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A Practical Scheme for Induction Motor Speed Sensorless Field Oriented Control

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Abstract—This paper deals with a practical scheme of induction motor drive speed sensorless control. It requires the presence of an adaptive flux observer. The speed estimator is based on the induction motor magnetic characteristic slope approximation to the mutual inductance value. Encouraging simulation results have been obtained.

NOMENCLATURE

$V, I(i), \varphi$	=	voltage, current, flux;
$\omega (\Omega)$	=	rotor speed;
R, L, M	=	resistance; inductance, mutual inductance;
T_r	=	rotor time constant, $T_r = L_r/R_r$;
σ	=	total leakage coefficient, $\sigma = 1 - M^2/L_s L_r$;
γ	=	constant, $\gamma = -(R_s + M^2 R_r / L_r^2) / \sigma L_s$;
K	=	constant, $K = M / \sigma L_s L_r$;
$\bar{x} (\hat{x})$	=	complex (estimated) value.

I. INTRODUCTION

Induction motor drive researches have been concentrated in the elimination of speed sensor at the motor shaft without deteriorating the dynamic performance of the drive control system. Different schemes for speed sensorless in field orientation have been proposed in the past few years. A state of the art of the sensorless control of ac drive is given in [1].

This paper presents a practical solution to estimate the induction motor speed in the rotor flux oriented control system frame taking into account rotor resistance variations. In fact, vector control strategy is very sensitive to flux estimation mainly altered by motor parameter variations. It was then imperative to identify them [2-3]. Moreover, fast flux estimation and parameter identification achieve high dynamic performance and speed accuracy. Such features have been considered in the proposed control scheme [4-5].

II. THE INDUCTION MOTOR MODEL

An induction motor model can be described in a stator reference frame with the following equation.

$$\begin{bmatrix} \frac{d\bar{i}_s}{dt} \\ \frac{d\bar{\varphi}_r}{dt} \end{bmatrix} = \begin{bmatrix} \gamma & \frac{K}{T_r} - j\omega K \\ \frac{M}{T_r} & -\frac{1}{T_r} + j\omega \end{bmatrix} \begin{bmatrix} \bar{i}_s \\ \bar{\varphi}_r \end{bmatrix} + \begin{bmatrix} 1 \\ 0 \end{bmatrix} \frac{\bar{V}_s}{\sigma L_s} \quad (1)$$

The rotor flux orientation is achieved by the following transformation from α - β to d - q frame.

$$\begin{bmatrix} x_d \\ x_q \end{bmatrix} = \frac{1}{\sqrt{\varphi_{r\alpha}^2 + \varphi_{r\beta}^2}} \begin{bmatrix} \varphi_{r\alpha} & \varphi_{r\beta} \\ -\varphi_{r\beta} & \varphi_{r\alpha} \end{bmatrix} \begin{bmatrix} x_\alpha \\ x_\beta \end{bmatrix} \quad (2)$$

This transformation leads to

$$\begin{cases} \varphi_{rd} = \sqrt{\varphi_{r\alpha}^2 + \varphi_{r\beta}^2} \\ \varphi_{rq} = 0 \end{cases} \quad (3)$$

Figure 1 shows the general structure in which flux and torque are respectively controlled by the currents i_{sd} and i_{sq} in the synchronous rotating frame fixed with flux.

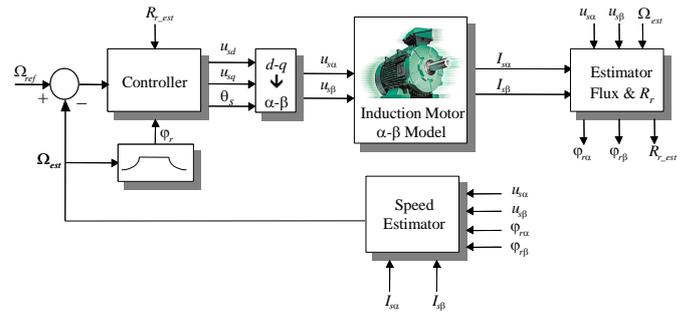


Fig. 1. Induction motor control structure.

III. THE SPEED ESTIMATOR

If we choose the motor speed ω as an intermediate variable, equation (1) could be written as

$$\begin{cases} \frac{d\bar{i}_s}{d\omega} \frac{d\omega}{dt} = \gamma \bar{i}_s + \left(\frac{1}{T_r} - j\omega \right) K \bar{\varphi}_r + \frac{\bar{V}_s}{\sigma L_s} \\ \frac{d\bar{\varphi}_r}{d\omega} \frac{d\omega}{dt} = \frac{M}{T_r} \bar{i}_s - \left(\frac{1}{T_r} - j\omega \right) \bar{\varphi}_r \end{cases} \quad (4)$$

From the above, the following equation is then obtained

$$\left(\frac{d\bar{\varphi}_r}{d\bar{i}_s} \right) \left(\gamma \bar{i}_s + \left(\frac{1}{T_r} - j\omega \right) K \bar{\varphi}_r + \frac{\bar{V}_s}{\sigma L_s} \right) = \frac{M}{T_r} \bar{i}_s - \left(\frac{1}{T_r} - j\omega \right) \bar{\varphi}_r \quad (5)$$

Where $d\bar{\varphi}_r / d\bar{i}_s$ represents the slope value for each point of the induction motor magnetic characteristic.

$$\bar{\varphi}_r = L_r \bar{i}_r + M \bar{i}_s \quad (6)$$

The magnetic saturation is neglected on the basis of a physical consideration. In fact, the induction motor temperature increase (thermal effect), due to its operation, will perturb and slow down the magnetic saturation process [5]. This has led us to consider M as a constant. In this case

$$\frac{d\bar{\varphi}_r}{d\bar{i}_s} = L_r \frac{d\bar{i}_r}{d\bar{i}_s} + M \quad (7)$$

This equation can be written as

$$\frac{d\bar{\varphi}_r}{d\bar{i}_s} = K_{sat} \quad (8)$$

Where K_{sat} represents the magnetic characteristic slope that changes with the motor load. In fact, the first part of (7) may be written as

$$L_r \frac{d\bar{i}_r}{d\bar{i}_s} = L_r \left(\frac{d\bar{i}_s}{dt} \right)^{-2} \left(\frac{d\bar{i}_r}{dt} \bullet \frac{d\bar{i}_s}{dt} + j \left(\frac{d\bar{i}_r}{dt} \times \frac{d\bar{i}_s}{dt} \right) \right) \quad (9)$$

In this case, it is numerically checked that (9) real part absolute value increases and does not exceed the mutual inductance value in loaded conditions. Furthermore, we have noticed that the imaginary part remains slight. Therefore, the two following statements are derived: $K_{sat} = M$ for unloaded motor operation; and $K_{sat} \neq 0$ for rated load motor operation.

With the above considerations and in the particular case of small and medium power induction motor, the following approximation could be written

$$K_{sat} \approx M \quad (10)$$

From (10) and (5), the motor speed estimator is extracted.

$$\hat{\omega} = \frac{1}{\left(K + \frac{1}{M} \right) \hat{\varphi}_r^2} \left[\left(\gamma - \frac{1}{T_r} \right) (\hat{i}_s \times \hat{\varphi}_r) + \frac{1}{\sigma L_s} (\bar{v}_s \times \hat{\varphi}_r) \right] \quad (11)$$

As it is shown by (11), the speed estimator is composed of two terms. The first one has the drawback of being sensitive to rotor and stator resistance that may vary owing to temperature. However, at motor rated load operation, we have

$$\frac{1}{\left(K + \frac{1}{M} \right) \hat{\varphi}_r^2} \ll 1 \quad (12)$$

Therefore, the parameter variation influence is attenuated.

VI. SIMULATION RESULTS

The proposed speed estimator has been simulated on a 4-kW induction motor (Fig. 1).

For simulation purposes, the transient behavior of the speed estimator is evaluated during 3 sec. At 1 sec, a 15 Nm load

torque is applied and removed at 2 sec. In these conditions, Fig. 2a shows that good performances are achieved and demonstrates the speed estimator robustness against parameter variations. Moreover, a change in the speed reference at 2.5 sec is well tracked. Figures 2b and 2c demonstrate the speed estimator capability for different induction motor operation conditions particularly those at low speed.

To illustrate the speed estimator robustness, computer simulations have been carried out in the case of an induction motor fed by a PWM-VSI inverter. Figure 2d clearly shows the robustness of the proposed speed estimator when a 15 Nm load torque is applied at 1 sec and removed at 2 sec and when the speed reference is change at 2.5 sec.

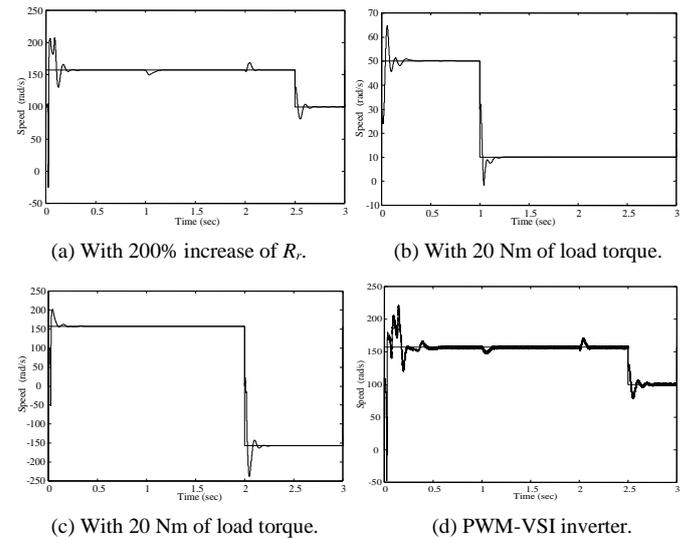


Fig. 2. Speed estimation.

V. CONCLUSION

This paper dealt with a practical speed estimator for induction motor sensorless control. It requires the presence of an adaptive flux observer. The speed estimator is based on the induction motor magnetic characteristic slope approximation to the mutual inductance value. Computer simulations show the feasibility and the robustness of the proposed speed estimator.

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