International Conference on Control, Engineering & Information Technology (CEIT'13)
Proceedings Engineering & Technology - Vol.3, pp. 169-173, 2013

Sonsorless Induction Motor Drive Based on Deadbeat Torque and Flux Control

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***Abstract*— A stator flux oriented induction motor (IM) drives with deadbeat direct torque and flux control is proposed in this paper as a new strategy of the DTC drive. This algorithm calculates the voltage vector necessary to be generated by the converter in order to eliminate the flux and the torque errors with the advantage of a fixed switching frequency and good transient and steady state performance. Gopinath Observer is designed for stator flux estimation. The rotor speed estimation is performed using a model reference adaptive system (MRAS) estimator. The simulation results indicate that the proposed strategy is capable of maintaining a constant switching frequency, eliminating the flux and torque ripples, the major drawback of the classical DTC, and showed the efficiency of the proposed method in transient and steady state regimes.**

***Keywords*— Deadbeat control, Gopinath Observer, MRAS, Induction motor.**

I. INTRODUCTION

In recent years significant advances have been made on the sensorless control of IMs. One of the most well-known methods used for the control of AC drives is the Direct DTC strategy developed by Takahashi in 1984 [1]. DTC of IM is known to have a simple control structure with comparable performance to that of the field-oriented control (FOC) technique developed by Blaschke in 1972 [2]. Unlike FOC method, DTC technique requires utilization of hysteresis band comparators instead of flux and torque controllers [3]. To replace the coordinate transformations and pulse width modulation (PWM) inverter, DTC involves the use of a look-up table to select the voltage vectors based on torque and stator flux magnitude error [4-7].

DTC improves the IM controller dynamic performance and reduces the influence of the parameter variation during the operation. However, the sensorless implementation of this technique suffers from variable switching frequency according to the motor speed and the hysteresis bands of torque and flux comparators [4-5], witch leads to non- sinusoidal phase currents and noticeable torque and stator flux ripples. In addition, DTC of IM requires an accurate knowledge of the magnitude and angular position of the controlled flux. In the classical DTC, the flux is conventionally obtained from the stator voltage model, using the measured stator voltages and currents. This method, uses an open loop pure integration suffering from the well known problems of integration effects in digital systems, especially at low speeds operation range [3]. An alternative method of direct torque control is based on the deadbeat (inverse) solution to the machine equations [6-7]. The deadbeat solution is similar to the classical direct torque control method in that it controls torque and stator flux directly, without an intermediate current loop. It is different, however, in the calculation of the voltage vector to be applied to the machine. In the deadbeat solution, an inverse model is used to calculate the theoretical voltage vector needed to move the machine torque and stator flux to the desired values in one sample period [7].

In the last decade, many researches have been carried on the design of sensorless control schemes of the IM. Most methods are basically based on the MRAS [8-9-10]. The basic MRAS algorithm is very simple but its greatest drawback is the sensitivity to uncertainties in the motor parameters. Other method based on the EKF algorithm is used [11-12]. The EKF is a stochastic state observer where nonlinear equations are linearized in every sampling period. An interesting feature of the EKF is its ability to estimate simultaneously the states and the parameters of a dynamic process. This is generally useful for both the control and the diagnosis purposes. In [9] and [13] the authors used the Luenberger Observer for state estimation of the IM. The Extended Luenberger Observer (ELO) is a deterministic observer which also linearizes the equations in every sampling period. There is other type of methods for state estimation that is based on the intelligent techniques is used by many authors [15-16]. Fuzzy logic and neural networks has been a subject of growing interest in recent years. Neural network and fuzzy logic algorithms are quite heavy for basic microprocessors. In addition, several papers provide sensorless control of IM that are based on the variable structure technique [17-18] and nonlinear Observers [19].

In this paper, to reduce torque and stator flux ripples and avoid the operation with a variable switching frequency, a DTC based on deadbeat strategy is used. Also, the pure integration method used in the classical DTC for stator flux estimation of IMs suffering from the well known problems of integration especially at low speed operation range is resolved by the use of the Gopinath Observer to estimate the stator flux linkages [19]. Besides, the speed estimation witch is performed using MRAS estimator. To keep the high performance of the control scheme, the control algorithm is updated by the estimated values at each sample time.

II. INDUCTION MOTOR MODEL

The induction motor mathematical model in space vector notation, established in the d-q coordinate system rotating at speed ω_s is given by the following equations.

$$\overline{V}_s = R_s\overline{i}_s + \dot{\overline{\psi}}_s + j\omega_s\overline{\psi}_s \tag{1}$$

$$\overline{V}_r = \overline{0} = R_r\overline{i}_r + \dot{\overline{\psi}}_r + j(\omega_s - \omega_r)\overline{\psi}_r \tag{2}$$

$$\overline{\psi}_s = L_s\overline{i}_s + L_m\overline{i}_r \tag{3}$$

$$\overline{\psi}_r = L_r\overline{i}_r + L_m\overline{i}_s \tag{4}$$

$$T_e = p(\overline{i}_s \times j\overline{\psi}_s) \tag{5}$$

Where $\overline{V}_s, \overline{V}_r$ are the stator and rotor voltage vectors respectively, $\overline{i}_s, \overline{i}_r$ are the stator and rotor current vectors, $\overline{\psi}_s, \overline{\psi}_r$ are the stator and rotor flux vectors, ω_s, ω_r are the stator and rotor speeds, R_s, R_r are the stator and rotor resistances, L_s, L_r are the stator and rotor self inductances, L_m is the mutual inductance and p is the pole-pair number.

Using the d-q coordinate system and separating the machine variables into their real and imaginary parts, the time-varying state space model of the induction motor is given by the following equation:

$$\begin{bmatrix} \dot{i}_{sd} \\ \dot{i}_{sq} \\ \dot{\psi}_{sd} \\ \dot{\psi}_{sq} \\ \dot{\omega}_r \end{bmatrix} = \begin{bmatrix} -\gamma i_{sd} + (\omega_s - \omega_r)i_{sq} + \frac{1}{\sigma L_s\tau_r}\psi_{sd} + \frac{\omega_r}{\sigma L_s}\psi_{sq} \\ -(\omega_s - \omega_r)i_{sd} - \gamma i_{sq} - \frac{\omega_r}{\sigma L_s}\psi_{sd} + \frac{1}{\sigma L_s\tau_r}\psi_{sq} \\ -R_s i_{sd} + \omega_s\psi_{sq} \\ -R_s i_{sq} - \omega_s\psi_{sd} \\ \frac{p^2}{J}\left(\psi_{sd}i_{sq} - \psi_{sq}i_{sd}\right) - \frac{p}{J}T_l - \frac{f_v}{J}\omega_r \end{bmatrix} + \begin{bmatrix} \frac{1}{\sigma L_s}V_{sd} \\ \frac{1}{\sigma L_s}V_{sq} \\ V_{sd} \\ V_{sq} \\ 0 \end{bmatrix} \tag{6}$$

Where τ_r is the rotor time constant, T_l is the load torque, J is the motor inertia, f_v is the viscous friction coefficient, $\sigma = 1 - L_m^2/(L_sL_r)$ is the total leakage coefficient and γ is a constant defined as : $\gamma = \frac{1}{\sigma}(R_s/L_s + R_r/L_r)$.

III. DEADBEAT CONTROL STRATEGY

In some applications, it is preferred to have constant switching frequency to obtain quasi-sinusoidal phase current [6]. Thus, a Direct Deadbeat Torque and Flux Control (DDTFC) strategy recently proposed was used. The voltage vector necessary to eliminate the flux and torque errors in one switching period is determined in a stator flux reference frame [7]. This calculated voltage vector serves as a reference to a PWM inverter. The flux and torque controllers are described below.

A. Flux control

In a stator flux reference frame, where the d-axis is aligned with the stator flux space vector, the q axis flux component is zero and the flux dynamics can be imposed only by the d axis stator voltage component [6].

$$V_{sd} = R_s i_{sd} + \dot{\psi}_{sd} \tag{7}$$

Considering T_s as the sampling time, Eq. (7) can be written as a discrete forward difference equation:

$$\psi_{sd}(k+1) = \psi_{sd}(k) + [\, V_{sd}(k) - R_s i_{sd}(k)\,]\, T_s \tag{8}$$

In order to impose a deadbeat flux control, it should be considered that $\psi_{sd}(k+1) = \psi_s^*$ then:

$$V_{sd}^*(\,k\,) = \frac{\psi_s^* - \hat{\psi}_{sd}(k)}{T_s} + R_s i_{sd}(k) \tag{9}$$

Where the symbols * and ^ denote the commanded value and the estimated value, respectively.

B. Torque control

Being p the pole pair number, the electromagnetic torque is determined by the following expression:

$$T_e(k) = p\psi_{sd}(k)i_{sq}(k) \tag{10}$$

Considering that the flux is conveniently controlled, the torque depends only on the q axis current. This current is strongly related to the stator flux speed as it can be seen from Eq. (11), rewritten in terms of stator flux and current components in a stator flux reference frame:

$$\omega_s = \omega_r + \frac{\frac{L_s}{\tau_r}i_{sq} + \sigma L_s\dot{i}_{sq}}{\psi_{sd} - \sigma L_s i_{sd}} \tag{11}$$

By the use of a sufficiently small sampling period, the current derivative in (11) can be written as $(i_{sq}(k+1) - i_{sq}(k))/T_s$. Imposing the electromagnetic torque to be equal to the commanded value in the next sampling instant results in:

$$T_e(k+1) = p\hat{\psi}_{sd}(k)i_{sq}(k+1) = T_e^* \tag{12}$$

Combining Eqs. (11) and (12) gives:

$$\omega_s^*(k) = \omega_r(k) + \frac{\frac{L_s}{\tau_r}i_{sq}(k) + \frac{\sigma L_s}{T_s}\left(\frac{T_e^*}{p\hat{\psi}_{sd}(t)} - i_{sq}(k)\right)}{\hat{\psi}_{sd}(k) - \sigma L_s i_{sd}(k)} \tag{13}$$

The q axis component of the commanded voltage is given by the following equation:

$$V_{sq}^*(k) = R_s i_{sq}(k) + \omega_s^*(k)\hat{\psi}_{sd}(k) \tag{14}$$

The equations (9) and (14) are applied to the PWM after a transformation to the abc-three-phase reference frame, the flux and torque will have the respective desired values at the next sampling instant. The necessary quantities to be measured for the technique implementation are the stator currents and voltages.

Besides the advantage regarding the simplicity of the presented algorithm, the implementation in synchronous reference frame avoids the problem of division by zero, witch occurs for stationary reference frame implementation. The additional rotational transformations do not represent an important computational effort for the digital processor [6].

Fig. 1 shows the block diagram of the sensorless DDTFC scheme for the induction motor drive.

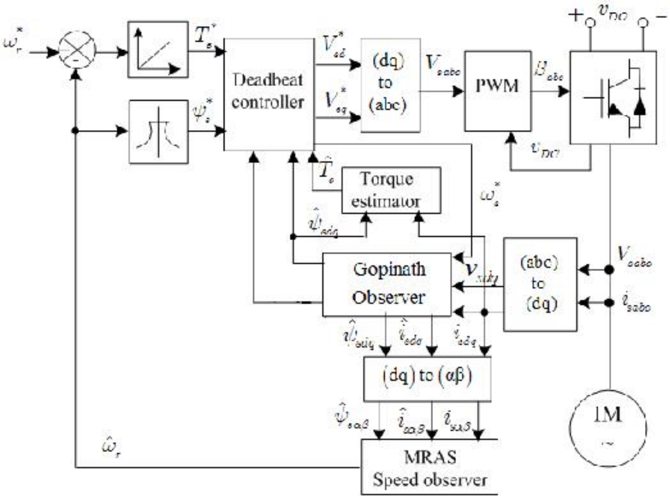

Fig. 1. Block diagram of the proposed sensorless deadbeat control strategy.

IV. GOPINATH OBSERVER

In order to evaluate the effect of the voltage measuring scheme and whether a flux observer is able or not to compensate for the voltage errors, a reduced order Gopinath stator flux observer was implemented. A feedback term based on the derivative of the stator current error was added to the stator voltage equation in a stator reference frame to improve the observer dynamic response.

$$\begin{cases} \dot{\vec{\psi}}_s = \vec{V}_s - R_s\vec{i}_s \\ \dot{\hat{\vec{\psi}}}_s = \vec{V}_s - R_s\vec{i}_s + L(\dot{\hat{\vec{i}}}_s - \vec{i}_s) \end{cases} \quad (15)$$

The stator current derivative in Eqn. (20) is calculated from stator voltage, stator current and stator flux, using the induction motor state model. Combining Eqs (2) and (4) gives

$$\dot{\vec{\psi}}_r = -\left(\frac{R_r}{L_r} - j\omega_r\right)\vec{\psi}_r + \frac{R_r}{L_r}L_m\vec{i}_s \quad (16)$$

From Eqs. (3) and (4), it is seen that

$$\vec{\psi}_r = \frac{R_r}{L_m}\vec{\psi}_s - \frac{L_r}{L_m}\sigma L_s\vec{i}_s \quad (17)$$

Substituting (17) in (16):

$$\dot{\vec{\psi}}_s = R_r\frac{L_m^2}{L_r^2}\vec{i}_s + \gamma L_s\dot{\vec{i}}_s - \left(\frac{R_r}{L_r} - j\omega_r\right)\left(\vec{\psi}_s - \gamma L_s\vec{i}_s\right) \quad (18)$$

Finally, from Eqs. (18) and (1), the stator current derivative can be estimated as

$$\begin{cases} \dot{\vec{i}}_s = \dfrac{1}{\gamma L_s}\left[\vec{V}_s - R_s'\vec{i}_s + \left(\dfrac{R_r}{L_r} - j\omega_r\right)\vec{\psi}_s - \left(\dfrac{R_r}{L_r} - j\omega_r\right)\gamma L_s\vec{i}_s\right] \\ \dot{\hat{\vec{i}}}_s = \dfrac{1}{\gamma L_s}\left[\vec{V}_s - R_s'\vec{i}_s + \left(\dfrac{R_r}{L_r} - j\omega_r\right)\hat{\vec{\psi}}_s - \left(\dfrac{R_r}{L_r} - j\omega_r\right)\gamma L_s\vec{i}_s\right] \end{cases} \quad (19)$$

Where

$$R_s' = R_s + R_r\frac{L_m^2}{L_r^2}$$

The stator flux observer error is governed by:

$$\begin{bmatrix} \dot{\varepsilon}_d \\ \dot{\varepsilon}_q \end{bmatrix} = \begin{bmatrix} \dot{\psi}_{sd} \\ \dot{\psi}_{sq} \end{bmatrix} - \begin{bmatrix} \dot{\hat{\psi}}_{sd} \\ \dot{\hat{\psi}}_{sq} \end{bmatrix} = A_{ob}\begin{bmatrix} \varepsilon_d \\ \varepsilon_q \end{bmatrix} \quad (20)$$

Where $A_{ob} = \dfrac{1}{\gamma L_s}L\begin{bmatrix} \dfrac{1}{\tau_r} & \omega_r \\ -\omega_r & \dfrac{1}{\tau_r} \end{bmatrix}$

L is the observer gain. Choosing, for example, $L = \begin{bmatrix} L_1 & 0 \\ 0 & L_1 \end{bmatrix}$ forces the observer poles to be

$$s_{1,2} = \frac{1}{\gamma L_s}\left(\frac{L_1}{\tau_r} \pm j\omega_r L_1\right) \quad (21)$$

In order to avoid differentiating, the stator current, the observer Eq. (15) is written in terms of new variable:

$$\hat{\vec{Y}} = \overrightarrow{\psi}_s + L\vec{i}_s \quad (22)$$

from (15), (18) and (22), a state equation to obtain $\hat{\vec{Y}}$ is:

$$\dot{\hat{\vec{Y}}} = A_{ob}\hat{\vec{Y}} - \left[A_{ob}(L + \gamma L_s I) + R_s I + \frac{R_s'}{\gamma L_s}L\right]\vec{i}_s + \left(I + \frac{1}{\gamma L_s}L\right)\vec{V}_s \quad (23)$$

where $I = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$

Equation (23) is solved each sampling period and the observer stator flux is then obtained from (22)

V. MRAS BASED ROTOR SPEED ESTIMATION

The MRAS technique is used in sensorless IM drives, at the first time, by Schauder [8]. Since this, it has been a topic of many publications [9-10]. The MRAS is important since it leads to relatively easy to implement system with high speed of adaptation for a wide range of applications. The basic scheme of the parallel MRAS configuration is given in Fig. 3. The scheme consists of two models; reference and adjustable ones and an adaptation mechanism. The block "reference model" represents the actual system having unknown parameter values. The block "adjustable model" has the same structure of the reference one, but with adjustable parameters instead of the unknown ones. The block "adaptation mechanism" estimates the unknown parameter using the error between the reference and the adjustable models and updates the adjustable model with the estimated parameter until satisfactory performance is achieved.

Using a proportional plus integral (PI) controller, the IM speed observer equation is given by (24) [14]:

$$\hat{\omega}_r = K_P(\varepsilon_\beta \hat{\psi}_{s\alpha} - \varepsilon_\alpha \hat{\psi}_{s\beta}) + K_I \int_0^t (\varepsilon_\beta \hat{\psi}_{s\alpha} - \varepsilon_\alpha \hat{\psi}_{s\beta})dt \qquad (24)$$

Where $\varepsilon_\alpha = i_{s\alpha} - \hat{i}_{s\alpha}$ and $\varepsilon_\beta = i_{s\beta} - \hat{i}_{s\beta}$. *Ki* and *Kp* are the integral and proportional observer gains. There values have an important role for the tracking performance of the speed estimation and the sensitivity to noises. Therefore, the integral gain *Ki* is chosen high for fast tracking of speed. While, a low proportional *Kp* gain is needed to attenuate high frequency signals denoted noises [14]. The stability and the convergence of this observer have been proven in several papers [9].

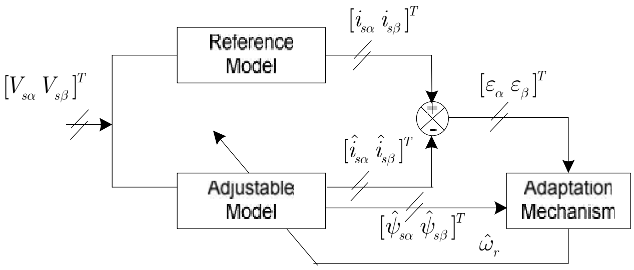

Fig. 3. MRAS Scheme for rotor speed estimation

VI. Simulation Results

The efficiency of the proposed control scheme has been verified using MATLAB/SIMULINK software package. Motor parameters used in simulations are given in the Appendix.

Fig. 4 shows the actual and the estimated *d* and *q* components of the stator flux. Fig.5 shows the actual and the estimated *d* and *q* components of the stator currents. The reversal speed response for a square wave reference trajectory from -8 rad/s to +8 rad/s and with no load is shown in Fig. 6. The electromagnetic torque is shown in Fig. 7. It is clearly shown that the estimated variables are in close agreement with the real ones for a no load, low and reversal speed reference.

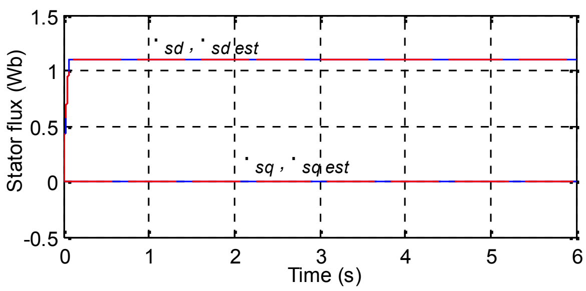

Fig. 4. The actual and estimated stator fluxes.

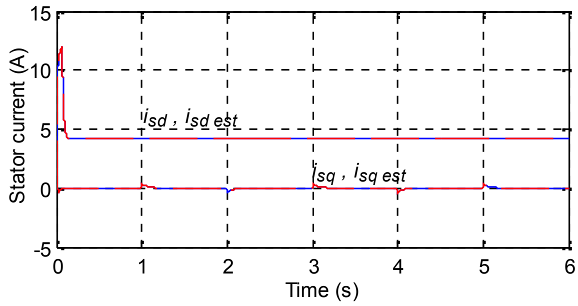

Fig. 5. . The actual and estimated stator currents.

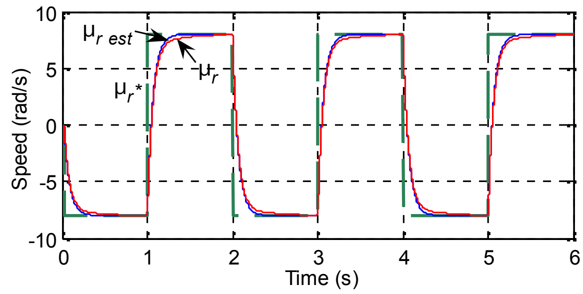

Fig. 6. The actual and estimated rotor speeds for a square reference trajectory and at low speed range

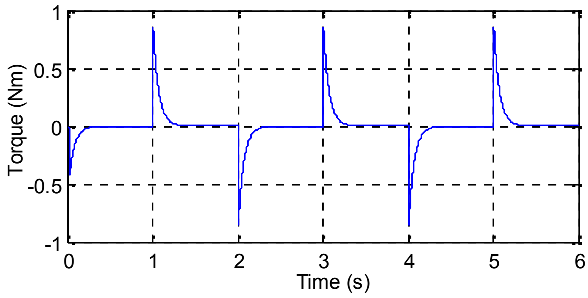

Fig. 7. Electromagnetic torque

The steady-state operation at maximum motor speed (157 rad/s) and under variable load torque is illustrated in Figs 8 and 9. The load torque impact on the speed estimation is studied under different level of load variations. The load torque is initially set to 0 Nm, then it is changed to 5 Nm at 1.5 s and to the rated value at 2.5 s and at 3.5 s it is kept again to 0 Nm. Fig 8 shows the actual and the estimated rotor speed. Fig. 9 shows simultaneously the load and the electromagnetic torques. The robustness of the MRAS speed estimation to the load torque variations is clearly visible.

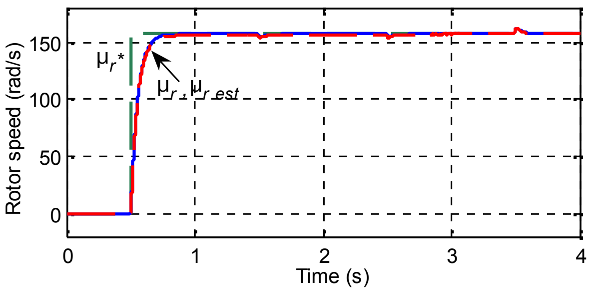

Fig. 8. The actual and estimated rotor speed.

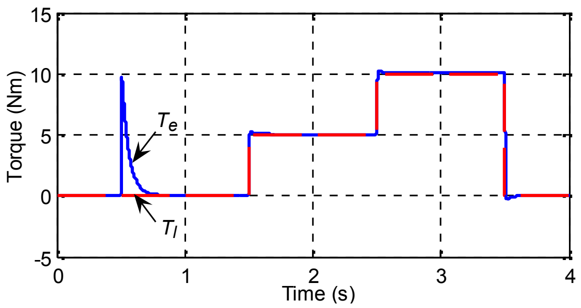

Fig. 9. The electromagnetic and load torques

VII. CONCLUSIONS

The paper deals with a new direct deadbeat torque and flux control of IM. This algorithm calculates the voltage vectors needed to move the machine torque and stator flux to the desired values in one sample period. It allows by means of pulse width modulation (PWM) block, constant switching frequency and overcome the major drawback of the classical DTC, witch is the torque and flux ripples. The Gopinath Observer is designed and developed for use in closed loop control for stator flux estimation.

Besides the advantages' regarding the simplicity of the presented algorithm, the implementation in synchronous reference frame avoids the problem of division by zero, witch occurs for stationary reference frame implementation. The additional rotational transformations do not represent an important computational effort for the digital processor.

Appendix

Table 1. Motor Data

Rated power	3 kW
Rated speed	1440 rpm
frequency	50 Hz
p	2
R_s	2.3 Ω
R_r	1.55 Ω
$L_s = L_r$	0.261 H
M	0.249 H
J	0.0076 kg.m^2

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