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Keywords

<<Sensorless Control>>, <<Induction Motor>>, <<Torque Compensation>>, <<MRAS Speed Estimation>>.

Abstract

This paper addresses the problem of controlling a three-phase induction motor without speed measurements and with torque compensation. A nonlinear feedback linearization control was designed for a fifth order induction motor model including electrical and mechanical dynamics. The proposed control scheme, which is an indirect field oriented control technique (IFOC) based on feedback linearization control (FLC), is made up of flux-speed controllers. In order to provide an estimated rotor speed for the control system, a model reference adaptive system (MRAS) algorithm based on reactive power was applied. This estimation passes through a Kalman filter, which is also used for the torque estimation. Some experimental results with TMS 320F2812 DSP are provided to verify the proposed system performance in a low speed over load variation conditions.

Introduction

Induction motors are widely used in position and speed servomechanism, due to their simple structure, low cost and robustness. In addition, theirs drives, based on full digital control, have reached a consecrated technological status. Ongoing research has concentrated on elimination of the motor speed/position sensor without deteriorating the dynamic performance of the drive control especially with uncertain load torque. This elimination increases the robustness and reliability of the whole system and reduces its cost.

Excellent surveys on nonlinear schemes for induction motor control design have been proposed in the literature for IM sensorless servomechanism. One of the proposed nonlinear control methodologies is based on feedback linearization control (FLC) as firstly introduced by [8]. FLC provides regulation of rotor speed, rotor flux amplitude decoupling using only stator current measurements and load torque compensation. Although [8] was not a sensorless control scheme, many other papers following FLC methodology have dealt with the application of sensorless IM drivers, such as [2] [3] [4] and [9]. Their contribution has basically been to “simplify” FLC hypotheses, sensorless algorithm and control gain design.
The aim of this paper is to present the development of speed sensorless feedback linearization control of an induction motor using model reference adaptive control for speed estimation – MRAS. The proposed scheme is based on a current/flux and speed tracking controller. Only three-phase IM currents are measured for feedback, in order to provide rotor speed value, a MRAS algorithm is used for speed estimation. Thus, the control algorithm is able to track the reference speed and to compensate the value of the load torque estimated by using a Kalman filter.

This paper is organized as follows: first, the fifth-order IM model is presented, after which, the system control is defined and its structure is developed, which is composed of a current controller, speed controller and flux controller, is presented. Then, the design of speed estimation algorithm based on the reactive power – MRAS is illustrated and the Kalman filter for torque estimation is explained. Finally, experimental results are depicted, showing the performance of the proposed speed sensorless control.

**Induction Motor Model**

A three-phase N pole pair induction motor is expressed in stationary reference frame (q,d), considering a linear magnetic circuit and balanced operation condition, by the fifth-order model [7] [10], as

\[
\begin{align*}
\rho I_q &= -a_{11} I_q - \omega I_d + a_{13} a_{11} \lambda_d - a_{13} N \omega \lambda_q + a_{14} V_q, \\
\rho I_d &= -a_{14} I_q + \omega I_d + a_{13} a_{11} \lambda_d + a_{13} N \omega \lambda_q + a_{14} V_d, \\
\rho \lambda_q &= -a_{11} \lambda_q - (\omega_q - N \omega) \lambda_d + a_{13} L_m I_q, \\
\rho \lambda_d &= -a_{11} \lambda_q + (\omega_q - N \omega) \lambda_d + a_{13} L_m I_d, \\
\rho \omega &= \frac{T_e}{J} - \frac{T_L}{J}.
\end{align*}
\]

(1) 
(2) 
(3) 
(4) 
(5)

in which, \(\rho\) represents the time-derivative operator.

The stator currents \((I_q, I_d)\) and the rotor fluxes \((\lambda_q, \lambda_d)\) are the state variables, the stator voltages \((V_q, V_d)\) are the control input, the load torque \(T_L\) is known and constant, \(\omega\) is the rotor speed and \(\omega_s\) is the stationary speed.

Defining electric torque \(T_e\) as

\[
T_e = \mu \cdot J \cdot (\lambda_q I_q - \lambda_d I_d).
\]

(6)

Definitions above are related to the mechanical and electrical parameters as \(a_{01} \equiv \frac{L_m L_s - L^2}{a_0}, \quad a_{11} \equiv \frac{R_s}{L_s}, \quad a_{13} \equiv \frac{L_r}{a_0}, \quad a_{14} \equiv \frac{L}{a_0}\) and \(\mu \equiv \frac{N L_m}{J L_s}\).

Where \(R_s, R_r, L_s, L_r\) are the stator/rotor resistances and inductances, \(L_m\) is the magnetizing inductance \(J\) is the rotor inertia and \(N\) is the number of pole pairs.
**Induction Motor Control**

The speed tracking, flux regulation control problem under speed sensorless conditions is formulated considering the IM model (1)-(5) under the following conditions:

- Stator currents are measurable.
- Motor parameters are known and considered constant.
- Load torque is estimated and it is applied after motor flux excitation.
- Initial conditions of the IM state variables are known.
- \(\lambda^*\) is the flux constant reference value, \(\hat{\omega}\) is estimated speed, \(\omega_{ref}\) is the smooth reference bounded speed signal.

The FLC control scheme considers flux vector alignment on direct axis ‘d’, i.e., \(\lambda_{d} = \lambda^*\) and \(\lambda_{q} = 0\) as a condition of asymptotic field orientation [3]. FLC equations are developed considering fifth-order IM model under the assumption that estimated speed tracks real as could be acceptable replace measured speed with the estimated \(\omega = \hat{\omega}\).

The proposed control scheme of speed-flux sensorless controller is similar a feedback linearization control. This general structure is based on a combination of FLC and a proportional integral compensation (PI). Firstly, FLC control is designed exploiting classical cascade structure for inner current and speed/flux control approach. Then, PI compensation is implemented to guarantee speed tracking in range of speed operation.

The following definitions are used in control scheme: \(I_{q}^*, I_{d}^*, e_{\omega} = \hat{\omega} - \omega_{ref}, i_{p} = I_{q} - I_{q}^*, i_{d} = I_{d} - I_{d}^*, T_{L}\), which represent reference currents, estimated speed tracking error, stator/rotor current errors relating reference to measure values and estimated load torque (using a Kalman filter).

The proposed controller scheme is defined as follow:

**Current Controller**

The current controller is defined as

\[
V_{q} = u_{q} - u_{q}^* \tag{7}
\]

\[
V_{d} = u_{d} - u_{d}^* \tag{8}
\]

Considering indirect field orientation as a framework, the first part is derived from equations (1) and (2)

\[
u_{q} = \frac{1}{a_{d}} \left( a_{2} I_{q}^* + \omega I_{d}^* + a_{1} \lambda \right) + \rho I_{q}^* \tag{9}
\]

\[
u_{d} = \frac{1}{a_{d}} \left( a_{2} I_{d}^* - \omega I_{q}^* - a_{1} \lambda + \rho I_{d}^* \right) \tag{10}
\]

\[
\omega = N \hat{\omega} + \frac{a_{1} I_{d}^*}{\lambda} + I_{q}^* \tag{11}
\]

The voltage values of \(\overline{u_{q}}\) and \(\overline{u_{d}}\) are defined as \(\overline{u_{q}} = (k_p + k_n/s)\overline{i_{q}}\) and \(\overline{u_{d}} = (k_p + k_n/s)\overline{i_{d}}\), which is composed of a proportional integral compensation with fixed gains \(k_p\) and \(k_n\). These gains are determined considering a simplified induction motor electrical model.
Flux Controller

The flux controller is designed from (4):

\[ I_\alpha^* = \left( a_{11} \lambda + \rho \lambda \right) \frac{1}{a_{11} L_{\alpha}} \tag{12} \]

Speed Controller

Hence, from (5) is defined the speed controller

\[ I_{q^*} = i_q - \bar{I}_q \tag{13} \]

in which, \( i_q = \frac{1}{\mu \lambda} \left( \frac{T_e}{J} + \rho \omega_{ref} \right) \) and \( \bar{I}_q = \left( k_{p_{iq}} + k_{i_{iq}} \right) \omega_{ref} \).

The \( i_q \) current is composed of a proportional integral compensation with fixed gains \( k_{p_{iq}} \) and \( k_{i_{iq}} \).

These gain values are determined considering an induction motor mechanical model. Figure 1 illustrates the proposed control block diagram.

**Speed Sensorless Control Algorithm**

Considering the induction motor model expressed in a stationary frame, two independent observers are derived to estimate the components of the counter-EMF vectors. Thus, the two models of instantaneous reactive power could be defined as a reference model and as an adjustable model. [1]

The estimated speed is produced by a proportional integral adaptation mechanism error of both models. The MRAS system can be drawn as in figure 2.
Fig. 2: MRAS Rotor Speed Estimation

Where $e_m$, $q_m$, $i_m$, $\hat{e}_m$ and $\hat{q}_m$, are respectively the vectors of electromotive-force, reactive power, magnetizing current, estimated electromotive-force, estimated reactive power. The vectors $v_{\alpha\beta}$, $i_{\alpha\beta}$ represent stator voltage and current relative to the $\alpha\beta$ referential.

A modification of the original MRAS Speed Estimation [1], such as was proposed in [6], is included. This modification proposed the use of state variable filters to obtain the differentiation presented on MRAS equations. The filters should also be applied to voltage signals to avoid addition noise and phase delay among the vectors. These filters are design by the discretization of the transfer function given by

$$
\frac{v_{\alpha\beta}}{v_{\alpha\beta}} = \frac{i_{\alpha\beta}}{i_{\alpha\beta}} = \frac{\hat{i}_{\alpha\beta}}{\hat{i}_{\alpha\beta}} = G_f(s) = \frac{\omega^2}{(s + \omega)^2}.
$$

The filter frequency $\omega_c$ should be around five times around the input frequency of $v_{\alpha}$, $v_{\beta}$, $i_{\alpha}$ and $i_{\beta}$.

### Kalman Filter

The reduced mechanical IM system can be represented by the following equations

$$
\begin{align*}
\rho \begin{bmatrix} \omega \\ T_e \end{bmatrix} &= \begin{bmatrix} -\frac{B}{J} & -\frac{1}{J} \\ 0 & 0 \end{bmatrix} \begin{bmatrix} \omega \\ T_e \end{bmatrix} + \frac{1}{J} \begin{bmatrix} 1 \\ 0 \end{bmatrix} T_e \\
y &= [1 \ 0] \begin{bmatrix} \omega \\ T_e \end{bmatrix} \end{align*}
$$

The Kalman Filter could be used to provide the value of torque load or disturbances $- T_e$. Since this mechanical system (15)-(16) is nonlinear, the Kalman filter linearizes the model at the actual operating point. In addition, this filter has the property of taken into account the signal noise, which could be generated from the pulse width modulation drivers.

Assuming the definitions $x = [\omega \ T_e]^T$, $A_m = \begin{bmatrix} -\frac{B}{J} & -\frac{1}{J} \\ 0 & 0 \end{bmatrix}$, $B_m = \begin{bmatrix} 1/J \\ 0 \end{bmatrix}$ and $C_m = [1 \ 0]$.

Then, the recursive equation for the discrete time Kalman Filter [5] is described by
\[ K(k) = P(k)C_m^T \left( C_mP(k)C_m^T + R \right)^{-1} \]  

(17) 

Where \( K(k) \) is the Kalman gain.

The covariance matrix \( P(k) \) is given by

\[ P(k+1) = (I - A_{m,T}) (P(k) - K(k)C_mP(k)) (I - A_{m,T})^T + (B_{m,T}Q(B_{m,T})^T \]  

(18) 

So that, the observed states are

\[ \hat{x}_k(k+1) = (I - A_{m,T}) \hat{x}_k(k) + B_{m,T}u(k) + (I - A_{m,T}) K(k) \left[ y(k) - C_m \hat{x}_k(k) \right] \]  

(19) 

Where \( \hat{x}_k(k) = \begin{bmatrix} \omega_{kalman} \\ T_e \end{bmatrix}^T \), the matrices \( R \) and \( Q \) are defined according to noise elements of predicted state variables, taking into account the noise in the measured signals \( R \) and in the model parameter deviations \( Q \).

**Experimental Results**

Examples showing the performance attainable with the proposed controller are illustrated below.

Experimental results were carried out on a motor: 1.5 cv, 380 V, 2.56 A, 60 Hz, \( R_e = 3.24 \) Ω, \( R_r = 4.96 \) Ω, \( L_r = 404.8 \) mH, \( L_s = 402.4 \) mH, \( L_m = 388.5 \) mH, \( N = 2, 188 \) rad/s. (Produced by Industries WEG S.A.)

The operation sequences, reported in figure 3, are as follow:

1) The motor is excited during an initial time interval (around of 7 s) using a smooth flux reference trajectory. The steady-state flux value is 0.388 Wb.
2) Starting, from zero initial value, the speed reference grows linearly until it reaches the reference value. Thus, the reference speed value is kept constant.
3) At time \( t = 21 \) s a constant load torque is applied.

Figure 3 and figure 4 show experimental results considered reference speed value of 18 rad/s and 45 rad/s respectively. These figures depicted speed measure \( \omega \) by an external encoder, estimated MRAS speed \( \omega_{mras} \), estimated torque (via Kalman filter) and measure \( I_s \) and \( I_p \) stator motor currents.

It can be seen from the experimental results that the dynamic performance of the system attain satisfactory. The system tracks a reference speed value even upon application of torque disturbances. Based on the speed figures, there appears to be an error between the encoder speed measurement and estimated MRAS speed, which presents a constant value with some oscillation during torque and speed transients. This error may have been due to instrumentation hardware system (such as measure of stator currents) or to parametric model variations.
Conclusion

The sensorless IM control scheme proposed was developed based on nonlinear control – FLC and is composed by flux-speed controller, derived from fifth order IM model. To provide speed value for control feedback, it involved a rotor speed observer using an algorithm MRAS based on reactive power. Experimental results in DSP TMS 320F2812 platform depicted proposed control performance.

In accordance with expectative, the proposed control scheme presents satisfactory performance at low speed and over load variation conditions. An error between measure speed and estimated speed was observed at both reference speeds. This error involved a constant steady-state value and could be corrected inserting an offset in the control feedback. Future efforts will be devoted to avoid this speed error as well as to improve robustness against variations in model parameters.

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Fig. 4: Experimental Results: $\omega_{ref} = 45$ rad/s

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