Sensorless Direct FOC with Takagi-Sugeno Fuzzy Controller for Three-Phase Induction Motor

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Abstract—This paper proposes a Takagi-Sugeno (T-S) Fuzzy controller to substitute the two PI current regulators present in conventional sensorless direct field oriented control (DFOC) scheme. The proposed controller calculates the reference quadrature components of the stator voltage vector in rotor-flux-oriented reference frame. The rule base for the proposed controller is defined in function of the stator current direct and quadrature component errors. Constant switching frequency and low torque ripple are obtained using space vector modulation technique. Performance of the proposed sensorless DFOC scheme with T-S fuzzy controller is analyzed in terms of rise time, settling time and torque ripple considering different operating conditions. The fast torque response and low torque ripple are obtained with the proposed scheme as is shown through numerical simulation results.

I. INTRODUCTION

The three-phase induction motors (IM) are widely used in industrial application, because of their simple and robust structure, higher torque-to-weight ratio, higher reliability and ability to operate in dangerous environment. However, because of the coupling between torque and flux, unlike DC motor, their control is a challenging task. A classical method of induction motor control is the field oriented control [1]. It leads to decoupling between the flux and torque, thus, resulting in improved dynamic response of torque and speed.

In general the use of fuzzy systems does not require the accurate mathematical model of the process to be controlled. Instead, it uses the experience and knowledge of the involved professionals to construct its control rule base. The fuzzy logic controller (FLC) to be investigated is the Sugeno’s type [2], although there exist other types, for example, the Mamdani’s [3] and the Yamakawa’s [4].

The fuzzy controllers has proved powerful in the power electronics area and control of electric machines as shown in various articles in the literature, e.g., in [5] and [6] the fuzzy logic controller (FLC) for the speed control is implemented using the field oriented control technique as it provides better control of motor with high dynamic performance. In [7], it is also proposed and another fuzzy speed controller and it is compared with a conventional PI controller, shown that this controller takes superior performance under various operating conditions, like step change in speed and torque reference.

Similarly, in [8], it is proposed the fuzzy speed controller, but this controller is applied in indirect field-oriented control scheme. This scheme was compared with two speed control techniques, scalar control and conventional indirect field-oriented control, showing its superiority especially with high dynamic disturbances.

In this paper is designed the T-S fuzzy controller to substitute the both PI current regulators present in conventional sensorless DFOC scheme. The T-S fuzzy controller calculates the quadrature components of the stator voltage vector represented in the rotor-flux-oriented reference frame. The rule base for the proposed controller is defined in function of the stator current direct and quadrature component errors. In this controller is used the trapezoidal and triangular membership functions for inputs fuzzification.

This paper is organized as follows. In section II the theoretical background of the three-phase induction motor model, and the direct field oriented control principle are presented. In section III the topology of the proposed sensorless control scheme is analysed and in section IV the proposed T-S fuzzy controller is described in details, mentioning different aspects of its design. Section V presents the simulations results of T-S fuzzy controller, and in the end, the conclusion is given in Section VI.

II. THEORETICAL BACKGROUND

A. Three-Phase Induction Motor Model

By the definitions of the fluxes, currents and voltages space vectors, the dynamical equations of the three-phase IM in stationary reference frame can be put into the following mathematical form [9]:

\[
\ddot{\psi}_s = R_s i_s + \frac{d\psi_s}{dt} 
\]  

\[
0 = R_r \ddot{i}_r + \frac{d\psi_r}{dt} - j\omega_r \psi_r 
\]  

\[
\ddot{\psi}_s = L_s \ddot{i}_s + L_m \dot{i}_r 
\]  

\[
\ddot{\psi}_r = L_r \ddot{i}_r + L_m \ddot{i}_s 
\]  

Where \( \ddot{\psi}_s \) is the stator voltage space vector, \( \ddot{i}_s \) and \( \ddot{i}_r \) are the stator and rotor current space vectors, respectively, \( \dot{\psi}_s \) and
\( \vec{\psi}_r \) are the stator and rotor flux space vectors, \( \omega_r \) is the rotor angular speed, \( R_s \) and \( R_r \) are the stator and rotor resistances, \( L_s \), \( L_r \) and \( L_m \) are the stator, rotor and mutual inductance, respectively.

The electromagnetic torque is expressed in terms of the cross-vectorial product of the stator and the rotor flux space vector.

\[
t_c = \frac{3}{2} P \frac{L_m}{L_r} \vec{\psi}_r \times \vec{i}_s \\
t_c = \frac{3}{2} P \frac{L_m}{L_r} (\psi_{rd} i_{sd} - \psi_{rq} i_{sq})
\]

Where \( P \) is a number of pole pairs, \( \psi_{rd} \) and \( \psi_{rq} \) are the quadrature components of the rotor flux, and \( i_{sd} \) and \( i_{sq} \) are the quadrature components of the stator current.

**B. Direct Field Oriented Control**

In the rotor-flux-oriented reference frame the quadrature component of the rotor flux disappears and a physically easily comprehensible representation of the relations between torque, flux and current components is obtained. This representation can be expressed in the following formula.

\[
\psi_{rd} = \frac{L_m}{1 + sT_r} i_{sd} \\
t_c = \frac{3}{2} P \frac{L_m}{L_r} \psi_{rd} i_{sq}
\]

Considering that \( \psi_{rd} = \psi_r \), we can rewrite this equation,

\[
t_c = \frac{3}{2} P \frac{L_m}{L_r} \psi_r i_{sq}
\]

Where \( s \) is a laplace operator, \( \psi_r \) is the rotor flux module, and \( T_r = L_r / R_r \) is a rotor time constant.

The equations (8) and (9) show that the component \( i_{sd} \) of the stator current can be used as a control quantity for the rotor flux \( \psi_{rd} \). If the rotor flux can be kept constant with the help of \( i_{sd} \), then the cross component \( i_{sq} \) plays the role of a control variable for the torque \( t_c \) [10].

**III. THE PROPOSED SENSORLESS FIELD ORIENTED CONTROL SCHEME**

Figure 1 shows the block diagram of the proposed sensorless DFOC scheme, this scheme only needs sense the DC link voltage and two phases of the stator currents of the three-phase IM to calculate the stator voltage, estimate the rotor flux and rotor speed. In the sensorless DFOC scheme the T-S fuzzy controller takes the direct component of the stator current error \( E_{i_{sd}} \) and the quadrature component of the stator current error \( E_{i_{sq}} \) as inputs. On the other hand, the stator voltage quadrature components are the outputs, these outputs are represented in the rotor-flux-oriented reference frame. In this paper the stator voltage to source the rotor flux estimation was calculated using the DC link voltage and the inverter switches states [11]. However, the voltage and current model of the three-phase induction motor are used to estimated the rotor flux [12].

**A. Stator Voltage Calculation**

The stator voltage calculation module use the DC link voltage \( V_{dc} \) and the inverter switch state \( (S_a, S_b, S_c) \) of the three leg two level inverter. The stator voltage vector \( \vec{u}_s \) is determined as in [11]:

\[
\vec{u}_s = \frac{2}{3} \left( (S_a - \frac{S_b + S_c}{2}) + j \frac{\sqrt{3}}{2} (S_b - S_c) \right) V_{dc}
\]

**B. Rotor Flux Estimation**

The rotor flux estimation is obtained from the stator flux, it is

\[
\vec{\psi}_s = \int (\vec{u}_s - R_s \cdot \vec{i}_s) dt
\]

The problem in this type of estimation is when in low speeds the back electromotive force (emf) depends strongly of the stator resistance, to resolve this problem is used the current model to improve the flux estimation as in [12]. The rotor flux represented in the rotor-flux-oriented reference frame is:

\[
\vec{\psi}_{rdq} = \frac{L_m}{1 + sT_r} (\psi_{rd} i_{sd} - \omega_r T_r \psi_{rq})
\]

Where \( T_r = L_r / R_r \) is the rotor time constant. In this reference frame \( \psi_{rq} = 0 \) and substituting this expression in the equation (12), it is:

\[
\psi_{rd} = \frac{L_m}{1 + sT_r} i_{sd}
\]

In the current model the stator flux is represented as:

\[
\vec{\psi}_s = \frac{L_m}{L_r} \vec{\psi}_r + \frac{L_s L_r - L_m^2}{L_r} \vec{\psi}_r
\]

Where \( \vec{\psi}_r \) is the rotor flux estimated in the equation (13).

The voltage model is based in the equation (1) and from there the stator flux in the stationary reference frame is:

\[
\vec{\psi}_s = \frac{1}{s} (\vec{\psi}_s - R_s \vec{i}_s - U_{comp})
\]

With the aim to correct the errors associated with the pure integration and the stator resistance measurement, the voltage model is adapted through the PI controller.

\[
U_{comp} = (K_p + K_i \frac{1}{s}) (\vec{\psi}_s - \vec{\psi}_r)
\]

The \( K_p \) and \( K_i \) coefficients are calculated with the recommendation proposed in [12]. The rotor flux \( \vec{\psi}_r \) in the stationary reference frame is calculated as:

\[
\vec{\psi}_r = \frac{L_r}{L_m} \vec{\psi}_s - \frac{L_s L_r - L_m^2}{L_m} \vec{\psi}_s
\]

The estimator scheme shows in Figure 2 works with a good performance in the wide range of speeds.
The rotor flux \( \psi_r \), output of the reference model, is calculated with the equation (17). The adjustable or adaptative model equation is simpler and is obtained from the current model of the machine equations in stationary reference frame [13] using stator currents and rotor angular speed

\[
\frac{d\psi_{ra}}{dt} = -\frac{1}{T_r} \dot{\psi}_{ra} + \frac{L_{mr}}{T_r} \dot{i}_s - \dot{\omega}_r \psi_{ra}^a
\]  

(18)

The superscript \( a \) denotes the stator flux calculated from the adaptative model. With the rotor flux estimation from two methods, the voltage model \( \psi_r \) (reference model) and the current model \( \psi_a \) (adaptative model), the rotor speed estimation \( \dot{\omega}_r \) can be calculated with a PI adaptation mechanism by

\[
\dot{\omega}_r = (K_p + \frac{K_i}{s}) e
\]  

(19)

where

\[
e = \psi_{rd}^a \psi_{rq} - \psi_{rq}^a \psi_{rd}
\]  

(20)

is the cross-error between the adjustable and reference models.

IV. TAKAGI-SUGENO FUZZY CONTROLLER

The first order Takagi-Sugeno Fuzzy controller proposed in this paper takes as inputs the direct component of the stator current error \( E_{i_{sd}} \) and the quadrature component of the stator current error \( E_{i_{sq}} \). Moreover, it takes as outputs the quadrature components of the stator voltage vector, represented in the rotor-flux-oriented reference frame. The first output \( (u_{sd}^*) \) is a linear combination of the inputs, similarly, the second output \( (u_{sq}^*) \) takes the similar linear combination used in the first output but with the coefficients interchanged as is shown in Figure 4.

All the Membership Functions (MF’s) used in the T-S fuzzy controller have triangular and trapezoidal shapes, as is shown in Figure 5 and Figure 6, because these functions are suitable for real-time operations [15].
The direct component of the stator voltage \( u_{sd}^* \) is determined by the rules of the form

\[
R_x : \text{if } E_{isd} \text{ is Isd and } E_{isq} \text{ is Isq then } u_{sd}^* = aE_{isd} + bE_{isq}
\]

However, the quadrature component of the stator voltage \( u_{sq}^* \) is determined by the rules of the form

\[
R_y : \text{if } E_{isd} \text{ is Isd and } E_{isq} \text{ is Isq then } u_{sq}^* = -bE_{isd} + aE_{isq}
\]

Where Isd = Isq = \{N, ZE, P\} are the fuzzy sets of the inputs and, \( a \) and \( b \) are coefficients of the first-order polynomial function typically present in the consequent part of the firs-order Takagi-Sugeno fuzzy controllers. For instance, when the consequent part of the rule is a real number it is a zero-order controller but If the consequent is a linear combination we have a first-order controller [16].

The **product** is the conjunction operator and the weighted average (**wtaver**) is the defuzzification method used to set the controller in the MATLAB fuzzy editor. The final T-S fuzzy controller was programmed in **C** to facilitate its future implementation in the digital signal processor (DSP) of Texas Instruments.

### V. Simulation Results

The simulations were performed using MATLAB simulation package which include Simulink block sets and fuzzy logic toolbox. The switching frequency of the three-phase two level inverter was set to be 10kHz, the direct component of the stator current (\( i_{sd}^* \)) was set to be 0.77 pu and the coefficients considered in the T-S fuzzy controller were \( a = 5.3 \) and \( b = 0.5 \). The motor parameters are given in Table I. In order to investigate the effectiveness of the proposed control system and in order to check the closed-loop stability of the complete system, we performed several tests at different dynamic operating conditions.
in this paper. All the test results show the good performance of the proposed sensorless Direct FOC scheme with T-S fuzzy controller.

TABLE I
THREE-PHASE INDUCTION MOTOR PARAMETERS

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated voltage (V)</td>
<td>220/60 Hz</td>
</tr>
<tr>
<td>Rated Power (HP)</td>
<td>1.5</td>
</tr>
<tr>
<td>Rated Torque (Nm)</td>
<td>6.1</td>
</tr>
<tr>
<td>Rated Speed (rad/s)</td>
<td>180/12</td>
</tr>
<tr>
<td>$R_s$, $R_r$ (Ω)</td>
<td>5.56, 4.25</td>
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<tr>
<td>$L_s$, $L_r$ (H)</td>
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<tr>
<td>$L_m$ (H)</td>
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</tr>
<tr>
<td>$J/K_g m^2$</td>
<td>0.02</td>
</tr>
<tr>
<td>P (pole pairs)</td>
<td>2</td>
</tr>
</tbody>
</table>

VI. CONCLUSION

In this paper was proposed the T-S fuzzy controller to substitute the two PI current regulator present in conventional sensorless DFOC scheme. This controller calculates the quadrature components of the reference stator voltage vector in rotor-flux-oriented reference frame, the direct component is a linear combination of the fuzzy controller inputs, and the quadrature component use the similar linear combination but with the coefficients interchanged, not to be necessary another different coefficients values. The low torque ripple and constant switching frequency were obtained using space vector modulation technique. Numerical simulations results at different operating conditions verify that the proposed sensorless DFOC scheme with T-S fuzzy controller achieved good performance measures such as rise time, settling time and torque ripple as expected. These results validate the proposed scheme.

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REFERENCES


