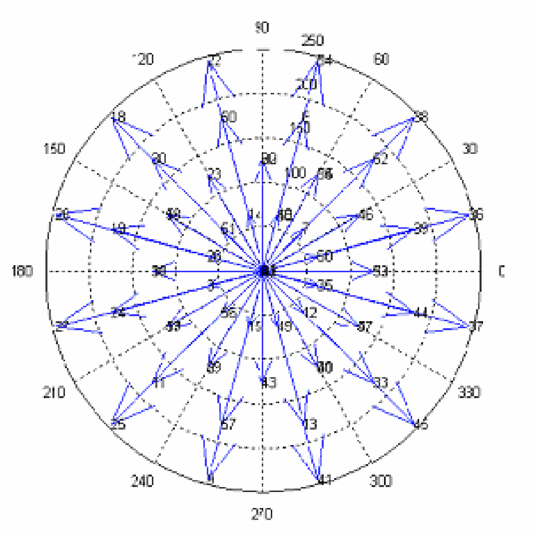


Fig. 2-b. Voltage vectors in the z1-z2 subspace.

Fig. 2. Voltage vectors of SPIM

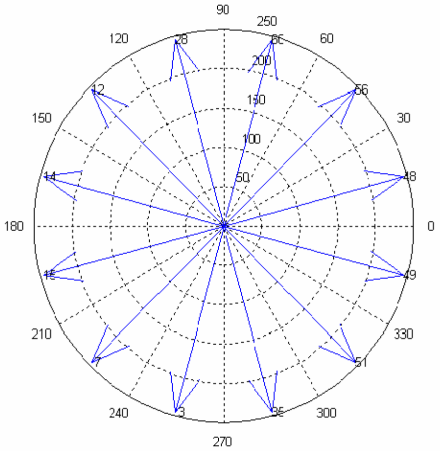


Fig. 3. Selected Voltage vectors in the α-β subspace.

The time interval corresponding to each voltage vector is calculated by the following set of system equation:

$$\begin{cases} T_1V_\alpha^1 + T_2V_\alpha^2 + T_3V_\alpha^3 + T_4V_\alpha^4 + T_5V_\alpha^5 = V_\alpha^* T_s \\ T_1V_\beta^1 + T_2V_\beta^2 + T_3V_\beta^3 + T_4V_\beta^4 + T_5V_\beta^5 = V_\beta^* T_s \\ T_1V_{zi}^1 + T_2V_{zi}^2 + T_3V_{zi}^3 + T_4V_{zi}^4 + T_5V_{zi}^5 = 0 \\ T_1V_{z2}^1 + T_2V_{z2}^2 + T_3V_{z2}^3 + T_4V_{z2}^4 + T_5V_{z2}^5 = 0 \\ T_1 + T_2 + T_3 + T_4 + T_5 = T_s \end{cases} \quad (16)$$

In this system of equations:

V_α^k : is projection of the *k*th voltage vector on the α-axis;

V_β^k : is projection of the *k*th voltage vector on the β-axis;

5 Double Frame Current Controller

Due to the small asymmetries in the stator windings and supply voltages, and depending to different operational conditions, the two stator windings sets have different currents. For compensating this asymmetry, according to [8], a double d-q synchronous frame current control can be used (Fig.4).

In this current control configuration, the currents of the two stator windings can be independently controlled, so the drive asymmetry will be compensated. This current control strategy is based on the two sets of d-q current components; each set is obtained from one stator

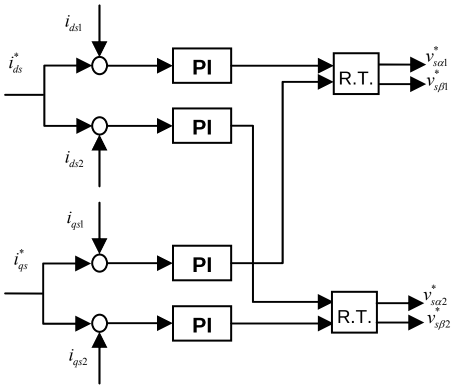


Fig. 4. Double *d*-*q* synchronous frame current control

windings sets (one from phases *abc*, and the other from phases *def*). By using Clarke transformation for every set of stator currents (17), and then applying the rotational transformation (18), the d-q current components of every stator windings sets will be produced. It should be noted that phase shift between the two sets of stator windings, must be applied in their rotational transformations.

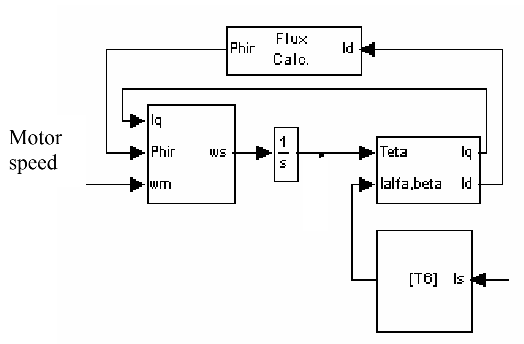
$$\begin{bmatrix} i_{\alpha 1} \\ i_{\beta 1} \end{bmatrix} = \sqrt{\tfrac{2}{3}} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \tag{17}$$

$$\begin{bmatrix} i_{d1} \\ i_{q1} \end{bmatrix} = \begin{bmatrix} Cos\theta & Sin\theta \\ -Sin\theta & Cos\theta \end{bmatrix} \begin{bmatrix} i_{\alpha 1} \\ i_{\beta 1} \end{bmatrix} \tag{18}$$

6 Simulation Results

In order to verify the efficiency of the proposed method in field orientation control of SPIM, we have developed a program based on the proposed algorithm. Equations (14) and (15c) have been used to estimate the orientation of the rotor flux and its amplitude, respectively (Fig. 5). We have used PI current and speed regulators. It should be noted that the source voltage is not measured and the stator current and angular speed are the only measures used in this control algorithm. The parameters of simulated motor are given in the Table 1. The overall system block diagram is shown in Fig. 6.

Fig. 5. Estimator of rotor flux orientation



shown in table 1. The overall system block diagram is shown in Fig. 6.

Fig. 7 illustrates the simulation results of the proposed method for a startup and speed inversion test. As can be seen, the speed follows the reference speed very well. In order to study the disturbance rejection using the proposed method, a new test is performed when the load varies at *t*=0.5 *sec*. Fig 8 shows the simulation results in this case. It can be seen that the speed is controlled very well and it follows its reference. The machine angular speed is well controlled.

As can be seen from these results, orientation in all cases is done well, and the harmonic currents are near zeros.

Table 1: Experimental Set up Parameters

VSI DC source voltage	600 V
no. of poles	8
Mutual inductance	51.3 mH
Stator resistance	2.34 Ω
Stator leakage inductance	6.7 mH
Rotor resistance	1.17 Ω
Rotor leakage inductance	6.7 mH
Inertia coefficient	0.03 Kg.m

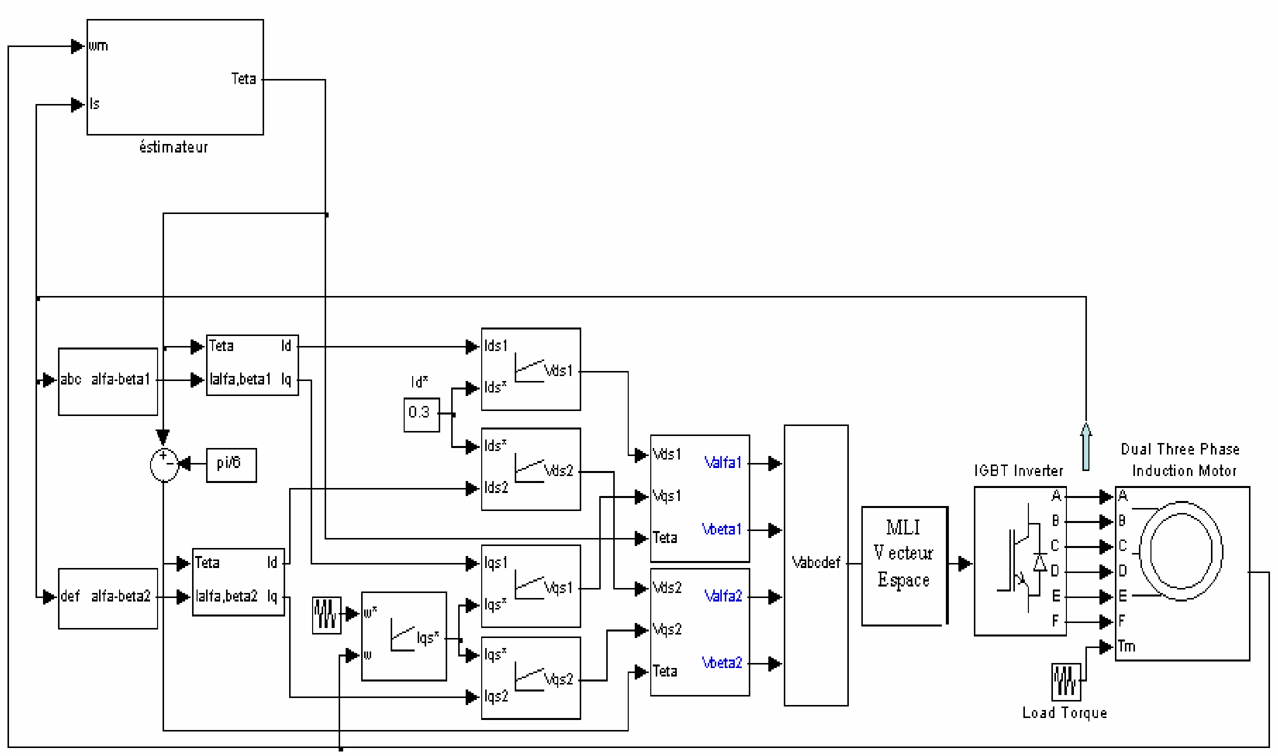


Fig. 6. block diagram of field orientation control of SPIM

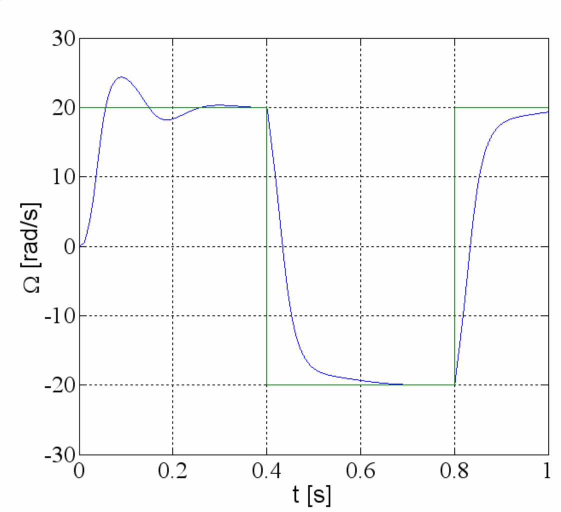


Fig. 7a: real speed.

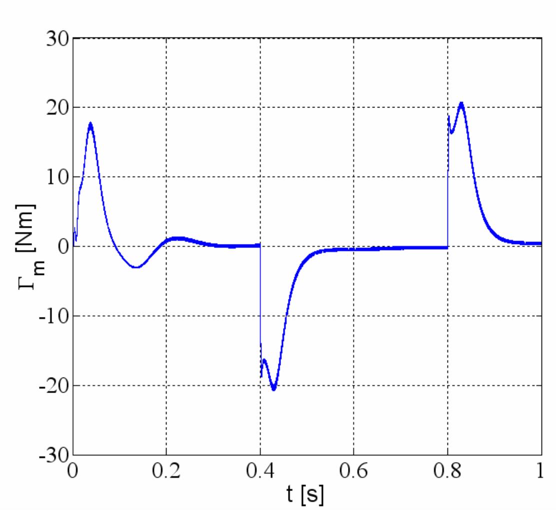


Fig. 7b: electromagnetic torque.

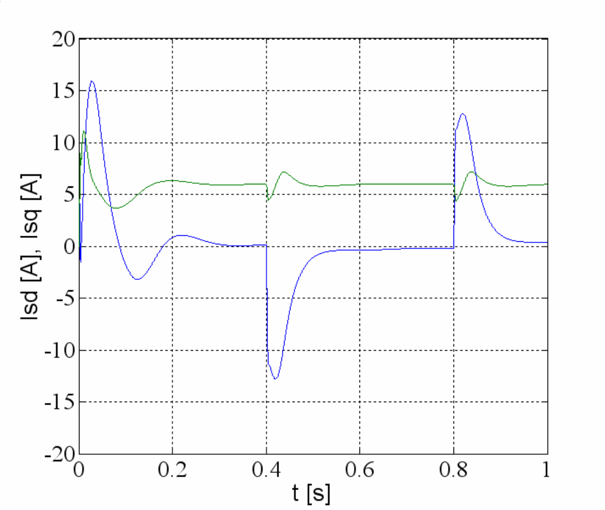


Fig. 7c: real direct and quadratic currents.

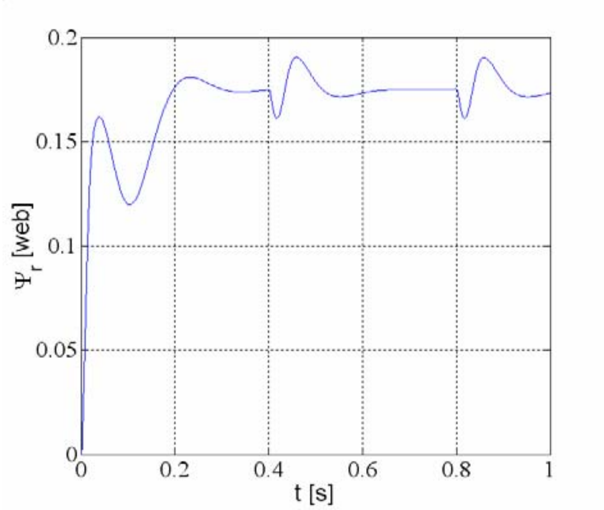


Fig. 7d: rotor flux estimation.

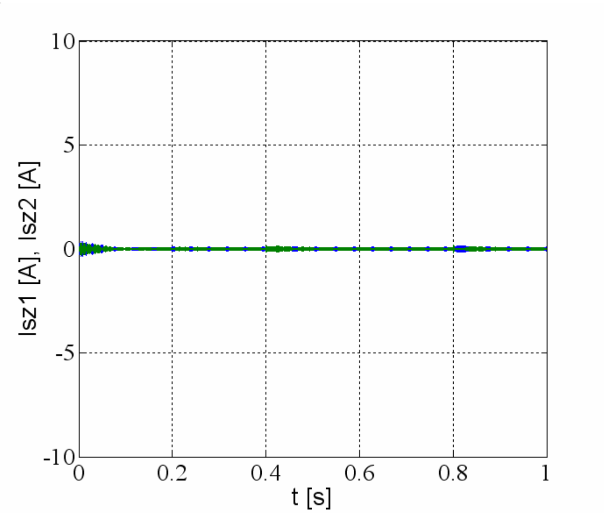


Fig. 7e: harmonic currents i_{z1} and i_{z2}

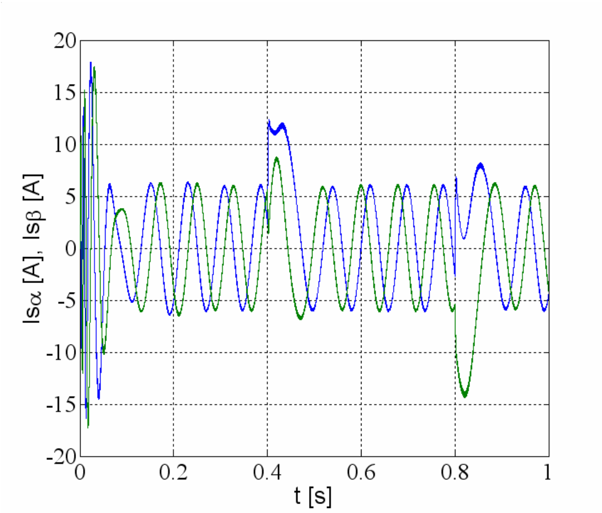


Fig. 7f: phase currents $i_{s\alpha}$ and $i_{s\beta}$

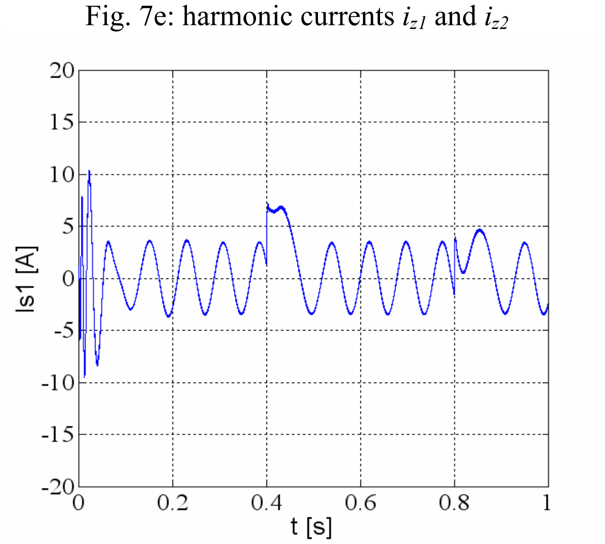


Fig. 7g: phase current i_{sa1}

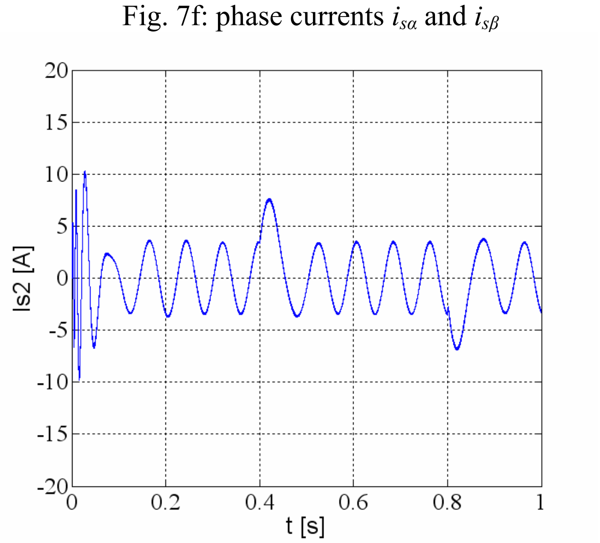


Fig. 7h: phase current i_{sa2}

Fig. 7. Simulation results for a startup and speed inversion test, Ω_{ref}=20 rad/sec.

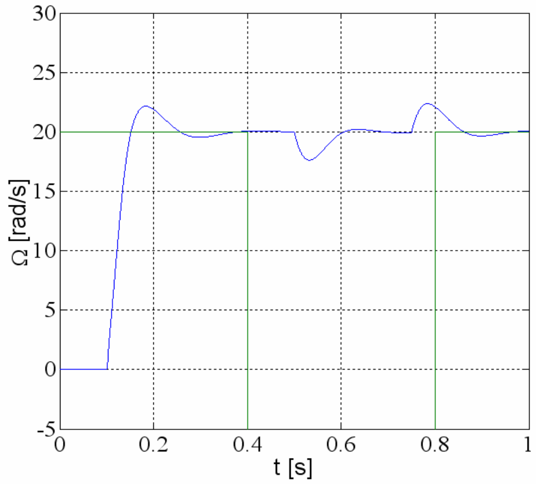


Fig. 8a: real speed.

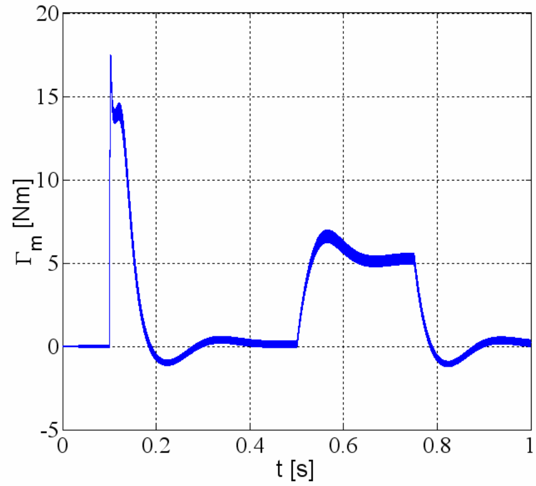


Fig. 8b: electromagnetic torque.

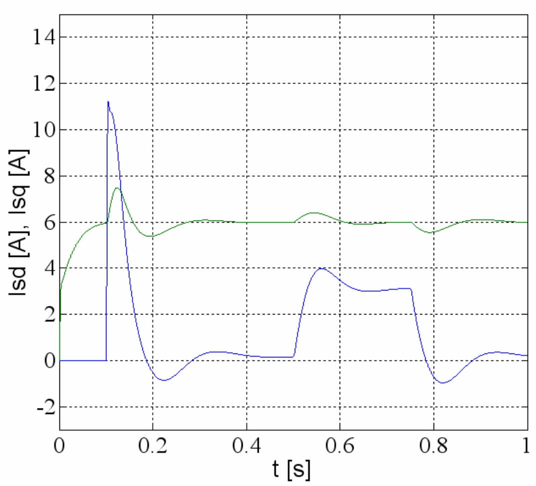


Fig. 8c: real direct and quadratic currents.

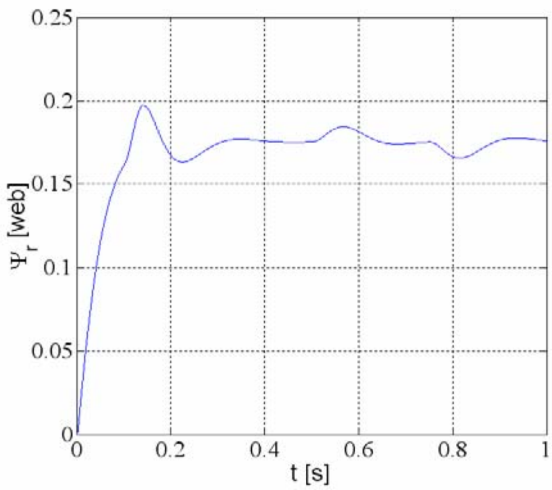


Fig. 8d: rotor flux estimation.

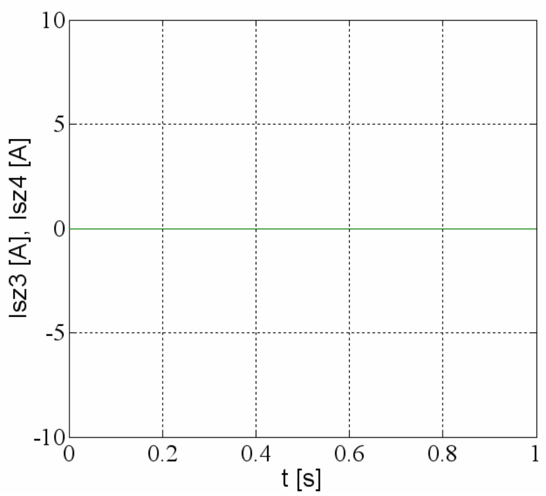


Fig. 8e: harmonic currents i_{z3} and i_{z4}

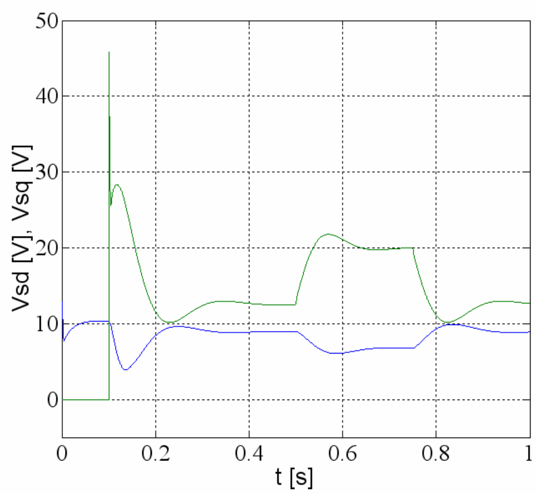


Fig. 8f: voltages v_{sd} and v_{sq}

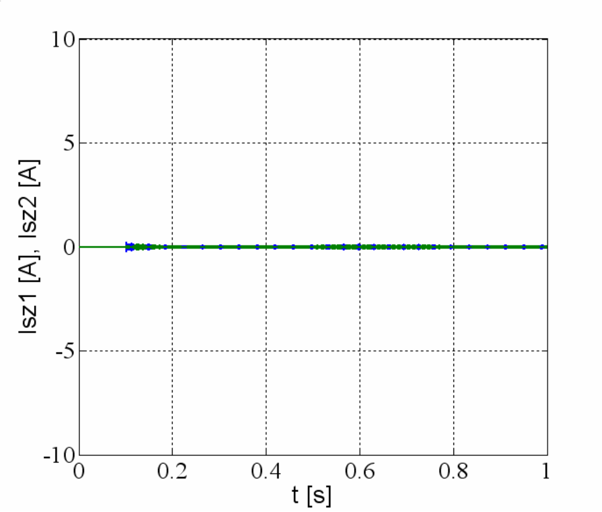


Fig. 8g: harmonic currents i_{z1} and i_{z2}

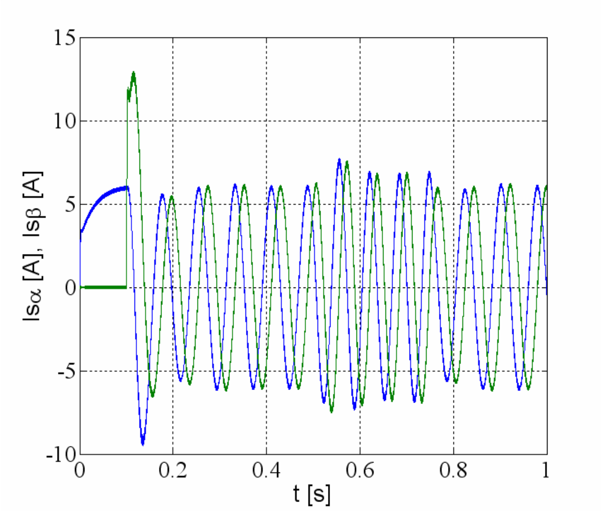


Fig. 8h: phase currents $i_{s\alpha}$ and $i_{s\beta}$

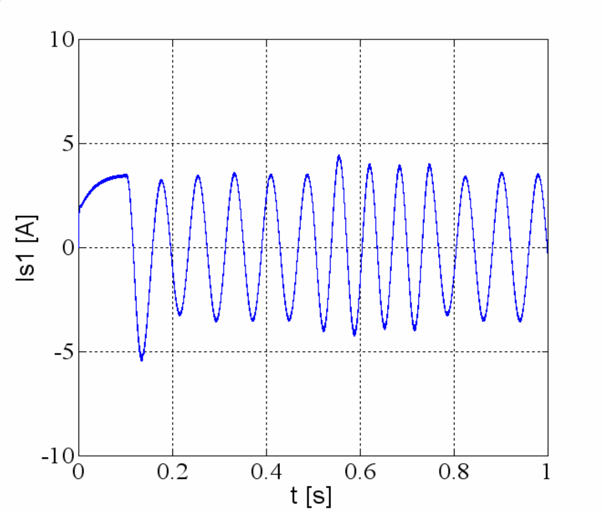


Fig. 8i: phase current i_{sa1}

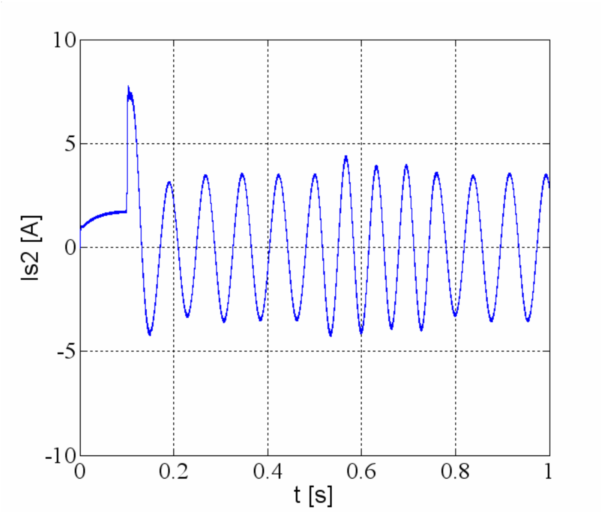


Fig. 8j: phase current i_{sa2}

Fig. 8- Simulation results for a load torque rejection test Ω_{ref}=20 *rad/s* and T_L=5 *Nm*.

7 Conclusion

In this paper, we have proposed a new method of field orientation control for SPIM based on space vector PWM and double d-q frame current controller. The proposed method uses a simple estimator for identifying rotor flux and its orientation. Space vector PWM technique based on the space vector decomposition is employed. For canceling unbalanced current sharing between the two three-phase windings sets due to the small system asymmetries, double d-q synchronous frame current controller is used.

We have tested this method under different operational conditions, by simulation. We have obtained good results in the different operational conditions. The simulation results show the efficiency of the proposed method.

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