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A FIELD ORIENTED BASED METHOD FOR DTC OF INDUCTION MOTOR

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Abstract **– This paper present a direct torque control, (DTC) method for the induction machine based on the field oriented control principle (FOC). Basically, we start from the voltage equation set of the induction motor referred to the same coordinate system dλ – qλ, oriented with respect to the stator flux linkage. Subsequently two control regimes may result: stator current control within dλ – axis, towards its limit, for a fast increase of the stator flux linkage and the simultaneous control of both flux linkage and torque. The proposed method provides better results during start-up and in the steady-state regime. This method may be used both in the lower speed and in higher speed domains, torque and flux linkage oscillations are diminished.**

Keywords: *induction motor, field oriented control, direct torque control.*

1. INTRODUCTION

In induction motors' direct torque control method, the electromagnetic torque is directly monitored through selecting inverter's switching vectors within a look-up table. Because considering only the space vector's location within sectors without considering precise space vector positioning within a sector, this selection, obviously, isn't the best choice, [1], [2]. The deadbeat solution has a fast torque response but requests complex calculations and pulse-width modulation to shaping the correct deadbeat voltage. A solution to simplify computations could be solving stator's voltage equation in field oriented coordinate system. In this approach, the computations of deadbeat voltage control are simpler because neglecting resistive voltage drops.

2. INDUCTION MOTOR'S MODEL

Voltage equations and flux-linkage equations of induction machine in the same arbitrary coordinate system, dλ - qλ are given by:

$$\begin{aligned} \underline{u}_{s\lambda} &= R_s \cdot \underline{i}_{s\lambda} + \frac{d\underline{\Psi}_{s\lambda}}{dt} + j\cdot\omega_\lambda \cdot \underline{\Psi}_{s\lambda} \\ 0 &= R'_r \cdot \underline{i}_{r\lambda} + \frac{d\underline{\Psi}_{r\lambda}}{dt} + j\cdot(\omega_\lambda - \omega_r)\cdot \underline{\Psi}_{r\lambda} \end{aligned} \tag{1}$$

$$\begin{aligned} \underline{\Psi}_{s\lambda} &= L_s \cdot \underline{i}_{s\lambda} + L_m \cdot \underline{i}_{r\lambda} \\ \underline{\Psi}_{r\lambda} &= L_m \cdot \underline{i}_{s\lambda} + L'_r \cdot \underline{i}_{r\lambda} \end{aligned} \tag{2}$$

where:

- $\underline{u}_s$ stator voltage;
- $\underline{i}_s$, $\underline{\Psi}_s$ stator current and stator flux linkage;
- $\underline{i}_r$, $\underline{\Psi}_r$ rotor current and stator flux linkage;
- ω_λ angular speed of the arbitrary rotating system;
- ω_r angular speed of the rotor;
- $L'_r = k^2 \cdot L_r \quad R'_r = k^2 \cdot R_r$ are rotor self inductance and rotor resistance referred to stator's number of windings respectively.

The instantaneous value of the electromagnetic torque may be expressed as the vector product of the stator-flux and stator current space vectors as follows:

$$T_e = \frac{3\cdot p}{2}\cdot\left(\underline{\Psi}_{s\lambda} \times \underline{i}_{s\lambda}\right) = \frac{3}{2}\cdot\frac{p}{\sigma\cdot L_s}\cdot\left(\underline{\Psi}_{r\lambda} \times \underline{\Psi}_{s\lambda}\right), \tag{3}$$

where $\sigma = 1 - L_m^2 / L_s \cdot L_r$ and p is the number of pair-poles.

The flux-leakage is given by:

$$\underline{\Psi}_{\sigma\lambda} = L_\sigma \cdot i_{s\lambda} = \underline{\Psi}_{s\lambda} - \underline{\Psi}_{r\lambda}, \tag{4}$$

where $L_{s\sigma} = \sigma\cdot L_s = L_s - L'_r$.

Equations (1) and (2) arranged as state-space equations lead to:

$$\frac{d}{dt}\begin{bmatrix}\mathbf{\Psi}_{s\lambda}\\ \mathbf{\Psi}_{r\lambda}\end{bmatrix} = \begin{bmatrix} -\left(\frac{1}{T'_s}\right)\cdot\mathbf{I} - \omega_\lambda\cdot\mathbf{J} & \frac{1}{T'_s}\cdot\mathbf{I} \\ \left(\frac{1-\sigma}{T'_s}\right)\cdot\mathbf{I} & -\left(\frac{1}{T'_r}\right)\cdot\mathbf{I} - (\omega_\lambda - \omega_r)\cdot\mathbf{J} \end{bmatrix} \times \begin{bmatrix}\mathbf{\Psi}_{s\lambda}\\ \mathbf{\Psi}_{r\lambda}\end{bmatrix} + \begin{bmatrix}\mathbf{I}\\ 0\end{bmatrix}\cdot\mathbf{u}_{s\lambda} = \mathbf{A}_\lambda\cdot\mathbf{x}_\lambda + \mathbf{B}_\lambda\cdot\mathbf{u}_{s\lambda}, \tag{5}$$

where:

$$\mathbf{x}_\lambda = \begin{bmatrix} \mathbf{\Psi}_{s\lambda} \\ \mathbf{\Psi}_{r\lambda} \end{bmatrix};$$

$$T_s' = \frac{L_\sigma}{R_s} = \frac{\sigma \cdot L_s}{R_s};$$

$$T_r' = \frac{\sigma \cdot L_r}{R_r};$$

$$\mathbf{I} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, \quad \mathbf{J} = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}.$$

If introducing another space vectors set $\underset{\sim}{\mathbf{x}}_\lambda$ as follows:

$$\underset{\sim}{\mathbf{x}}_\lambda = \begin{bmatrix} \mathbf{\Psi}_{s\lambda} \\ \mathbf{\Psi}_{\sigma\lambda} \end{bmatrix} = \begin{bmatrix} \mathbf{I} & 0 \\ \mathbf{I} & -\mathbf{I} \end{bmatrix} \cdot \mathbf{x}_\lambda, \tag{6}$$

then the state-space equation may be written as follows:

$$\frac{d\underset{\sim}{\mathbf{x}}_\lambda}{dt} = \\ = \begin{bmatrix} -\omega_\lambda \cdot \mathbf{J} & -\left(\frac{1}{T_s'}\right) \cdot \mathbf{I} \\ \left(\frac{1}{T_r}\right) \cdot \mathbf{I} - \omega_r \cdot \mathbf{J} & -\left(\frac{1}{T_s'} + \frac{1}{T_r'}\right) \cdot \mathbf{I} - (\omega_\lambda - \omega_r) \cdot \mathbf{J} \end{bmatrix} \times \\ \times \underset{\sim}{\mathbf{x}}_\lambda + \begin{bmatrix} \mathbf{I} \\ \mathbf{I} \end{bmatrix} \cdot \mathbf{u}_{s\lambda} \tag{7}$$

where:

$$T_r = \frac{L_r}{R_r} = \frac{T_r'}{\sigma}.$$

3. STATOR-FLUX FIELD-ORIENTED DTC FOR INDUCTION MOTORS

Voltage source inverters (VSI) are commonly used in induction motors electric drives. These inverters have eight-commutation states, (Fig. 1). The stator space vector's magnitude is $U = 2/3 \cdot U_d$, where U_d is the DC-link voltage. This value is almost equal to the peak-value of the rated phase-voltage of the induction motor. In Fig. 1, d and q are the axis of the stationary reference frame and a_s, b_s and c_s are the axis of the three-phase windings of the stator.
If resistance voltage drops are neglected and the axis $d\lambda$ is oriented with respect to the stator flux-linkage (Fig. 1) then results:

$$\dot{\underset{\sim}{\mathbf{x}}}_\lambda = \begin{bmatrix} -\omega_\lambda \cdot \mathbf{J} & 0 \\ -\omega_r \cdot \mathbf{J} & -(\omega_\lambda - \omega_r) \cdot \mathbf{J} \end{bmatrix} \cdot \underset{\sim}{\mathbf{x}}_\lambda + \\ + \begin{bmatrix} \mathbf{I} \\ \mathbf{I} \end{bmatrix} \cdot \underset{\sim}{\mathbf{u}}_{s\lambda} \tag{8}$$

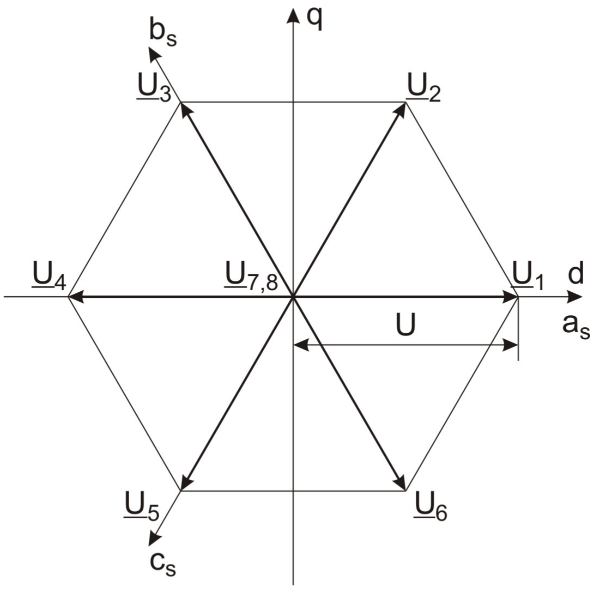

Fig. 1. Stator voltage vectors from the three-phase VSI.

Equation (8) is simpler because the system's matrix does not contain any time-constant.
From equation (8) results:

$$\frac{d\Psi_s}{dt} = u_{sd\lambda}, \tag{8.a}$$

$$0 = -\omega_\lambda \cdot \Psi_s + u_{sq\lambda}, \tag{8.b}$$

$$\frac{d\Psi_{\sigma d\lambda}}{dt} = (\omega_\lambda - \omega_r) \cdot \Psi_{\sigma q\lambda} + u_{sd\lambda}, \tag{8.c}$$

$$\frac{d\Psi_{\sigma q\lambda}}{dt} = -\omega_r \cdot \Psi_s - (\omega_\lambda - \omega_r) \cdot \Psi_{\sigma d\lambda} + u_{sq\lambda}. \tag{8.d}$$

The relation between flux-linkage space vectors in (8.a) - (8.b) is illustrated in Fig. 2.
The electromagnetic torque may be expressed from equation (3) as follows:

$$T_e = \frac{3 \cdot p}{2} \cdot \frac{\Psi_s \cdot \Psi_{r\lambda} \cdot \sin\gamma}{L_\sigma} = \\ = \frac{3 \cdot p}{2} \cdot \frac{\Psi_s \cdot \Psi_{\sigma\lambda q}}{L_\sigma}. \tag{9}$$

Equation (10) relates that rotor's flux built-up is essential to producing torque into an induction machine. The rotor-flux increases slower than the stator-flux because the rotor time-constant is greater than the stator time-constant. There are two control procedures to enhance torque according to the rotor-flux magnitude:
1. Only stator-current component with respect to $d\lambda$ axis is controlled until the rotor-flux achieves the

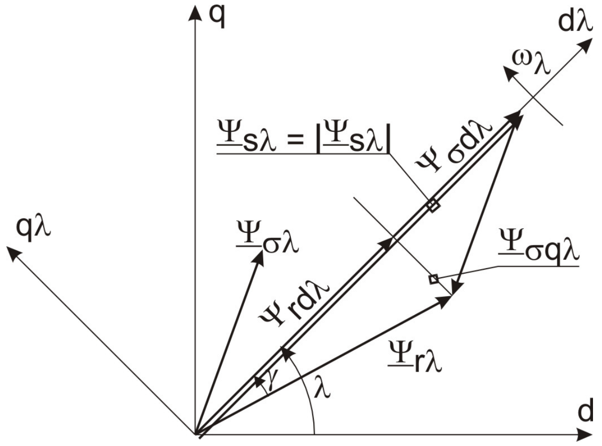

Fig. 2. Flux-linkage space vectors (oriented with respect to the stator flux-linkage).

value $\Psi_{rd\lambda} < \Psi_s^* - \Psi_{lim}$; Ψ_s^* is the reference value of the stator-flux and $\Psi_{lim} = \Psi_{\sigma d\lambda} = L_s \cdot i_{\sigma d\lambda}$, where $i_{sd\lambda}$ is the maximum stator current.

2. When flux is exceeding the value, $\Psi_s^* - \Psi_{lim}$ the controller is switching the system to the second regime; in this regime, flux and torque are simultaneously controlled.

The computation of the stator voltage vector is based on a predictive diagram. The purpose of this computation is the deadbeat control of flux-linkage and torque over a constant sample period. Considering a sample period sufficiently short, then the voltage at the motor's terminals should verify the following condition [5]:

$$\left(\left. \frac{dx}{dt} \right|_{\underline{u}_{s\lambda}^*} \right) \cdot T_S = x^* - x(t_k). \qquad (10)$$

From relations (8.a) – (8.b) and (9) the flux-linkage and torque transformations according to the operation regimes are obtained:

Regime 1: $$\frac{d\Psi_{\sigma d\lambda}}{dt} = u_{sd\lambda} \qquad (11.a)$$

Regime 2:

$$\frac{d}{dt}\begin{bmatrix} \Psi_s \\ \dfrac{L_\sigma \cdot T_e}{\frac{3}{2} \cdot p \cdot \Psi_s^*} \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ \dfrac{\Psi_{\sigma q\lambda}}{\Psi_s^*} & \dfrac{\Psi_{rd\lambda}}{\Psi_s^*} \end{bmatrix} \times \times \begin{bmatrix} u_{sd\lambda} \\ u_{sq\lambda} - u_\omega \end{bmatrix} \qquad (11.b)$$

where: $u_\omega = \omega_r \cdot \Psi_s$.

The stator voltage for the deadbeat response is obtained by the substitution of the relations (11.a) and (11.b) into the relation (10):

$$\underline{u}_{s\lambda}^* = f_s \cdot \Delta\underline{\Psi}_{u\lambda} + \underline{u}_{\omega\lambda} \qquad (12)$$

Regime 1: $$\Delta\underline{\Psi}_{u\lambda} = \Psi_{lim} - \Psi_{\sigma d\lambda} \qquad (12.a)$$

$$\underline{u}_{\omega\lambda} = 0 \qquad (12.b)$$

Regime 2:

$$\Delta\underline{\Psi}_{u\lambda} = \Psi_s^* - \Psi_s + j \cdot \left(\frac{\Psi_s^*}{\Psi_{rd\lambda}} \right) \times \times \left(\Psi_{\sigma q\lambda}^* - \Psi_{\sigma q\lambda} \right) \qquad (12.c)$$

$$\underline{u}_{\omega\lambda} = j \cdot u_\omega \qquad (12.d)$$

where:

- f_s the sampling frequency, $1/T_s$;
- $\Delta\underline{\Psi}_{u\lambda}$ is the unified flux error;
- $\underline{u}_{\omega\lambda}$ is the speed voltage.

and:

$$\Psi_{\sigma q\lambda}^* = \frac{L_\sigma \cdot T_e^*}{\frac{3}{2} \cdot p \cdot \Psi_s^*}.$$

Estimated waveforms of a no-load start-up at referred constant torque of 18 Nm are presented in Fig 3. Plots are scaled in per-units, $\Psi_s^* = 1.0$ p.u. and the flux limit is set to 0.3 p.u. The two operation regimes may be clearly distinguished. In the first operating regime, the control algorithm is automatically limiting the current. After the rotor-flux component with respect to the dλ axis achieves 0.7 p.u., the

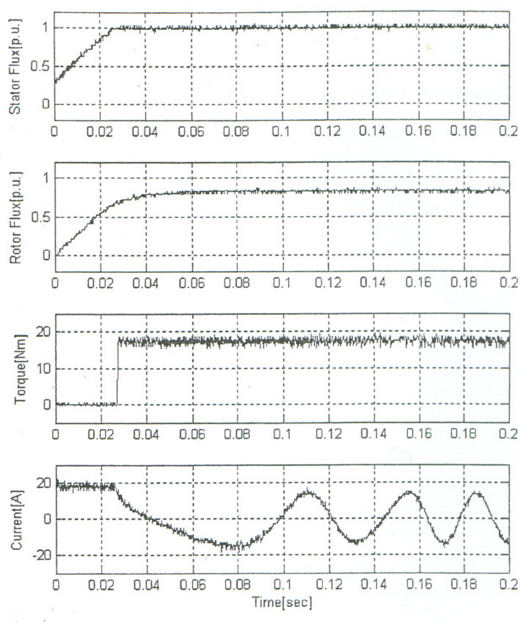

Fig. 2. Waveforms at no-load start-up with constant referred torque.

controller is switching the system to the second regime and produces torque at a constant (referred) flux.

CONCLUSIONS

DTC conventional has less selectable voltage space vectors. This produces flux and torque ripples. With the method presented above better performances at start-up and during the stationary regime are obtained. The transition from one regime to the other may be easily recognized if monitorizing rotor-flux magnitude. The method may be implemented both at low speed and at high-speed regimes. Flux and torque ripples are decreased.

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