ADAPTIVE FUZZY SLIDING-MODE CONTROL OF SPEED SENSORLESS UNIVERSAL FIELD ORIENTED INDUCTION MOTOR DRIVE WITH ON-LINE STATOR RESISTANCE TUNING*

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Abstract— In this paper, a speed sensorless induction motor drive is introduced which is direct vector controlled in a universal field-oriented (UFO) reference frame. This chosen reference frame is easily linked with direct and indirect rotor, stator and air gap field orientation control schemes of the induction machine (IM) drives using a stator to rotor virtual turn ratio. Based on partial feedback linearization theory, a nominal speed controller is designed first, regardless of the system parameter uncertainties and external load torque disturbance. Then, in the perturbed condition, a so-called totally invariant sliding mode (SM) controller is employed that has low SM chattering with no reaching phase. SM controller ensures the sliding motion through the entire drive system state trajectory. In addition, a fuzzy-adaptive algorithm is used to estimate the optimum upper bound of the motor lumped uncertainty function due to the rotor moment of inertia, friction coefficient and external load torque disturbance. As a result, the proposed controller combines the merits of fuzzy inference mechanism and the adaptive algorithm. Furthermore, a model reference adaptive system (MRAS) is derived to estimate the rotor speed and a simple method is also proposed for on-line estimation of the stator resistance. Finally, the effectiveness of the proposed speed controller scheme is verified by computer simulation.

Keywords— Induction motor, sliding mode, speed controller, sensorless, stator resistance

1. INTRODUCTION

Induction machine drives controlled by the field-orientation technique have been employed in high performance industrial applications. The field-orientation control scheme enables the control of the induction motor in the same way as a separately excited dc motor.

The two major methods of high performance control of induction motor drives are the direct field oriented control (DFOC) and the indirect field oriented control (IFOC). The IFOC is generally called slip frequency control, while in a DFOC system, the flux angle is employed directly in decoupling the torque and flux components [1-2].

The principle of a universal field oriented (UFO) controller for induction machine drives was proposed by Dr. De Doncker [3-4], and is compatible with direct and indirect rotor, stator and the air gap flux field-oriented control schemes of induction machine drives.

The main drawback of induction motor field oriented systems is that the performance of these systems will be degraded due to the motor parameters uncertainties, as well as due to the external load torque disturbance. Hence, many researchers have attempted to propose various methods to solve this

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problem [5-7]. The sliding mode control has been known as one of the most effective methods to overcome the mentioned problems [7, 8].

The variable structure control (VSC) strategy, using the sliding-mode control, has received much attention in recent years for controlling ac servo drive systems [5, 7, 8].

In general, the design of an SM controller can be divided into two phases: The SM reaching (or hitting) phase and the SM control. Before the system reaches the switching surface, the system structure is switched when the system state crosses the predetermined hyper-plane, so that the state slides along the reference trajectory. Once the controlled system’s state enter the SM, the dynamics of the system are determined by the choice of the sliding hyper plane and are immune to the system parameter uncertainties and disturbances. In real control systems such as ac motor drive systems, the SM switching rate should be limited around the sliding line, and limiting in the cycle occurs even in the steady-state. It may be noted that a high SM switching rate causes SM chattering, which has harmful effects on motor drive system performance such as torque pulsations, current harmonics and acoustic noise, etc.

A so-called totally invariant structure SM controller has been described in [7] which has no reaching phase and produces low SM chattering. In general SM control, the upper bound of uncertainties which include parameter variations and external load torque disturbance, must be available. However, the bound of the uncertainties is difficult to obtain in advance for practical applications. If the SM control gain is chosen to be much greater than the upper bound of lumped uncertainty, the system control will be stable, however the steady state error will exist [7, 8].

For SM control to gain less than the upper bound of lumped uncertainty, the system control will not be stable and robust against the uncertainties and external disturbances. To overcome this problem an adaptive fuzzy SM controller (AFSC) is used to estimate the optimum upper bound of lumped uncertainties [8]. Using the sliding surface in [7] and the UFO control approach presented in [3], this paper addresses the following research work.

i) A speed sensorless DFOC induction motor drive system is introduced which is field-oriented in an arbitrary reference frame proposed by Dr. De Doncker [3-4].

ii) Based on partial feedback linearization theory, a nominal speed controller is designed first, regardless of the motor parameters uncertainties and the unknown load torque. Then, in the perturbed system condition when uncertainties occur, a so-called totally invariant sliding mode structure controller is employed which guarantees the total sliding motion of the induction motor drive, as well as assures the perturbed drive system with the exact performance designed by the nominal controller.

iii) An adaptive fuzzy algorithm is derived to estimate the optimum upper bound of system lumped uncertainty [8].

iv) A MRAS- based estimator is used to estimate the rotor speed. One may note that the stator resistance error has great impact on the performance of a MRAS-based speed estimator, especially in a low speed operating region [13, 14]. To overcome this problem, a simple method is proposed for online estimation of this parameter.

v) A step trapezoidal PWM voltage source inverter (VSI) is used to feed the IM drive, which compared to SPWM VSI has a dc link utilization factor about (17.3) percent higher. In addition, it produces fewer torque pulsations [11]. The remainder of this paper is organized as follows:

In section II, the system dynamic model of the UFO IM drive is described. In section III, an adaptive fuzzy sliding mode speed controller is introduced. The rotor speed and stator resistance estimations are discussed in section IV. In section V, the parameter sensitivity of the sensorless UFO controller is presented. System simulation is shown in VI and finally in section VII, conclusions are verified.
2. DYNAMIC MODEL OF UFO INDUCTION MOTOR DRIVE SYSTEM

a) Basic theory

Figure 1 illustrates the stator \( D_{s} - Q_{s} \) : axes, the rotor \( d_{r} - q_{r} \) : axes and an arbitrary rotating reference frame which is defined by \( d_{i} - q_{i} \) : axes.

\[
v_{sd} = R_{s}i_{sd} + \frac{d\Psi_{sd}}{dt} - \omega_{i}\Psi_{sq} \tag{1}
\]

\[
v_{sq} = R_{s}i_{sq} + \frac{d\Psi_{sq}}{dt} + \omega_{i}\Psi_{sq} \tag{2}
\]

\[
0 = R_{r}i_{rd} + \frac{d\Psi_{rd}}{dt} - (\omega_{i} - \omega_{r})\Psi_{rq} \tag{3}
\]

\[
0 = R_{r}i_{rq} + \frac{d\Psi_{rq}}{dt} + (\omega_{i} - \omega_{r})\Psi_{rd} \tag{4}
\]

Fig. 1. Representation of an arbitrary reference frame

The stator and rotor flux linkages are

\[
\Psi_{sd}^{i} = L_{s}i_{sd}^{i} + L_{h}i_{rd}^{i} \tag{5}
\]

\[
\Psi_{sq}^{i} = L_{s}i_{sq}^{i} + L_{h}i_{rq}^{i} \tag{6}
\]

\[
\Psi_{rd}^{i} = L_{r}i_{rd}^{i} + L_{h}i_{sd}^{i} \tag{7}
\]

\[
\Psi_{rq}^{i} = L_{r}i_{rq}^{i} + L_{h}i_{sq}^{i} \tag{8}
\]

With

\[
L_{s} = L_{h} + L_{ls}, \quad L_{r} = L_{h} + L_{lr} \tag{9}
\]

All parameters are referred to the stator, and indices \( s \) and \( r \) show the stator and rotor quantities, respectively. Also, the induction machine torque and the mechanical load equations are given by

\[
T_{e} = \frac{3P}{2}(\Psi_{sd}^{i}i_{sq}^{i} - \Psi_{sq}^{i}i_{sd}^{i}) \tag{10}
\]

\[
J\frac{d\omega_{m}}{dt} = T_{e} - T_{L} \tag{11}
\]

Where \( R_{s} \) : Stator resistance, \( R_{r} \) : Rotor resistance, \( L_{h} \) : Magnetizing inductance, \( L_{r} \) : Rotor inductance, \( L_{S} \) : Stator inductance, \( L_{ls} \) : Stator leakage inductance, \( L_{lr} \) : Rotor leakage inductance, \( P \) : Pole pair number, \( J \) : Moment of inertia, \( \omega_{m} \) : Mechanical rotor angular velocity, \( \theta_{i} \) : Electrical rotor displacement, \( \theta_{e} \) : Electrical angle displacement between reference frame \( i \) and the two axis stationary reference frame and \( \omega_{i} \) is the electrical synchronous speed and \( T_{L} \) : Load Torque.
Considering a virtual stator to rotor turn ratio defined by (a), Eqs. (2), (3), (7) and (8) are referred to the stator to obtain [3]

\[
0 = R_r^i i_{rd}^i + \frac{\partial \Psi_{rd}^i}{\partial t} - (\omega_1 - \omega_r) \Psi_{rq}^i \\
0 = a^2 R_r^i i_{rq}^i + \frac{\partial \Psi_{rq}^i}{\partial t} + (\omega_1 - \omega_r) \Psi_{rd}^i
\]

With

\[
\Psi_{rd}^i = (a^2 L_r - aL_h) i_{rd}^i + aL_h (i_{sd}^i + i_{rd}^i) \\
\Psi_{rq}^i = (a^2 L_r - aL_h) i_{rq}^i + aL_h (i_{sq}^i + i_{rq}^i) \\
\Psi_{ad}^i = aL_h (i_{sd}^i + i_{rd}^i) \\
\Psi_{aq}^i = aL_h (i_{sq}^i + i_{rq}^i)
\]

Where \( R_r^i = a^2 R_r \), \( \Psi_{rd}^i = a \Psi_{rd}^i \), \( \Psi_{rq}^i = a \Psi_{rq}^i \), \( i_{rd}^i = \frac{i_{rd}}{a} \), \( i_{rq}^i = \frac{i_{rq}}{a} \).

Equations (5) and (6) are rewritten as

\[
\Psi_{sd} = (L_s - aL_h) i_{sd} + \Psi_{ad} \\
\Psi_{sq} = (L_s - aL_h) i_{sq} + \Psi_{aq}
\]

In order to link \( d_i \) axis of the arbitrary reference frame to a flux vector \( \varphi_a \), one usually states that the \( q_i \) component of this flux vector equals zero or in other words from (16) and (17), \( \varphi_{aq} = 0 \), \( \varphi_{ad} = \varphi_a \). For example, in the classical theory of the IM rotor flux field oriented (RFO) controller, one states that the component of the rotor flux vector \( \varphi_r \) equals zero. This means that the RFO controller derived from the machine equations automatically uses the rotor flux as the reference vector for decoupling the circuit. In steady-state conditions, for different values of parameter “a” given in Table 1, the flux vector “\( \varphi_a \)” can be linked with the well known specific synchronous frames in which the direct and indirect field orientation of the IM drive is most likely to be performed. These are termed the RFO, the air gap flux (AFO) and the stator flux field (SFO) reference frames.

\[\text{b) Calculations of flux vector } \varphi_a\]

Consider (1), (2), (18) and (19) be transformed from the \( d_i - q_i \) axes reference frame on the stator D-Q axes reference frame. The \( D_s \) axis and \( Q_s \) axis components of the stator flux linkages at the K-th sampling instant can be derived as:

\[
\varphi_{sd}^s (k) = \varphi_{sd}^s (k-1) + (v_{sd}^s (k-1) - R_s i_{sd}^s (k-1)) T_s \\
\varphi_{sq}^s (k) = \varphi_{sq}^s (k-1) + (v_{sq}^s (k-1) - R_s i_{sq}^s (k-1)) T_s
\]

Where \( T_s \) is the sampling interval and the variables with subscript “k-1” are the previous samples.

Also

\[
\varphi_{sd}^s = (L_s - aL_h) i_{sd}^s + \varphi_{ad}^s \\
\varphi_{sq}^s = (L_s - aL_h) i_{sq}^s + \varphi_{aq}^s
\]
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\[ \varphi^s_{ad} = aL_h(i^s_{ad} + i^s_{rd}) \]  
\[ \varphi^s_{aq} = aL_h(i^s_{aq} + i^s_{rq}) \]

where

\[ i^s_{rd} = \frac{i^s_{rd}}{a}, \quad i^s_{rq} = \frac{i^s_{rq}}{a} \]  

At kth sampling instant, \( \varphi^s_{ad}(k) \) and \( \varphi^s_{aq}(k) \) are obtained from (20) and (21). Once the components of the stator flux linkage are obtained, then \( \varphi_{ad}(k) \) and \( \varphi_{aq}(k) \) are calculated from (22) and (23). At kth sampling instant, the magnitude and position of flux vector \( \varphi_a \) are obtained as

\[ \varphi_{ak}(k) = \sqrt{\varphi^2_{ad}(k) + \varphi^2_{aq}(k)} \]  
\[ \rho_t(k) = \arctan \frac{\varphi^2_{aq}(k)}{\varphi^2_{ad}(k)} \]

Also, the electrical angular speed of rotating flux vector \( \varphi_a \), or in other words the speed of arbitrary reference frame is:

\[ \omega_t(k) = \frac{\rho_t(k) - \rho_t(k-1)}{T_s} \]

<table>
<thead>
<tr>
<th>Turn ratio</th>
<th>( \Psi_a )</th>
<th>Flux selected</th>
</tr>
</thead>
<tbody>
<tr>
<td>( a = \frac{L_h}{L_r} )</td>
<td>( \Psi_r )</td>
<td>Rotor flux</td>
</tr>
<tr>
<td>( a = 1 )</td>
<td>( \Psi_h )</td>
<td>Air gap flux</td>
</tr>
<tr>
<td>( a = \frac{L_s}{L_h} )</td>
<td>( \Psi_s )</td>
<td>Stator flux</td>
</tr>
</tbody>
</table>

c) Design of a decoupler for DUFO system

Linking (14-17) and (20) gives

\[ \Psi'_{rd} = (a^2L_r - aL_h)[\frac{\varphi_{ad}}{aL_h} - i_{sq}] + \Psi_{ad} \]  
\[ \Psi'_{rq} = -(a^2L_r - aL_h)i_{sq} \]

Combining (12) and (13) with (22) and (23)

\[ \omega_{sl} = (\omega_r - \omega_p) = \frac{(1 + \sigma_{st}T_s p)i_{sq}}{\frac{T_r \Psi_{ad}}{aL_h} - \sigma_{st}T_s i_{sd}} \]  
\[ (1 + \sigma_{st}T_s p)i_{sd} = \frac{1}{aL_h}(1 + T_s p)\Psi_{ad} + \omega_s \sigma_{st} T_s i_{sq} \]
Where $T_r = \frac{L_r}{R_r}$, $\frac{d}{dt}$, $\omega_{sl}$ is the rotor slip angular velocity and $\sigma_{rl} = \frac{aL_r - L_h}{aL_r}$ denotes the rotor leakage inductance coefficient. Equation (33) shows that a coupling exists between the stator current component $i_{sq}$ and the UFO flux orientation $\Psi_{ad}$. Therefore, any change in $i_{sq}$ will cause a transient in $\Psi_{ad}$. To solve this problem, a decoupler is designed in the following way:

Assume that

$$i_{sd} = \tilde{i}_{sd} + i_{dc}$$  \hspace{1cm} (34)

Substituting (34) into (33) yields

$$(1 + \sigma_{rl}T_r p)i_{sd} - \frac{1}{aL_h}(1 + T_r p)\Psi_{ad} + (1 + \sigma_{rl}T_r p)i_{dc} - \omega_{sl}\sigma_{rl}T_r i_{sq} = 0$$  \hspace{1cm} (35)

In order to decouple $\Psi_{ad}$ from $i_{sq}$, it is required to have

$$(1 + \sigma_{rl}T_r p)i_{dc} - \omega_{sl}\sigma_{rl}T_r i_{sq} = 0$$  \hspace{1cm} (36)

and

$$i_{dc} = \frac{\omega_{sl}\sigma_{rl}T_r i_{sq}}{(1 + \sigma_{rl}T_r p)}$$  \hspace{1cm} (37)

### 3. PROPOSED SPEED CONTROLLER DESIGN

**a) Nominal controller**

Considering Eqs. (11) and (21), assuming a zero disturbance load torque and the motor nominal parameters, the mechanical Equation of the UFO IM drive becomes

$$\frac{d\omega_n}{dt} = -\frac{B_n}{J_n} \omega_n(t) + \frac{K_f}{J_n} i_{sq}^* = A_{pn}\omega_m(t) + B_{pn} U(t)$$  \hspace{1cm} (38)

Where $A_{pn} = -\frac{B_n}{J_n}$, $B_{pn} = \frac{K_f}{J_n}$, $K_f = \frac{3}{2} p\Psi_a^*$ and $U(t)$ is the control effort.

Note that the superscript “*” represents the reference commands and subscript “n” represents the nominal stator or rotor quantities. With reference to [10], using the partial feedback linearization theory, a nominal speed controller is designed as

$$U_n = -B_{pn}^{-1} A_{pn} \omega_m + B_{pn}^{-1} [\dot{\omega}_m^* - k_v e]$$  \hspace{1cm} (39)

Substituting (38) into (39), the speed equation error is

$$\dot{e} + k_v e = 0$$  \hspace{1cm} (40)

Where $e = \omega_m - \omega_m^*$.

From (40), the desired speed error dynamic can be achieved by adjusting the parameter $k_v$.

The above nominal controller cannot guarantee the robust and stable performance of the UFO IM against the load torque disturbance and uncertainties that exist in the mechanical parameter (J, B). This problem can be solved by employing a so-called totally invariant structure SM controller, which has no reaching phase and produces small SM chattering.
b) Sliding-mode controller

Consider the unpredictable perturbation effect due to the unknown external load torque and variations that exist in \( (J, B) \). Using Eqs. (30), the perturbed system dynamic is given by

\[
\frac{d\omega_m}{dt} = (A_{pn} + \Delta A)\omega_m(t) + (B_{pn} + \Delta B)U(t) + (D_{pn} + \Delta D)T_i
\]

Where \( D_{pn} = -\frac{1}{J} \) and \( P(t) \) is called the lumped uncertainty defined by

\[
P(t) = \Delta A\omega_m(t) + \Delta B U(t) + (D_{pn} + \Delta D)T_i
\]

It is assumed that \( P(t) \) is bounded such that

\[
|P(t)| < \rho
\]

Where \( \rho \) is a positive constant value.

Now, assume the following sliding surface

\[
S(t) = C(e) - C(e_0)e_0 - \int_0^t \frac{\partial C}{\partial e} A e d\tau
\]

Where \( C = B_{pn}^{-1}e \), \( A = -k_v \) and \( e_0 \) is the initial speed error.

From (36), \( S(0) = 0 \) and also

\[
\dot{S}(t) = \frac{\partial C}{\partial e} \dot{e} - \frac{\partial C}{\partial e} Ae = B_{pn}^{-1}(\dot{e} + k_v e) = 0 \quad \text{for} \quad t > 0
\]

As a result, \( S(t) = 0 \) for all \( t \geq 0 \). This means that the sliding surface, regardless of the system uncertainties and disturbance load torque, has no reaching phase as in the traditional sliding-mode control [10].

When uncertainties occur, from (41) and (44) one can obtain that

\[
\dot{S}(t) = B_{pn}^{-1} P(t) \quad \text{for} \quad t > 0
\]

In addition, \( S(t) \neq 0 \) for \( t > 0 \).

In such a condition, the nominal controller \( U_n \) cannot satisfy the perturbed system dynamic shown in (41), and therefore the speed controller will not be capable of preserving the drive system robustness and stability. To overcome this drawback, an additional curbing controller is designed to ensure that for all time \( t > 0 \), the motor drive system dynamics curbs onto the \( S(t) = 0 \), making the states of the induction motor follow the nominal trajectories, controlled by the nominal controller shown in (39). This controller is chosen as [7]

\[
U_b = -\rho B_{pn}^{-1} \text{sgn}(S(t))
\]

Combining the curbing controller \( U_b \) with the nominal controller \( U_n \) described by (31), the total system control input is obtained as

\[
U = -B_{pn}^{-1} A_{pn} \omega_m + B_{pn}^{-1} [\omega_m^* - k_v e] - \rho B_{pn}^{-1} \text{sgn}(S(t))
\]

Such a control input signal has considerable SM chattering. One easy way to solve this problem may be by replacing the sign function \( \text{sgn}(S) \) for a boundary layer saturated function \( \text{sat}(\frac{S}{\phi}) \) [10].
c) Adaptive fuzzy sliding mode controller

An adaptive fuzzy SM control system which combines the merits of the SM control of [7], the fuzzy inference mechanism and the adaptive algorithm is described in [8]. The adaptive fuzzy estimator is used to detect the upper bound of lumped uncertainty shown in (42) in order to reduce the SM controller gain which is defined by parameter $\rho$.

The fuzzy inference mechanism is improved by adapting the centers of the membership functions to estimate the optimal bound of lumped uncertainty. Compared to a conventional estimator, the fuzzy-inference mechanism uses prior expert knowledge to accomplish control object more efficiently. Hence in this paper, a fuzzy algorithm is employed to estimate the upper bound of lumped uncertainty.

The block diagram of Fig. 2 is proposed to obtain the optimum value for the SM controller gain. In this figure, (s.f) denote the fuzzy scaling factors.

![Fig. 2. Block diagram of fuzzy control gain estimator](image)

Also, the membership functions of Fig. 3 are used for the fuzzy sets of sliding surface $S, \dot{S}$ and control gain ($\rho$).

![Fig. 3. Membership functions of fuzzy sets](image)

Moreover, the fuzzy sets corresponding to fuzzy control rules are

N: negative  Z: zero  P: positive

NH: negative huge, NB: negative big, NM: negative medium, NS: negative small, PH: positive huge, PB: positive big, PM: positive medium, PS: positive small, ZE: zero. From Fig. 3, the three fuzzy subsets, N, Z, and P are used for defining $S$ and $\dot{S}$. Therefore, nine fuzzy inference rules result and are given in Table 2. Fuzzy output can be calculated by the center of area (COA) defuzzification as [8].

$$\rho = \frac{\sum_{i=1}^{9} w_i c_i}{\sum_{i=1}^{9} w_i} = \frac{[w_1 \cdots w_9]}{[w_1 \cdots w_9]} = \nu^T W$$  \hspace{1cm} (50)
Where \( \nu = [c_1 \ldots c_9] \) is an adjustable parameter vector, \( c_1 \) through \( c_9 \) are the center of the membership functions of \( \rho \) and vector

\[
W = [w_1, \ldots, w_9] / \sum_{i=1}^{9} w_i
\]

is a firing strength vector. The adjustable parameter vector, \( \nu \), is obtained as

\[
\nu = \alpha B_{mn}^{-1} |S(t)| W
\]

Then, the upper bound of lumped uncertainty \( (\rho) \) estimated by (42) will be optimum. One may also be noted that in Eq. (44), \( \alpha \) is a positive constant value which is determined by a trial and error method. In addition, for an optimum value of \( \rho \), the minimum control effort \( (i_{rQ}^*) \) can be achieved. Furthermore, when the system parameter variations or the external disturbance occur \( (S(t) \neq 0) \), the value of \( P(t) \) will change; thereby the fuzzy inference mechanism and the adaptive law will be excited to find a new value for \( \rho \).

### Table 2. The resulting fuzzy inference rules

<table>
<thead>
<tr>
<th>S</th>
<th>( \dot{S} )</th>
<th>P</th>
<th>Z</th>
<th>N</th>
</tr>
</thead>
<tbody>
<tr>
<td>P</td>
<td>NH</td>
<td>NB</td>
<td>NM</td>
<td></td>
</tr>
<tr>
<td>Z</td>
<td>NS</td>
<td>ZE</td>
<td>PS</td>
<td></td>
</tr>
<tr>
<td>N</td>
<td>PM</td>
<td>PB</td>
<td>PH</td>
<td></td>
</tr>
</tbody>
</table>

### 4. PARAMETER ESTIMATION

#### a) Rotor speed estimator

The MRAS approach has the immediate advantage of the model being simple, very easy to implement and has direct physical interpretation. A well known model reference adaptive system (MRAS) based speed estimator is used for the UFO IM drive which has the following reference and adjustable models, [9]

**A:**

\[
\frac{d}{dt} \Psi_{rD}^s = \frac{L_r}{L_h} [v_{rD}^s - (R_s + \sigma \frac{d}{dt}) i_{rD}^s]
\]

\[
\frac{d}{dt} \Psi_{rQ}^s = \frac{L_r}{L_h} [v_{rQ}^s - (R_s + \sigma \frac{d}{dt}) i_{rQ}^s]
\]

**B:**

\[
\frac{d}{dt} \hat{\Psi}_{rD}^s = (-\frac{1}{T_r}) \hat{\Psi}_{rD}^s - \omega_r \hat{\Psi}_{rQ}^s + \frac{L_h}{T_r} i_{sD}^r
\]

\[
\frac{d}{dt} \hat{\Psi}_{rQ}^s = (-\frac{1}{T_r}) \hat{\Psi}_{rQ}^s + \omega_r \hat{\Psi}_{rD}^s + \frac{L_h}{T_r} i_{sQ}^r
\]

Where \( \sigma = L_s \frac{L_h^2}{L_r^2}, \hat{\Psi}_{rd}^s \) and \( \hat{\Psi}_{rq}^s \) are the rotor estimated flux linkages in the stator \( D_s - Q_s \) axes reference frame.

Using the above models, the block diagram of Fig. 4 illustrates the implementation of the MRAS-based speed estimator. As can be seen on Fig. 4, the error between rotor flux linkage estimation is used to drive a suitable adaptive mechanism, which generates the estimated rotor speed \( \hat{\omega}_r \).
It has been proved that if a proportional plus integral (PI) controller is used [9]

$$\dot{\omega}_r(t) = k_p e + k_I \int_0^t e \, d\tau$$  \hspace{1cm} (57)

Then, the observer, which is a nonlinear feedback system, is stable if

$$e = \Psi_{rQ}^s \dot{\Psi}_{rQ}^s - \Psi_{rD}^s \dot{\Psi}_{rD}^s$$  \hspace{1cm} (58)

**b) Stator resistance estimator**

The rotor time constant $\tau_r$ has a negligible effect on the accuracy of the MRAS-based speed estimator, while the stator resistance $R_s$ has great influence on the operation of this speed observer. This influence becomes more serious in the low speed region [13]. This occurs because the voltage drop on $R_s$ cannot be neglected in this operation region. To solve this problem, the following simple method is proposed for online estimation of $R_s$.

Referring to Fig. 5, a particular synchronous rotating reference frame is introduced in which its $q_e$ axis coincides with the stator current vector $I_s$. Using this reference frame, ignoring the deviative terms $\frac{d\Psi_{sd}^e}{dt}$ and $\frac{d\Psi_{sq}^e}{dt}$, Eqs. (1) and (2) become

$$v_{sd}^e = -\omega_{es} \Psi_{sq}^e$$  \hspace{1cm} (59)

$$v_{sq}^e = R_s i_{sq}^e + \omega_{es} \Psi_{sd}^e$$  \hspace{1cm} (60)

From (52), one can obtain

$$R_s = (v_{sq}^e - \omega_{es} \Psi_{sd}^e) / i_{sq}^e$$  \hspace{1cm} (61)

where $\omega_{es}$ is the angular speed of the stator current vector which in Fig. 5 is linked by $q_e$ axis.

Fig. 5. A synchronous rotating reference frame for stator resistance estimating.

**5. PARAMETER SENSITIVITY OF THE DUFO SYSTEM**

**a) Parameter sensitivity of the selected reference frame**

In a DUFO system, the choice of the reference frame has an important impact on the IM drive system performance. From (24) and (29), considering the parameter “a” given in Table 1, one can conclude that
the estimator stator flux is only sensitive to variation of the stator resistance, while the estimated air gap flux is parameter dependent on the stator resistance and stator leakage inductance. On the other hand, the estimation of the rotor flux requires the knowledge of the machine inductances, especially the total leakage inductance $\sigma L_s$.

Since the estimation of the stator flux is insensitive to the machine total leakage inductance, and since the variation of the stator leakage flux compared to the stator flux linkage is small, one can say that the steady state performance of the DSFO and DAFO systems are insensitive to the IM inductances. This means the DSFO and DAFO controllers do not require sophisticated parameter estimation methods or model-based observers to reduce robust and precise torque control [4]. It may be noted that from the view point of control theory, the performance of a feedback control system relies on the accuracy of the feedback signal. To obtain accurate and robust control characteristics, the most accurate signal should be chosen as the feedback signal. As mentioned earlier, the stator flux and air gap flux can be estimated more accurately than the rotor flux from the stator measured currents and voltages, so it is natural to use the stator or the air gap flux as the feedback signal, which leads to the implementation of DSFO or DAFO control systems.

In [4] it has been described that for a given stator flux the pullout torque in the IM DSFO or DAFO control systems exist. However, these pullout torques do not vary with errors in the estimated machine total leakage inductance. In addition, the torque limits are for most machines well above the rated torque (more than two times) and do not impose a significant restriction on the drive system. Moreover, since the sensitive elements, namely the leakage inductances, are in the decouplers of the DSFO and DAFO systems, the drive leakage inductance will affect only the transient response of these elements. But in the steady state, the detuning effect will be corrected by the feedback control of the estimated stator flux (for the DSFO system) and the estimated air gap flux for the DAFO systems, yielding accurate and robust control characteristics. The advantage of DSFO and DAFO controllers are more significant in the field weakening region, where the voltage is limited, hence both the stator flux and air gap flux are limited. In this case, the maximum torque capability of the DSFO and DAFO systems are approximately equal to that of a properly tuned DRFO system or greater than that of a detuned DRFO system [12].

In contrast to DSFO and DAFO systems, the DRFO controllers are sensitive to detuned machine leakage inductances, namely the total leakage inductance $\sigma L_s$. Clearly a perfect DRFO IM drive system has no pullout torque phenomenon. However, in the case of detuning (magnetic saturation, temperature and skin effects), this system can also experience stability problems similar to pullout torque, which is a function of the error in $\sigma L_s$. Obviously the larger the error in this parameter, the smaller the pullout torque. In a detuned DRFO system, the detuning pullout torque determines the boundary of static stability. If the torque command exceeds the pull out torque, the IM drive system will be statically unstable, and DAFO control systems compared to DRFO systems because of their precise and robust control, always retain maximum torque capability with a limited stator flux. A properly tuned DRFO controller can achieve the same maximum torque capability, but lower torque-ampere ratio results for reduced load torque. A detuned DRFO system has reduced maximum torque capability and a significantly lower torque-ampere ratio. Large detuning may even result in static instability. The steady state performance of the DSFO and DAFO are clearly superior to that of the DRFO system below the based speed at rated flux and above base speed in the field weakening region. However, in practice, the implementation of a DRFO system is simpler than the DSFO and DAFO control systems.
In the IM DRFO systems, it is required for the rotor actual flux to be the feedback control signal. The only solution is to use a flux observer to estimate this quantity. The observer needs to know the accurate values of the machine parameters and on the other hand, in estimation of these parameters, it is required to use the actual rotor flux. As a result, it is impossible or at least extremely difficult to detect the rotor flux and machine leakage inductances simultaneously.

One may note that in DRFO systems, operating below the base speed at the rated rotor flux, the machine inductances remain approximately constant [13-14]. However it is argued that the effects of load induced cross saturation (due to a non sinusoidal distribution) of the saturated flux, axial saturation and skew saturation are such that machine inductances, especially the magnetizing reactance, may have about 5% variation. Such a small variation cannot really affect the estimated rotor flux, hence in below the base speed, the effects of machine inductance errors have been few in practice, which is an encouraging result.

One can conclude that in DRFO systems, in above the base speed operating regions (in the field weakening region), using a machine saturation model is mandatory. In this paper we limit ourselves only in detecting the stator resistance and rotor speed simultaneously. It is worth mentioning that we have recently proposed a novel method for simultaneous estimation of the rotor flux, rotor speed, stator and rotor resistances, using the sliding mode observers.

**b) Parameter sensitivity of the MRAS-based speed estimator**

In [14] it has been shown that in a MRAS-based speed estimator, the estimator speed ripple is higher when the machine magnetizing reactance is overestimated. Therefore in UFO systems, a maximum of reduction in \( L_h \) below base speed and at the rated flux vector \( \varphi_\alpha \), has little effect on the performance of the MRAS-based speed estimator. In addition, the estimation effects of \( \pm 10\% \) variation in \( \sigma L_r \) and \( \sigma L_s \) are not observed in practice, perhaps because the errors in \( R_s \), albeit very small, are sufficient to mask them.

Moreover, since it is \( T_r \) or \( R_r \) that occurs in the adjustable model of the MRAS-based speed estimator, the effect of errors in \( R_r \) itself is perhaps rather superfluous. However in the previous reported research works, the predicted effects of \( R_r \) errors like the effect of \( \sigma L_r \) and \( \sigma L_s \), which have in practice been small, have been validated and is an encouraging result. The parameter errors which have substantial effects have been found to be \( R_s \).

**6. SYSTEM MODELING**

**a) System description**

The block diagram depicted in Fig. 6 shows the MRAS-based sensorless control of a DUFO IM drive, fed by a PWM VSI. With reference to this figure, the stator d-q axes reference voltages are obtained in the following way:

\[
\begin{align*}
V_{sd} &= \hat{V}_{sd} + V_{sd}^m, \\
V_{sq} &= \hat{V}_{sq} + V_{sq}^m
\end{align*}
\]

(62)

where \( \hat{V}_{sd} \) and \( \hat{V}_{sq} \) are the output signals of the d-q: axes PI controllers and \( V_{sd}^m \) and \( V_{sq}^m \) are the d-q: axes movement voltage signals which are produced in a block termed “movement voltage production block”.

The movement voltage terms are expressed as

\[
\begin{align*}
V_{sd}^m &= -\omega_e \phi_{sd}, \\
V_{sq}^m &= \omega_e \phi_{sq}
\end{align*}
\]

(63)
Linking (18), (19), (20) and (60) yields

\[ v_{sq}^m = -\omega_c (L_s - aL_h) i_{sq} \]  \hspace{1cm} (64)

\[ v_{sq}^m = \omega_c (L_s - aL_h) i_{sq} + \omega_c |\varphi_a| \]  \hspace{1cm} (65)

In the above equations, the stator measured voltages and currents and the flux vector \( \varphi_a \) can be obtained based on the theory described in the part b of section 2.

**b) Simulation results**

A C++ step by step computer program was developed to model the drive system control shown in Fig. 6. The flowchart of this program is shown in Fig. 7. In this program, the system equations are solved using a static runge-kutta fourth order method. Computer simulation results were obtained for a three-phase cage induction machine with parameters shown in Table 3. Using these parameters, the gains of conventional controllers and the gains of the AFSM controller were obtained by the trial and error method and shown in Table 4.

A simulation test was performed first, in order to obtain the IM drive performance in the DRFO, DSFO and DAFO reference frames. These results were obtained based on the motor nominal parameters (the conventional PI controllers were tuned to the motor nominal parameters), in the case of an exponential speed reference command from zero to 200 rad/sec (electrical) at 0 sec., a stepwise load torque from zero to 25n.m (motor rated torque) at \( t=1 \) sec., an exponential speed reference command from zero to \(-200\) rad/sec (electrical) at \( t=3 \) sec and a step down load torque from 25n.m to \(-25\) n.m at \( t=3.5 \) sec.

From these results, one can see that the direct vector control of the IM drive has been achieved in the stator, rotor and air gap field orientation reference frames. That is because in the IM steady state condition, the average value of \( \varphi_{aq} \) tends to be zero, besides the average value of flux vector \( \varphi_a \) accurately track the respected reference command signal \( \varphi_{ad}^* \).

In addition, the simulated results shown in Fig. 8 were obtained on the condition of an exponential reference command from zero to 150 rad/sec (electrical) at zero sec and rise up from zero to 150rad/sec at \( t=0.5 \) sec., a step up load torque from zero to 25 n.m and for \( J=5J_n \), \( B=5B_n \) at 0.5 sec. In this test, the speed PI controller was tuned to the motor nominal parameters. From these results one can see that while with AFSM controller a good disturbance rejection can be achieved, but with a PI controller, a speed overshoot of about 33.3 percent and a somewhat long settling time is obtained. Although the drive system will finally become stable, such a high speed overshoot needs a high motor accelerating torque, and as a result, a high motor inrush current which is certainly harmful in practice. It must be noted that it is not practical to tune the speed controller gains to the uncertain parameter variations and external load torque online.

Moreover, the third simulation test was carried out to show the impacts of the stator resistance variations on the performance of the MRAS-based sensorless control IM drive. As Figs. 1 and 2 show, for a 10 percent linear increase in the stator resistance (from 0.66\( \Omega \) to 0.72\( \Omega \)), a high frequency, high amplitude speed ripple is produced which can make the drive system unstable. Furthermore, the capability of the adaptive fuzzy inference mechanism is demonstrated by the simulated results shown in Fig. 11. Figure 11a shows that if at \( t=0.45\) sec. the value of SM controller gain is stepped down to half times its initial value (selected at \( t=0 \) sec.), by using the conventional SM controller the drive the system becomes...
unstable. Also in Fig. 11b, it is shown that at t=0.45 sec, the SM controller gain is stepped up to two times its initial value. With the conventional SM controller, the drive system becomes stable but steady and state error will exist in the rotor speed. In addition, Fig. 11c confirms that by using the AFSMC, the mentioned problems can be solved. Using the AFSMC, the SM controller gain “\( \rho \)” is optimally estimated based on upper bound lumped uncertainty, as previously described in part C of section 3.

Table 3. Induction Machine parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( L_m ) (mH)</td>
<td>58.2</td>
</tr>
<tr>
<td>( L_i ) (mH)</td>
<td>4.29</td>
</tr>
<tr>
<td>( R_s ) (Ω)</td>
<td>0.6592</td>
</tr>
<tr>
<td>( J ) (Kg-m²)</td>
<td>0.006</td>
</tr>
<tr>
<td>( f_n ) (Hz)</td>
<td>50</td>
</tr>
<tr>
<td>( V ) (L-L) (Volt)</td>
<td>250</td>
</tr>
</tbody>
</table>

Table 4. Coefficients of the system PI and FASM controllers

<table>
<thead>
<tr>
<th>Coefficient</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \phi = 0.7 )</td>
<td></td>
</tr>
<tr>
<td>Boundary layer width</td>
<td></td>
</tr>
<tr>
<td>( K_\rho = 50 )</td>
<td></td>
</tr>
<tr>
<td>( K_f = 5 )</td>
<td></td>
</tr>
<tr>
<td>( \alpha = 3 )</td>
<td></td>
</tr>
</tbody>
</table>

Fig. 6. Block diagram of the drive system control
Adaptive fuzzy sliding-mode control of speed sensorless…

Fig. 7. Computer flowchart

1. start
2. Give initial values and machine parameters
3. Rotor flux field orientation reference frame
4. Choose field orientation reference frame
5. Stator flux field orientation reference frame
6. Air gap flux field orientation reference frame
7. Flux orientation estimator
8. Stator resistance estimator
9. Rotor speed estimator
10. Calculate the stator reference currents in the UFO reference frame
11. Calculate the stator reference voltages in the UFO reference frame
12. Model the PWM-VSI
13. Solve the system equations
14. If time exceeding
15. Yes
16. End
17. No

Fig. 8. Rotor speed responses with PI and AFSM controllers due to parameter uncertainties and disturbance load torque

a) Rotor speed with PI controller
b) Rotor speed with AFSM controller
Fig. 9. Machine drive performance in the UFO reference frame with AFSMC

a) Reference and estimated speeds

b) Estimated speed error

c) Motor electromagnetic developed torque

d) Stator phase (A) current

e) Rotor flux vector component in the q axis of DRFO reference frame

f) Rotor flux vector component in the d axis of DRFO reference frame

g) Magnetizing flux vector component in the q axis of DAFO reference frame

h) Magnetizing flux vector component in the d axis of DAFO reference frame

i) Stator flux vector component in the q axis of DSFO reference frame

j) Stator flux vector component in the d axis of DSFO reference frame
Adaptive fuzzy sliding-mode control of speed sensorless…

7. CONCLUSIONS

The DFO of a speed sensorless induction machine drive system has been presented in a universal field orientated reference frame. The chosen reference frame can be easily linked with the well-known stator, rotor and air gap field orientated control schemes of the induction motor drives. A totally invariant structure sliding mode speed controller has been employed that has no reaching phase and produces minimum SM chattering. The SM controller makes the states of the induction motor follow its nominal trajectories, obtained by a nominal LQ feedback controller. The nominal controller is designed on the basis of partial feed back linearization theory. An adaptive fuzzy interference mechanism has been used for optimal estimation of the upper bound of the IM lumped uncertainty function. Using this mechanism, the SMC chattering is effectively reduced and the drive system performance becomes robust and stable against the motor parameter uncertainties and disturbance load torque. A simple method has also been proposed for on-line estimation of the stator resistance that makes the MRAS-based speed estimator insensitive to the stator resistance variations. Finally, a software package has been developed which can be used for direct vector controlling of the speed sensorless induction machine drives in each of the rotor, stator and air gap flux field-oriented reference frames.

REFERENCES


