Sensorless Indirect Stator Field Orientation Speed Control for Single-Phase Induction Motor Drive

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***Abstract*—The industrial requirements for the control of an induction machine without a mechanical sensor continue to be of interest, as evidenced by the most recent publications. The focus is on improvements of control without a mechanical sensor. A new method for the implementation of a sensorless indirect stator-flux-oriented control (ISFOC) of a single-phase induction motor (SPIM) drive is proposed in this paper. The proposed method of rotor speed estimation is based only on the measurement of the main and auxiliary windings stator currents and that of a reference *q*-axis current generated by the control algorithm. The error of the measured *q*-axis current from its reference value feeds the proportional plus integral controller, the output of which is the estimated slip angular frequency. Experimental results for sensorless ISFOC speed control of a SPIM drive are presented and analyzed using a dSPACE system with DS1104 controller board based on the digital signal processor TMS320F240. Digital simulation and experimental results are presented to show the improvement in performance of the proposed sensorless algorithm.**

***Index Terms*—Indirect stator-flux-oriented control, sensorless vector control, single-phase induction motor, speed estimation.**

I. Introduction

SINGLE-PHASE induction motors (SPIM) are traditionally used in constant speed home appliances, usually in locations where only single-phase energy supply is available without any type of control strategy. They are found in air conditioners, washers, dryers, industrial machinery, fans, blowers, vacuum cleaners, and many other applications. Variable speed controls of electrical motors are widely employed in industrial applications because of the obvious energy-saving benefits. The cost reduction and high efficiency of power electronic and microelectronics devices are motivating to implement a SPIM drives in both industrial and domestic applications.

During recent years, many research laboratories have focussed on variable-speed drives, especially for the SPIM, and major improvements have been achieved. The availability of low-cost static converters makes possible the economic use of energy and improvement of the quality of the electromagnetic torque in SPIM [1]–[3]. They are three power electronic converter topologies of two-phase inverters for SPIM: two-leg, three-leg, and four-leg inverters. In recent years, the topology with three-leg two-phase with six-transistor bridge voltage source inverter for SPIM drive systems has been preferred by many researchers compared to the other topologies [1]–[13]. This suitable topology to supply the SPIM with two-orthogonal voltages system is cheaper than that of the four-leg inverter, and it gives a better performance in terms of harmonic distortion of the output voltage when compared to that of the two-leg inverter.

Nowadays, field-oriented controlled (FOC) induction motors are widely adopted to obtain high-dynamic performance in drive systems. The FOC represents a better solution to satisfy industrial requirements. The asymmetry of the SPIM has an important impact on the design of the control strategies. However, the stator flux model requires appropriate variable changes [5]. The drawback of this method is that the rotor speed of the SPIM must be measured, which requires a speed sensor. A sensorless system where the speed is estimated instead of measured would considerably reduce the cost and complexity of the drive system.

In the existing literature, many approaches have been suggested for sensorless vector speed control of SPIM drives in [14]–[17]. Some suggested methods for speed estimation using a machine model fed by stator quantities are parameter dependent; therefore, parameter errors can degrade speed control performance [18]. In paper [14], the authors evaluate a sensorless indirect rotor FOC in which the rotor flux vector frequency is estimated directly from measurable stator currents and voltages but is dependent on SPIM parameters. The sensorless speed control strategy using MRAS techniques is based on the comparison between the outputs of two estimators when motor currents and voltages must still be measured [6], [19]. The MRAS algorithm sensorless speed vector control of three-phase induction motor drive is sensitive to resistance variation [20].

In this paper, we propose a contribution to the issue of speed sensorless indirect stator-flux-oriented control (ISFOC) of SPIM drive based on [18]. The published paper [18] investigates the sensorless speed control of three-phase induction motor drive. The SPIM model equations are more complex than that of the three-phase induction machines, because the main and auxiliary stator windings have different resistances and inductances. However, the use of field orientation control for an unbalanced single-phase machine requires special attention,

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because the mathematical model for this type of machine is similar to that of an asymmetrical two-phase machine. Moreover, particular converter topology and control are used to supply the SPIM based on three-leg to generate two-phase voltage source inverter in which sinusoidal pulsewide modulation (PWM) is applied.

The estimated speed is obtained only from the measurement of the main and auxiliary windings stator currents and that of a reference q-axis current generated by the control algorithm. A speed estimation method is proposed to overcome the problems of system complexity and cost. Simulation and experimental results are presented to demonstrate the main characteristics of the proposed drive system. The sensorless speed control algorithm is employed in this work and is implemented at rated, low, and zero speed operation.

II. SPIM Model

The dynamic model for the single-phase induction machine in a stationary reference frame can be described by the following equations:

$$v_{sd}^s = R_{sd} i_{sd}^s + \frac{d\phi_{sd}^s}{dt} \tag{1}$$

$$v_{sq}^s = R_{sq} i_{sq}^s + \frac{d\phi_{sq}^s}{dt} \tag{2}$$

$$0 = R_r i_{rd}^s + \frac{d\phi_{rd}^s}{dt} + \omega_r \phi_{rq}^s \tag{3}$$

$$0 = R_r i_{rq}^s + \frac{d\phi_{rq}^s}{dt} - \omega_r \phi_{rd}^s \tag{4}$$

$$\phi_{sd}^s = L_{sd} i_{sd}^s + M_{srd} i_{rd}^s \tag{5}$$

$$\phi_{sq}^s = L_{sq} i_{sq}^s + M_{srq} i_{rq}^s \tag{6}$$

$$\phi_{rd}^s = L_r i_{rd}^s + M_{srd} i_{sd}^s \tag{7}$$

$$\phi_{rq}^s = L_r i_{rq}^s + M_{srq} i_{sq}^s \tag{8}$$

$$T_e = n_p (M_{srq} i_{sq}^s i_{rd}^s - M_{srd} i_{sd}^s i_{rq}^s) \tag{9}$$

where v_{sd}^s, v_{sq}^s, i_{sd}^s, i_{sq}^s, ϕ_{sd}^s, ϕ_{sq}^s, ϕ_{rd}^s, and ϕ_{rq}^s are the d-q axes voltages, currents, and fluxes of the stator and rotor in the stator reference frame; L_{sd}, L_{sq}, L_r, M_{srd}, and M_{srq} denote the stator and rotor self and mutual inductances; R_{sd}, R_{sq}, and R_r denote d-q axes stator and rotor resistances; and ω_r, T_e, and n_p are the rotor angular frequency, the electromagnetic torque, and the pole pairs.

Equations (1)–(8) present the model of an asymmetrical two-phase machine due to the unequal resistances and inductances of the main and auxiliary windings. This asymmetry causes an oscillating term in the electromagnetic torque [1]. As was done in [5] to drive the symmetrical model, here too, the mutual inductances will be employed to define a transformation for the stator variables. This transformation is given by

$$\begin{bmatrix} i_{sd}^s \\ i_{sq}^s \end{bmatrix} = T \begin{bmatrix} i_{sd1}^s \\ i_{sq1}^s \end{bmatrix}, \qquad \begin{bmatrix} v_{sd}^s \\ v_{sq}^s \end{bmatrix} = T^{-1} \begin{bmatrix} v_{sd1}^s \\ v_{sq1}^s \end{bmatrix} \tag{10}$$

and

$$\begin{bmatrix} \phi_{sd}^s \\ \phi_{sq}^s \end{bmatrix} = T^{-1} \begin{bmatrix} \phi_{sd1}^s \\ \phi_{sq1}^s \end{bmatrix}$$

where

$$T = \begin{bmatrix} 1 & 0 \\ 0 & k \end{bmatrix} \quad \text{and} \quad k = \frac{M_{srd}}{M_{srq}}.$$

Using (1)–(10), the new mathematical model of the SPIM in the stator reference frame can be described by the following equations:

$$v_{sd1}^s = R_{sd} i_{sd1}^s + \frac{d\phi_{sd1}^s}{dt} \tag{11}$$

$$v_{sq1}^s = R_{sd} i_{sq1}^s + \frac{d\phi_{sq1}^s}{dt} + (k^2 R_{sq} - R_{sd}) i_{sq1}^s \tag{12}$$

$$0 = R_r i_{rd}^s + \frac{d\phi_{rd}^s}{dt} + \omega_r \phi_{rq}^s \tag{13}$$

$$0 = R_r i_{rq}^s + \frac{d\phi_{rq}^s}{dt} - \omega_r \phi_{rd}^s \tag{14}$$

$$\phi_{sd1}^s = L_{sd} i_{sd1}^s + M_{srd} i_{rd}^s \tag{15}$$

$$\phi_{sq1}^s = L_{sd} i_{sq1}^s + M_{srd} i_{rq}^s + (k^2 L_{sq} - L_{sd}) i_{sq1}^s \tag{16}$$

$$\phi_{rd}^s = L_r i_{rd}^s + M_{srd} i_{sd1}^s \tag{17}$$

$$\phi_{rq}^s = L_r i_{rq}^s + M_{srd} i_{sq1}^s \tag{18}$$

$$T_e = n_p M_{srd} (i_{sq1}^s i_{rd}^s - i_{sd1}^s i_{rq}^s). \tag{19}$$

III. Indirect Stator-Flux-Oriented Control

Using (15), (16), and (19), electromagnetic torque as a function of stator fluxes and stator currents can be written as

$$T_e = n_p (\phi_{sd1}^s i_{sq1}^s - \phi_{sq1}^s i_{sd1}^s + \Delta T) \tag{20}$$

where $\Delta T = (k^2 L_{sq} - L_{sd}) i_{sq1}^s i_{sd1}^s$.

In the same way, using (13)–(18), we can determine the dynamic model that relates the stator flux to the stator currents.

$$\begin{aligned} &\frac{d\phi_{sd1}^s}{dt} + \frac{1}{\tau_r}\phi_{sd1}^s + \omega_r \phi_{sq1}^s \\ &\quad = \frac{L_{sd}}{\tau_r} i_{sd1}^s + \sigma_d L_{sd} \frac{di_{sd1}^s}{dt} + \omega_r k^2 \sigma_q L_{sq} i_{sq1}^s \end{aligned} \tag{21}$$

$$\begin{aligned} &\frac{d\phi_{sq1}^s}{dt} + \frac{1}{\tau_r}\phi_{sq1}^s - \omega_r \phi_{sd1}^s \\ &\quad = k^2 \frac{L_{sq}}{\tau_r} i_{sq1}^s + k^2 \sigma_q L_{sq} \frac{di_{sq1}^s}{dt} - \omega_r \sigma_d L_{sd} i_{sd1}^s \end{aligned} \tag{22}$$

where

$$\sigma_d = 1 - \frac{M_{srd}^2}{L_{sd} L_r} \quad \sigma_q = 1 - \frac{M_{srq}^2}{L_{sq} L_r} \quad \tau_r = \frac{L_r}{R_r}.$$

The vector model for the stator-flux control written for an arbitrary frame (denoted by the superscript a) using (22) and