Adaptive Speed Controller for Vector Control Drives of Induction Machines Using Fuzzy-Sliding Mode Method

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ABSTRACT

In this paper a new speed controller for vector control of induction machine is introduced. To do this a proportional-integral adaptive-sliding mode controller is designed and fuzzy rules are used for the adaptation method. Consequently dynamical response of the system is improved using fuzzy rules. In addition the implementation of such a method by digital processing instrument is simple. Stability of the system is utilized in achieving the adaptation rule therefore the system will be stable. A proportional-integral controller is used for the speed controller of usual vector control so calculation of its proportional and integral gain values is difficult to maintain the stability, however, the new controller does not have the stability problem. To evaluate the effectiveness of the new controller an experimental set-up was prepared and simulation and experimental results are presented at the last section of the paper.

KEYWORDS: Adaptive controller, Fuzzy-sliding mode control, Vector control.

INTRODUCTION

Electrical machines are highly non-linear systems especially when they are accompanied by control systems like the speed, position and torque control which carry mutual dynamics. Therefore, non-linear behaviors such as bifurcated, chaotic, sub-harmonic and, etc are mostly probable in these dynamical systems. On the other hand, from practical point of view, these time responses are not acceptable except the stable ones. It is obvious that these non-linear behaviors cause fluctuation in electromagnetic torque, speed of the rotor, current components, supply voltages, flux linkages and, etc. Some papers show the appearance of strange time responses in electrical machines drives such as induction machines, synchronous reluctance machines, permanent magnet synchronous machines, DC machines, brushless DC machines [1-7].

Nowadays, vector control drives of induction machine are more suitable in different applications such as trains and conveyer belts because of their suitable dynamical behavior plus having tough structure. Nevertheless, these drives may represent stable, unstable and chaotic response. In this regard, there are some researches that show bifurcated and chaotic response of these drives emanating from different reasons like inverter malfunction or even because of inherent non-linearity of the system [16-18].

A proportional-integral speed controller is used in usual vector control of induction machines [19]. Many papers show there is a range of proportional and integral gain values that causes unstable and chaotic response for each drive's characteristics. Finding this range requires cumbersome and difficult mathematical procedures [17]. Adaptive controller is a method having self-tuning ability of the proportional and integral controller gain. Sliding mode controller is used in some researches and fuzzy rules are applied to reduce the chattering phenomena caused by sliding mode control [20-23]. Dynamical response can be improved if adaptation method is used for controller. Therefore, in this paper a new adaptation method is utilized to maintain the stability and also improve the dynamical response of the system. In addition, along with the implementation of the sliding mode controller as adaptation rule, fuzzy rules are applied to enhance the quality of the dynamical response. In the last section of this paper experimental and simulated results are presented for usual vector control, sliding mode vector control and the proposed method. It is shown that the dynamical response of the new method is better than the others.

2. Adaptive Controller Using Sliding Mode and Fuzzy Rules

2.1 Sliding Mode Controller

Consider a non-linear system like:

\[ \dot{x} = F(x) + G(x)u(t) \]  

The aim of the controller design is to calculate the \( u(t) \) in such a way that the origin is a stable equilibrium point of the system. In sliding mode controller design, \( u(t) \) is calculated so that the system trajectory is placed on a stable dynamic like \( S(x) \). The following two conditions must be satisfied in the calculation of \( u(t) \).

a- The system trajectory reaches to sliding surface \( (S(x)) \).

b- The system trajectory remains there forever.

to do this, \( u(t) \) is regarded as \( u(t) = u_{eq}(t) + u_{nf}(t) \). In this equation \( u_{nf}(t) \) and \( u_{eq}(t) \) are regarded to satisfy the conditions of (a) and (b), respectively. Suppose a Lyapunov function as:

\[ V(x) = \frac{1}{2} s^{2}(x) \]
Therefore, its time derivative will be:

\[
\frac{d}{dt} V(x) = S(x) \cdot \frac{d}{dt} S(x) = S(x) \left( \frac{d}{dx} S(x) \right)^T \frac{dx}{dt}
\]  

(3)

According to the Lyapunov theory the system trajectory will reach to \( S(x) \) if Eq. (3) is negative everywhere in the space because the \( V(x) \) is positive and its time derivative is negative. \( u_{hf}(t) \) is considered for this goal in design of the controller. Also the system trajectory will remain on the \( S(x) \) when:

\[
\frac{d}{dt} S(x) = 0 = S(x) = 0
\]  

(4)

\( u_{eq}(t) \) is calculated to satisfy these conditions. \( u_{hf}(t) \) does not have any effect on the response of the system near the origin because it has high frequency therefore:

\[
\frac{d}{dt} S(x) = \left( \frac{d}{dx} S(x) \right)^T \frac{dx}{dt} = \left( \frac{d}{dx} S(x) \right)^T \left( F(x) + G(x) u_{eq} \right) = 0 \Rightarrow
\]  

(5)

As said above, \( u_{hf}(t) \) must be chosen such a way that the system trajectory reaches to \( S(x) \). One option for \( u_{hf}(t) \) is the \( \text{Sign}(x) \) function. By considering \( u_{hf}(t) = u_{hf}(t) = -K \times G^T(x)S_{\text{Sign}}(S(x)) \), it can be result as:

\[
\frac{d}{dt} V(x) = S(x) \cdot \frac{d}{dt} S(x) = S(x) \left( \frac{d}{dx} S(x) \right)^T \left( F(x) + G(x) \left( u_{eq} + u_{hf} \right) \right) =
\]  

(6)

\[
S(x) \left( \frac{d}{dx} S(x) \right)^T \left( F(x) + G(x) \left( u_{eq} - K \times G^T(x)S_{\text{Sign}}(S(x)) \right) \right) =
\]

\[
S(x) \left( \frac{d}{dx} S(x) \right)^T \left( F(x) + G(x) u_{eq} - K \times \left( \frac{d}{dx} S(x) \right)^T \|G(x)\|^2 \text{Sign}(S(x))S(x) \right) =
\]

\[-K \times \left( \frac{d}{dx} S(x) \right)^T \|G(x)\|^2 \|S(x)\| \]

\( S(x) \) is chosen so that \( \left( \frac{d}{dx} S(x) \right)^T \) will be a vector with positive constants. Eq. (6) shows that the time derivative of the Lyapunov function is negative therefore the condition (a) will be satisfied.

### 2.2 Adaptive Sliding Mode Controller and Fuzzy Rules

Without losing the generality, suppose that the goal of the control is adjusting two variables of the system such as \( \dot{x}_1 \) and \( \dot{x}_2 \) to their reference values such as \( \dot{x}_{1\text{ref}} \) and \( \dot{x}_{2\text{ref}} \). Furthermore, suppose that the sliding mode controller is used to control the \( \dot{x}_1 \) thus the principles were that mentioned in the previous section can be used to control it. Proportional-integral controller is used to control the \( \dot{x}_2 \) and according to the adaptive control theory both prorportional gain \( (k_p) \) and integral gain \( (k_i) \) can vary therefore in Eq. (1), \( u(t) \) can be written as:

\[
u(t) = k_p (\dot{x}_{2\text{ref}} - \dot{x}_2) - k_i \int (\dot{x}_{2\text{ref}} - \dot{x}_2) dt
\]  

(7)

The new system can be written as:

\[
\dot{x} = F(x) + G(x) \left( k_p (\dot{x}_{2\text{ref}} - \dot{x}_2) - k_i \omega_1 \right) \]

\[
\omega_1 = \dot{x}_{2\text{ref}} - \dot{x}_2
\]  

(8)

Sliding mode concept can be used to calculate the adaptation rules for adjusting these gains. A sliding surface like \( S_2 \) is used to achieve the adaptation rule of \( k_i(t) \) (\( S_1 \) is used to control the \( \dot{x}_1 \)), i.e.:

\[
S_2 = x_m - x_m
\]  

(9)

\( x_m \) is a variable that \( k_i \) is in its dynamic. According to sliding mode method \( k_i(t) \) can be divided as:

\[
k_{i(t)} = k_{ineq} + k_{ithf}(t)
\]  

(10)
\( k_{i_{\text{seq}}} \) can be calculated to satisfy the Eq. (5). \( k_{i_{\text{old}}}(t) \) can be regarded as \(-K_2 \times G^T(x) \text{Sign}(S_2(x))\), and \( K_2 \) can be computed using Eq. (6). Finally, the adaptation rule of integral gain is:

\[
k_{i_{\text{old}}}(t) = k_{i_{\text{seq}}} - K_2 \times G^T(x) \text{Sign}(S_2(x))
\]

(11)

\( K_2 \) must be a great integer to ensure the reachability to \( S_2 \), however, the bigger the value of integral gain the slower response of the system. To solve this problem this value should be decreased. Reaching to \( S_2 \) must be supervised during decreasing. It means that \( S_2 \) value and its time derivative should be checked. Fuzzy rules can be used to implement such an adjusting. Fig. (1) shows the rules table. In this table \( x_1 \) and \( x_2 \) are replaced with \( S_2 \) and \( \frac{d}{dt}S_2 \). This table is built using "if" and "then" rules. For example

if \( S_2 \) is negative medium (NM) and \( \frac{d}{dt}S_2 \) is positive small (PS)

then \( K_2 \) is positive medium (PM)

This rule is shown in this table.

Figure 1: Rule table.

3. Fuzzy-Sliding Mode Control of Rotor Field Oriented Vector Control

Eq. (12) shows dynamical equations of induction machine in rotor field oriented frame [19].

\[
\begin{align*}
\frac{d}{dt} \begin{bmatrix} i_{sx} \\ i_{sy} \\ \psi_r \\ \omega_r \end{bmatrix} &= \begin{bmatrix} (I_s \iota_s + I_m \iota_r) i_{sx} + (\omega_m I_s i_{sy}) i_{sx} + r_r \psi_r \\ -(I_s \iota_s + I_m \iota_r) i_{sy} - (\omega_m I_s i_{sy}) i_{sy} + (I_r \omega_m - L_r \omega_r) \psi_r \\ I_q L_r - I_m^2 \\ -r_r \left( \frac{1}{I_q} \psi_r - \frac{I_m}{I_q} i_{sx} \right) \end{bmatrix} + \frac{1}{L_q L_r - I_m^2} \begin{bmatrix} u_{sx} \\ u_{sy} \end{bmatrix} \\
&= \begin{bmatrix} I_q L_r - I_m^2 \\ -r_r \left( \frac{1}{I_q} \psi_r - \frac{I_m}{I_q} i_{sx} \right) \end{bmatrix} \left( \frac{3 P I_m}{2} \right) \psi_r i_{sy} - T_e - B_0 \omega_r
\end{align*}
\]

In this equation \( i_{sx}, i_{sy}, \psi_r \) and \( \omega_r \) are direct and quadrature components of stator current, rotor flux and speed, respectively. \( u_{sx} \) and \( u_{sy} \) are direct and quadrature components of stator voltage. Also, \( L_q' \) is stator transient inductance and other parameters are explained in table 1. Suppose that speed adjustment is the goal of the control system and speed controller is a proportional-integral adaptive controller [19], therefore:

\[
\begin{align*}
u_{sx} &= r_s k_{p_{sx}} \psi_r \text{ref} - \omega_m I_s i_{sx} \\
u_{sy} &= r_s k_{i_{sx}} (\omega_r \text{ref} - \omega_r) + r_s k_{i_{sx}} \omega_r + \omega_m I_s i_{sx}
\end{align*}
\]
\[
\begin{align*}
\frac{d}{dt} \begin{pmatrix} i_{sx} \\ i_{sy} \
\end{pmatrix} &= \frac{1}{\omega_r} \left( r_i l_{sxref} - \omega_m l_i l_{sy} - (L_r r_x + L_m r_y)i_{sx} + \omega_m l_i l_{sx} + r_r \psi_r \right) \\
&\quad + r_r k_p (\omega_{ref} - \omega_r) + r_r k_i \omega_r + \omega_m i_{sx} - (L_r r_x + L_m r_y)i_{sy} - \omega_m l_i l_{sx} + (L_r \omega_m - L_r \omega_r) \psi_r \\
&\quad + \frac{P}{2} \left( 3 P I_m \frac{l_r}{l_r} \psi_r l_{sx} - T_L - B \omega_r \right) \\
&\quad + \omega_{ref} - \omega_r 
\end{align*}
\]

(14)

Fig. (3) shows the speed control diagram. \( i_{sx} \), \( i_{sy} \) and \( \omega_r \) are considered as \( \tilde{x}_1 \), \( \tilde{x}_2 \) and \( x_m \) according to the section 2.2.

Two sliding surfaces are regarded to achieve the rules of the control:

\[
S_1 = i_{sx}^* - i_{sx}
\]

(15)

\[
S_2 = i_{sy}^* - i_{sy}
\]

(16)

In these equations \( i_{sx}^* \) and \( i_{sy}^* \) are the required direct and quadrature current components to achieve the necessary rotor flux and speed. \( i_{sx}^* \) can be calculated using reference rotor flux:

\[
i_{sx}^* = \frac{\psi_{ref}}{l_m}
\]

(17)

and \( i_{sy}^* \) can be calculated from mechanical equation:

\[
J \frac{d \omega_r}{dt} = \frac{P}{2} \left( T_e - T_L - B \omega_r \right)
\]

(18)

Replacing the electrical torque with rotor flux and stator current components results:

\[
J \frac{d \omega_r}{dt} = \frac{P}{2} \left( 3 P I_m \frac{l_r}{l_r} \psi_r l_{sx} - T_L - B \omega_r \right)
\]

(19)

In steady state, \( i_{sy}^* \) can be achieved as:
\( 0 = \frac{P}{2} \left( \frac{3P_m}{2I_r} \psi \omega - T_i - B_0 \omega \right) \)
\[
i_{sy} = \frac{2I_r}{3P_m} \frac{T_i + B_0 \omega}{\psi} \tag{20}\]

\( i_{sxref} \) and \( k_{io} \) are regarded as:
\[
i_{sxref} = i_{sxrefeq} + i_{sxrefh} \tag{21}\]
\( k_{io} = k_{ioeq} + k_{ioh} \)

\( k_{psw} \) is set to constant value. Two Lyapunov functions are regarded as:
\[
V(S_1) = \frac{1}{2} S_1^2 \tag{22}\]
\[
V(S_2) = \frac{1}{2} S_2^2 \tag{23}\]

\( i_{sxrefh} \) can be achieved using the first Lyapunov function.
\[
S_1 = - \frac{d}{dt} i_{sx} = \frac{L_s}{L_s - L_m^2} \left( \omega_m L_r i_{sx} + (I_r L_s + L_m r_p) i_{sx} - \omega_m L_r i_{sy} - r_p \psi \right) \tag{24}\]

On the other hand, the system dynamic on the \( S_1 \) and \( S_2 \) surfaces are:
\[
S_1 = 0 \Rightarrow i_{sx} = i_{sx}^* \quad \psi_t = \psi_{ref} \tag{25}\]
\[
S_2 = 0 \Rightarrow i_{sy} = i_{sy}^* \tag{26}\]

Replacing these values in Eq. (23) results:
\[
i_{sxrefeq} = \frac{\omega_m L_r i_{sy}^* + (I_r L_s + L_m r_p) i_{sy}^* - \omega_m L_r i_{sy}^* - r_p \psi_{ref}}{L_s L_r - L_m^2} \tag{27}\]

\( k_{io} \) can be calculated by the following operation procedure on \( S_2 \).
\[
k_{ioeq} = \frac{\frac{d}{dt} S_2 = 0 \Rightarrow}{L_s L_r - L_m^2} \tag{28}\]

\( i_{sxrefh} \) can be assumed as \( K_3 \times \text{Sign}(S_1) \), therefore:
\[
S_1 \frac{d}{dt} S_1 = \frac{-r_s K_3 |S_1| + (L_s - L_m L_p) \left( \omega_m i_{sy} - \omega_m i_{sx}^* \right) S_1 - (I_r L_s + L_m r_p) S_2^2 - r_p (\psi_t - \psi_{ref}) S_1}{L_s L_r - L_m^2} \tag{29}\]

As it can be seen in Eq. (27), if \( K_3 \gg 1 \) this equation is negative and hence the system trajectory reaches to \( S_1 \). To calculate the \( k_{ioh} \) the same procedure must be done. \( k_{ioh} \) is regarded as \( K_2 \times \text{Sign}(S_2) \) therefore:
\[
f_{r} k_{psw}(\omega_{ref} - \omega_t) + \left( (L_s L_p - L_m Z_0) + (L_r L_s + L_m r_p) Z_2 - (L_s Z_0 - L_m Z_4) \right) \]
\[
Z_1 = \omega_m r_s i_{sx} \omega_t - \omega_m i_{sx}^* \]
\[
Z_2 = i_{sy} \omega_t - i_{sy}^* \]
\[
Z_3 = \omega_m \psi_{ref} \omega_t - \omega_m \psi_{ref}^* \]
\[
Z_4 = \omega_m \psi_{ref} \omega_t - \omega_m \psi_{ref}^* \]
\[
\omega_{isl} = \omega_{i} \tag{29}\]

As seen again choosing \( K_3 \gg 1 \) results negative time derivative and the system trajectory will reach \( S_2 \) too.
As mentioned above both $K_1$ and $K_2$ values must be great integer to ensure negative time derivative of the Lyapunov functions. This issue can cause the system response to be too slow because bigger integral controller gain results slower response of the system. To solve this problem both $K_1$ and $K_2$ values must be reduced and also the sliding mode condition should be satisfied. In other word, this values can be decreased if the time derivatives of Lyapunov function is negative. To do this Fuzzy rules are used. Figs. (4-a) and (4-b) shows how $K_1$ and $K_2$ values are adjusted by fuzzy rules. In both cases the time derivative of current components are required. When digital sampling is done, this procedure is easy to perform. Eq. (29) shows the time derivatives of the current components.

\[
\frac{di}{dt} \cong \frac{\Delta i}{\Delta t} = \frac{i(t + \Delta t) - i(t)}{\Delta t}
\]

In this equation $T_s$ is the sampling interval.

4. Numerical Analysis and Experimental Results

An experimental prototype was prepared by the authors to verify the analysis. A laboratory 350 W, 4-pole, 3-ph induction motor whose parameters and characteristics are shown in table 1 was used.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$r_s$</td>
<td>stator resistor</td>
<td>34.0 Ω</td>
</tr>
<tr>
<td>$r_r$</td>
<td>rotor resistor</td>
<td>16.2 Ω</td>
</tr>
<tr>
<td>$L_s$</td>
<td>stator inductance</td>
<td>0.847 H</td>
</tr>
<tr>
<td>$L_r$</td>
<td>rotor inductance</td>
<td>0.847 H</td>
</tr>
<tr>
<td>$L_m$</td>
<td>magnetizing inductance</td>
<td>0.774 H</td>
</tr>
<tr>
<td>$J$</td>
<td>moment of inertia</td>
<td>0.0015 kgm²/s</td>
</tr>
<tr>
<td>$B$</td>
<td>friction coefficient</td>
<td>0.001 Nkgm/s</td>
</tr>
<tr>
<td>$T_n$</td>
<td>no-load torque</td>
<td>0.1 Nm</td>
</tr>
</tbody>
</table>

Two AT91SAM7S64 microcontrollers are used to implement rotor speed calculation and filtering, speed controller, and processing operations including reference voltage values calculation for inverter, pulse-width-modulation (PWM), DC bus control, reference values setting such as reference speed and reference flux linkage and, etc. This microcontroller can work with 55 MHz clock signal, but they were used with 48 MHz clock and it has ARM core therefore it can be used for this drive which needs high speed processing operations. A 2000 pulse/revolution speed encoder is connected to the shaft of the rotor to measure the speed of the rotor. Its pulses after a voltage level adjusting are used for counter of the microcontroller. Microcontroller calculates the speed and after filtering by an FIR\(^1\) filtering and a low-pass filter which neutralizes the quantization error and noise provides the actual speed. A logic operation board is designed to operate on PWM pulses produced by microcontroller. These pulses after high speed optocouplers are delivered to the power section. An Inverter with IGBT\(^2\) switches, RC\(^3\) snubber circuit, high speed anti-parallel power diodes was designed to supply the motor. PWM pulses are delivered to the IGBT driver circuits to turn the IGBT

\(^1\) Finite Impulse Response
\(^2\) Isolated Gate Bipolar Transistor
\(^3\) Resistance-Capacitance
switches on or off. Moreover, a DC bus with 530 volts voltage is designed using three phase voltage rectifier, NTC\textsuperscript{4} resistors, relay, capacitors, discharge circuit and, etc. Two Hall effect sensors (ACS712) are used to measure the currents. These measurements after filtering are used to calculate the direct and quadrature current components. In addition, they are applied to achieve the rotor flux position. Fig. (5), shows the block diagram of experimental setup.

![Figure 5: Experimental Block Diagram.](image)

Figs. (6-a), and (6-b) show the simulated and experimental speed responses of the system for usual vector control, respectively. The dashed line shows the reference speed in Fig. (6-a). As seen in Fig. (6-b), the speed of the motor follows the reference speed, but its dynamical quality is not good. Figs. (7-a) and (7-b) show the simulated and experimental speed responses of the system for sliding mode control of vector control, respectively. The adaptation rules for this controller are shown in Eqs. (21), (25), and (26). Although its dynamical response is better than the previous controller, it is not good enough.

![Figure 6: Speed response of vector control.](image)

\textsuperscript{4} Temperature Coefficient
Figs. (8-a) and (8-b) show the simulated and experimental speed responses of the system for the proposed adaptive-fuzzy sliding mode vector control, respectively. The block diagram of this controller is shown in Fig. (3).

In this case, the $K_1$ and $K_2$ are adjusted according to Figs. (4-a) and (4-b). In spite of better dynamical response, implementation of this controller is simple when digital implementation is used. The reason is that in digital processing the time derivative of current components required for Eq. (29) is achieved just by difference in two consecutive sampling values. It must be noted that the noise can destroy the time derivative therefore using filters and several consecutive samples instead of one sample are necessary. As seen its dynamical response is very good. Fig. (9) shows the rotor flux vector phase angle for two different rotor speeds which shows the phase angle of the rotor flux is followed by the control system.
5. Conclusion
The new method causes stable response since the rules of controller gain tuning is sliding mode controller. Furthermore, its dynamical response is better than the usual sliding mode controller of vector control drives because the fuzzy rules are considered in tuning of controller gains. Stability is checked using the fuzzy rules therefore its stability will be maintained. In addition, implementation of the proposed method by digital processing devices is simple so it is another merit of the new controller.

REFERENCES