J. Basic. Appl. Sci. Res., 3(2)928-936, 2013
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ISSN 2090-4304
Journal of Basic and Applied Scientific Research
www.textroad.com

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 inegral 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 \cite{Ref8,Ref9}, error in parameter's values like rotor and stator resistances [10-15] or even because of inherent nonlinearity of the system [16-18].

A proportional-integral speed controller is used in usual vector control of induction machines [19]. Many researches 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:

$$\dot{\mathbf{x}} = \mathbf{F}(\mathbf{x}) + \mathbf{G}(\mathbf{x})u(t) \tag{1}$$

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 trajector is placed on a stable dynamic like $S(\mathbf{x})$. The following two conditions must be satisfied in the calculation of $u(t)$.

a- The system trajectory reaches to sliding surface ($S(\mathbf{x})$).

b- The system trajectory remains there forever.

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

$$V(\mathbf{x}) = \frac{1}{2}S^2(\mathbf{x}) \tag{2}$$

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Therefore, its time derivative will be:

$$\frac{d}{dt}V(\mathbf{x}) = S(\mathbf{x})\frac{d}{dt}S(\mathbf{x}) = S(\mathbf{x})\left(\frac{d}{dx}S(\mathbf{x})\right)^{T}\frac{d\mathbf{x}}{dt} \tag{3}$$

According to the Lyapunov theory the system trajectory will reach to $S(\mathbf{x})$ if Eq. (3) is negative everywhere in the space because the $V(\mathbf{x})$ 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 $S(\mathbf{x})$ when:

$$\begin{aligned} S(\mathbf{x}) &= 0 \\ \frac{d}{dt}S(\mathbf{x}) &= 0 \end{aligned} \tag{4}$$

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

$$\begin{aligned} \frac{d}{dt}S(\mathbf{x}) &= \left(\frac{d}{dx}S(\mathbf{x})\right)^{T}\frac{d\mathbf{x}}{dt} = \left(\frac{d}{dx}S(\mathbf{x})\right)^{T}\left(\mathbf{F}(\mathbf{x}) + \mathbf{G}(\mathbf{x})u_{eq}\right) = 0 \Rightarrow \\ u_{eq} &= \left(\left(\frac{d}{dx}S(\mathbf{x})\right)^{T}\mathbf{G}(\mathbf{x})\right)^{-1}\left(\left(\frac{d}{dx}S(\mathbf{x})\right)^{T}\mathbf{F}(\mathbf{x})\right) \end{aligned} \tag{5}$$

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

$$\begin{aligned} \frac{d}{dt}V(\mathbf{x}) &= S(\mathbf{x})\frac{d}{dt}S(\mathbf{x}) = S(\mathbf{x})\left(\frac{d}{dx}S(\mathbf{x})\right)^{T}\left(\mathbf{F}(\mathbf{x}) + \mathbf{G}(\mathbf{x})\left(u_{eq} + u_{hf}\right)\right) = \\ & S(\mathbf{x})\left(\frac{d}{dx}S(\mathbf{x})\right)^{T}\left(\mathbf{F}(\mathbf{x}) + \mathbf{G}(\mathbf{x})\left(u_{eq} - K \times \mathbf{G}^{T}(\mathbf{x})\mathrm{Sign}(S(\mathbf{x}))\right)\right) = \\ & S(\mathbf{x})\left(\frac{d}{dx}S(\mathbf{x})\right)^{T}\left(\mathbf{F}(\mathbf{x}) + \mathbf{G}(\mathbf{x})u_{eq}\right) - K \times \left(\frac{d}{dx}S(\mathbf{x})\right)^{T}\|\mathbf{G}(\mathbf{x})\|^{2}\mathrm{Sign}(S(\mathbf{x}))S(\mathbf{x}) = \\ & -K \times \left(\frac{d}{dx}S(\mathbf{x})\right)^{T}\|\mathbf{G}(\mathbf{x})\|^{2}\|S(\mathbf{x})\| \end{aligned} \tag{6}$$

$S(\mathbf{x})$ is chosen so that $\left(\frac{d}{dx}S(\mathbf{x})\right)^{T}$ 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 $\hat{x}_1$ and $\hat{x}_2$ to their reference values such as $\hat{x}_{1ref}$ and $\hat{x}_{2ref}$. Furthermore, suppose that the sliding mode controller is used to control the $\hat{x}_1$ thus the principles were that mentioned in the previous section can be used to control it. Proportional-integral controller is used to control the $\hat{x}_2$ and according to the adaptive control theory both prorortional gain (k_p) and integral gain (k_i) can vary therefore in Eq. (1), $u(t)$ can be written as:

$$u(t) = k_p(\hat{x}_{2ref} - \hat{x}_2) - k_i\int(\hat{x}_{2ref} - \hat{x}_2)dt \tag{7}$$

The new system can be written as:

$$\begin{aligned} \dot{\mathbf{x}} &= \mathbf{F}(\mathbf{x}) + \mathbf{G}(\mathbf{x})\left(k_p(\hat{x}_{2ref} - \hat{x}_2) - k_i\omega_i\right) \\ \dot{\omega}_i &= \hat{x}_{2ref} - \hat{x}_2 \end{aligned} \tag{8}$$

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 $\hat{x}_1$). i.e.:

$$S_2 = x_m^* - x_m \tag{9}$$

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

$$k_{i\omega}(t) = k_{i\omega eq} + k_{i\omega hf}(t) \tag{10}$$

$k_{i\omega eq}$ can be calculated to satisfy the Eq. (5). $k_{i\omega hf}(t)$ can be regarded as $-K_2 \times \mathbf{G}^T(\mathbf{x})\mathrm{Sign}(S_2(\mathbf{x}))$, and K_2 can be computed using Eq. (6). Finally, the adaptation rule of integral gain is:

$$k_{i\omega}(t) = k_{i\omega eq} - K_2 \times \mathbf{G}^T(\mathbf{x})\mathrm{Sign}(S_2(\mathbf{x})) \tag{11}$$

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. Fig. (2) shows the implementation of fuzzy rules to adjust K_2. These procedures must be done for K_1 and S_1.

x_2 \ x_1	NB	NM	NS	Z	PS	PM	PB
NB	PB	PB	PB	PB	PB	PB	PB
NM	PB	PB	PM	PM	PM	PB	PB
NS	PB	PM	PS	PB	PS	Z	NS
Z	PB	PM	PS	Z	NS	NM	NB
PS	PS	PM	NS	NB	NS	NM	NB
PM	NB	NM	NM	NM	NM	NB	NB
PB	NB	NB	NB	NB	NB	NB	NB

Figure 1: Rule table.

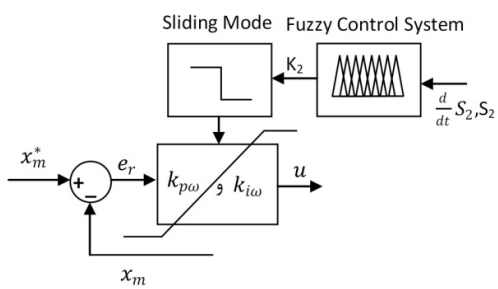


Figure 2: Fuzzy adjusting of K_2.

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].

$$\frac{d}{dt}\begin{pmatrix} i_{sx} \\ i_{sy} \\ \psi_r \\ \omega_r \end{pmatrix} = \begin{pmatrix} \dfrac{-(L_r r_s + L_m r_r)i_{sx} + (\omega_{mr} L_r L'_s)i_{sy} + r_r \psi_r}{L_s L_r - L_m^2} \\ \dfrac{-(L_r r_s + L_m r_r)i_{sy} - (\omega_{mr} L_r L'_s)i_{sx} + (L_{lr}\omega_{mr} - L_r \omega_r)\psi_r}{L_s L_r - L_m^2} \\ -r_r\left(\dfrac{1}{L_r}\psi_r - \dfrac{L_m}{L_r} i_{sx}\right) \\ \dfrac{P}{2J}\left(\dfrac{3P}{2}\dfrac{L_m}{L_r}\psi_r i_{sy} - T_L - B\omega_r\right) \end{pmatrix} + \frac{1}{L_s L_r - L_m^2}\begin{pmatrix} u_{sx} \\ u_{sy} \\ 0 \\ 0 \end{pmatrix} \tag{12}$$

In this equation i_{sx}, i_{sy}, ψ_r and ω_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'_s 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:

$$u_{sx} = r_s i_{sxref} - \omega_{mr} L'_s i_{sy}$$
$$u_{sy} = r_r k_{p\omega}(\omega_{ref} - \omega_r) + r_s k_{i\omega}\omega_i + \omega_{mr} L_s i_{sx} \tag{13}$$

$$\frac{d}{dt}\begin{pmatrix} i_{sx} \\ i_{sy} \\ \psi_r \\ \omega_r \\ \omega_i \end{pmatrix} = \begin{pmatrix} \dfrac{r_s i_{sxref} - \omega_{mr} L'_s i_{sy} - (L_r r_s + L_m r_r) i_{sx} + \omega_{mr} L_r L'_s i_{sy} + r_r \psi_r}{L_s L_r - L_m^2} \\ \dfrac{r_r k_{p\omega}(\omega_{ref} - \omega_r) + r_s k_{i\omega} \omega_i + \omega_{mr} L_s i_{sx} - (L_r r_s + L_m r_r) i_{sy} - \omega_{mr} L_r L'_s i_{sx} + (L_{lr}\omega_{mr} - L_r \omega_r)\psi_r}{L_s L_r - L_m^2} \\ -r_r\left(\dfrac{1}{L_r}\psi_r - \dfrac{L_m}{L_r} i_{sx}\right) \\ \dfrac{P}{2J}\left(\dfrac{3P}{2}\dfrac{L_m}{L_r}\psi_r i_{sy} - T_L - B\omega_r\right) \\ \omega_{ref} - \omega_r \end{pmatrix} \tag{14}$$

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

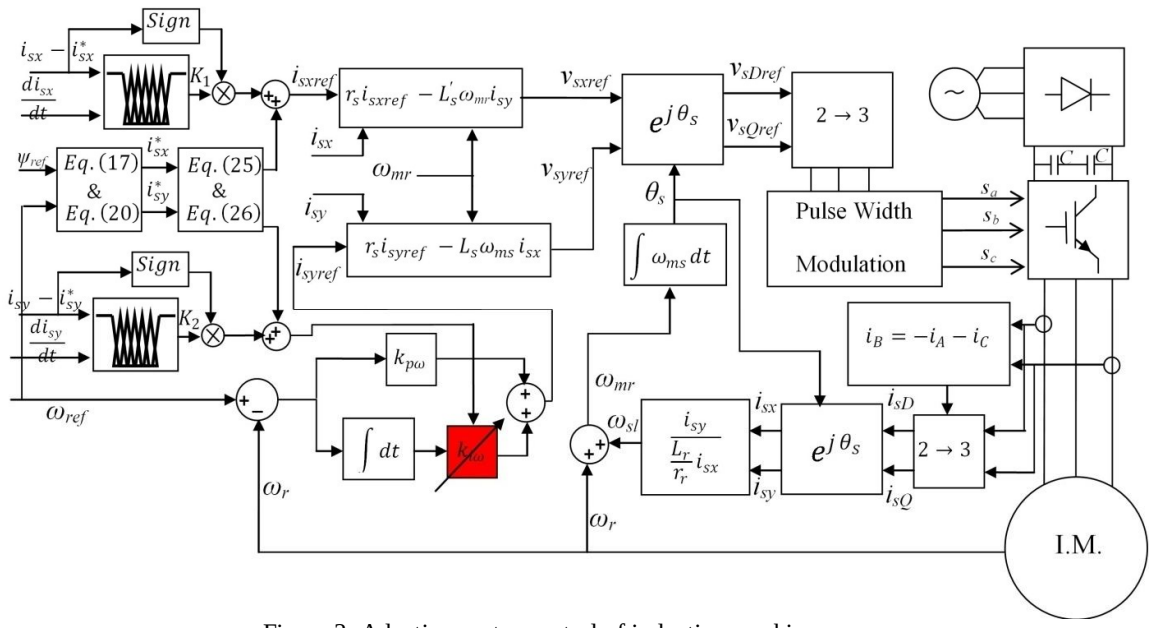


Figure 3: Adaptive vector control of induction machine.

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

$$S_1 = i^*_{sx} - i_{sx} \tag{15}$$

$$S_2 = i^*_{sy} - i_{sy} \tag{16}$$

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:

$$i^*_{sx} = \frac{\psi_{ref}}{L_m} \tag{17}$$

and i^*_{sy} can be calculated from mechanical equation:

$$J\frac{d\omega_r}{dt} = \frac{P}{2}(T_e - T_L - B\omega_r) \tag{18}$$

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

$$J\frac{d\omega_r}{dt} = \frac{P}{2}\left(\frac{3PL_m}{2L_r}\psi_{rx} i_{sy} - T_L - B\omega_r\right) \tag{19}$$

In steady state, i^*_{sy} can be achived as:

$$0 = \frac{P}{2}\left(\frac{3PL_m}{2L_r}\psi_{ref}i_{sy}^* - T_L - B\omega_{ref}\right)$$
$$i_{sy}^* = \frac{2L_r}{3PL_m}\frac{T_L + B\omega_{ref}}{\psi_{ref}} \tag{20}$$

i_{sxref} and $k_{i\omega}$ are regarded as:

$$i_{sxref} = i_{sxrefeq} + i_{sxrefhf}$$
$$k_{i\omega} = k_{i\omega eq} + k_{i\omega hf} \tag{21}$$

and $k_{p\omega}$ is set to constant value. Two Lyapunov functions are regarded as:

$$V(S_1) = \frac{1}{2}S_1^2$$
$$V(S_2) = \frac{1}{2}S_2^2 \tag{22}$$

$i_{sxrefhf}$ can be achieved using the first Lyapunov function.

$$\dot{S}_1 = -\frac{d}{dt}i_{sx} =$$
$$-\frac{r_s i_{sxrefeq} - \omega_{mr}L_s' i_{sy} - (L_r r_s + L_m r_r)i_{sx} + \omega_{mr}L_r L_s' i_{sy} + r_r\psi_r}{L_s L_r - L_m^2} = 0 \tag{23}$$

On the other hand, the system dynamic on the S_1 and S_2 surfaces are:

$$S_1 = 0 \Rightarrow i_{sx} = i_{sx}^*, \psi_r = \psi_{ref}$$
$$S_2 = 0 \Rightarrow i_{sy} = i_{sy}^* \tag{24}$$

Replacing these values in Eq. (23) results:

$$i_{sxrefeq} = \frac{\omega_{mr}^* L_s' i_{sy}^* + (L_r r_s + L_m r_r)i_{sx}^* - \omega_{mr}L_r L_s' i_{sy}^* - r_r\psi_{ref}}{L_s L_r - L_m^2} \tag{25}$$

$k_{i\omega}$ can be calculated by the following operation procedure on S_2.

$$\frac{d}{dt}S_2 = 0 \Rightarrow$$
$$k_{i\omega eq} = \frac{-\omega_{mr}L_s i_{sx} + (L_r r_s + L_m r_r)i_{sy} + \omega_{mr}L_r L_s' i_{sx} - (L_{lr}\omega_{mr} - L_r\omega_r)\psi_r}{r_s\omega_i^*} \tag{26}$$

$i_{sxrefhf}$ can be assumed as $K_1 \times \mathrm{Sign}(S_1)$, therefore:

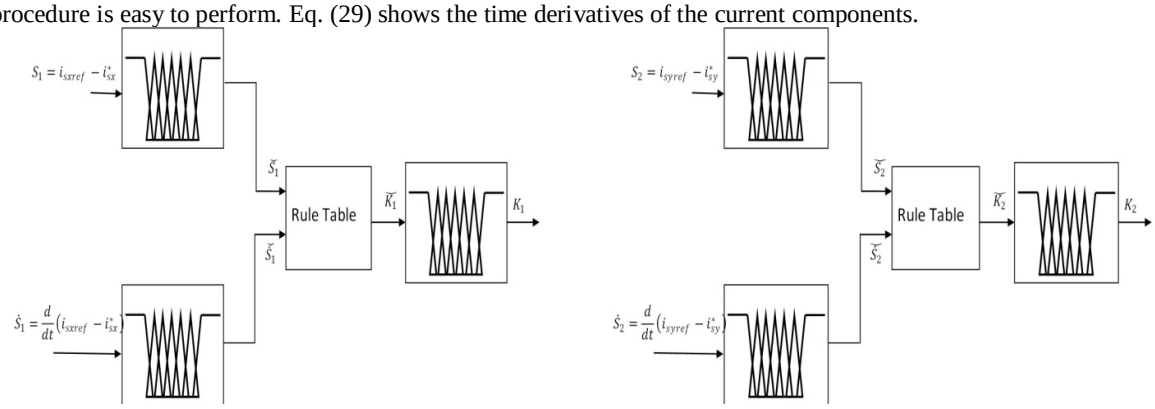
$$S_1\frac{d}{dt}S_1 = \frac{-r_s K_1|S_1| + (L_s' - L_r L_s')\left(\omega_{mr}i_{sy} - \omega_{mr}^* i_{sy}^*\right)S_1 - (L_r r_s + L_m r_r)S_1^2 - r_r(\psi_r - \psi_{ref})S_1}{L_s L_r - L_m^2} \tag{27}$$

As it can be seen in Eq. (27), if $K_1 \gg 1$ this equation is negative and hence the system trajectory reaches to S_1. To calculate the $k_{i\omega hf}$ the same procedure must be done. $k_{i\omega hf}$ is regarded as $K_2 \times \mathrm{Sign}(S_2)$ therefore:

$$S_2\frac{d}{dt}S_2 = \frac{r_s K_2}{L_s L_r - L_m^2}|S_2| +$$
$$\frac{r_r k_{p\omega}(\omega_{ref} - \omega_r) + ((L_r L_s' - L_s)Z_1 + (L_r r_s + L_m r_r)Z_2 - (L_{lr}Z_3 - L_r Z_4)}{L_s L_r - L_m^2}$$
$$Z_1 = \omega_{mr}^* i_{sx}^* \omega_{isl} - \omega_{mr} i_{sx}$$
$$Z_2 = i_{sy}^* \omega_{isl} - i_{sy}$$
$$Z_3 = \omega_{mr}^* \Psi_{ref}\omega_{isl} - \omega_{mr}\Psi_r$$
$$Z_4 = \omega_r^* \Psi_{ref}\omega_{isl} - \omega_r\Psi_r$$
$$\omega_{isl} = \frac{\omega_i}{\omega_i^*} \tag{28}$$

As seen again choosing $K_2 \gg 1$ results negative time derivative and the system trajectory will reache S_2 too.

As mentioned above both K_1 and K_2 values must be great integer to ensure negative time dirivative 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 dirivatives of Lyapunov function is negative. To do this Fuzzy rules are used. Figs. (4-a) and (4-b) shows how K_1 and K_2values are adjusted by fuzzy rules. In both cases the time dirivative 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.



(a) Adjusting the K_1. (b) Adjusting the K_2.

Figure 4: Using fuzzy rules to adjust the gains.

$$\frac{d}{dt} i_{sx} \cong \frac{\Delta i_{sx}}{\Delta t} = \frac{i_{sx}(t + T_s) - i_{sx}(t)}{T_s}$$
$$\frac{d}{dt} i_{sy} \cong \frac{\Delta i_{sy}}{\Delta t} = \frac{i_{sy}(t + T_s) - i_{sy}(t)}{T_s} \tag{29}$$

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 1: Induction motor parameters (380V, 50Hz)

Parameter	Description	Value
r_s	stator resistor	34.0 Ω
r_r	rotor resistor	16.2 Ω
L_s	stator inductance	0.847 H
L_r	rotor inductance	0.847 H
L_m	magnetizing inductance	0.774 H
J	moment of inertia	0.0015 Kgm
B	friction coefficient	0.001 Kgm/s
T_L	no-load torque	0.1 Nm

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[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.

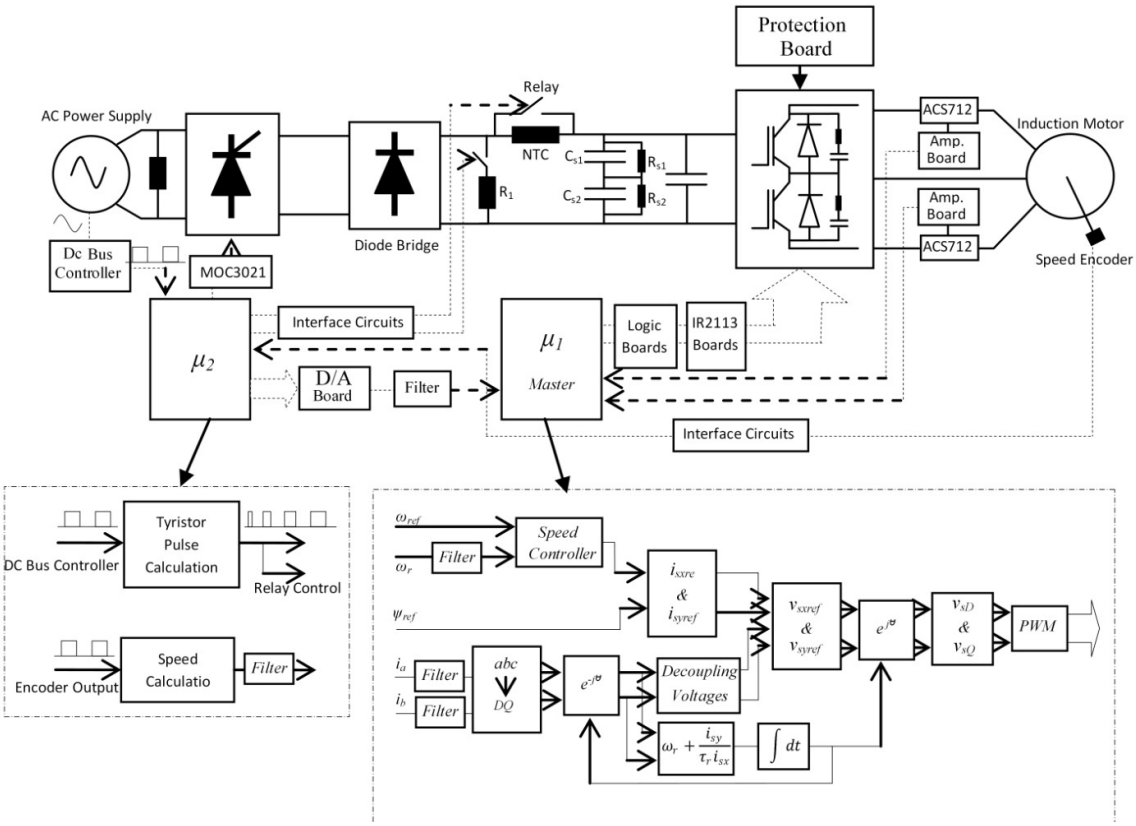
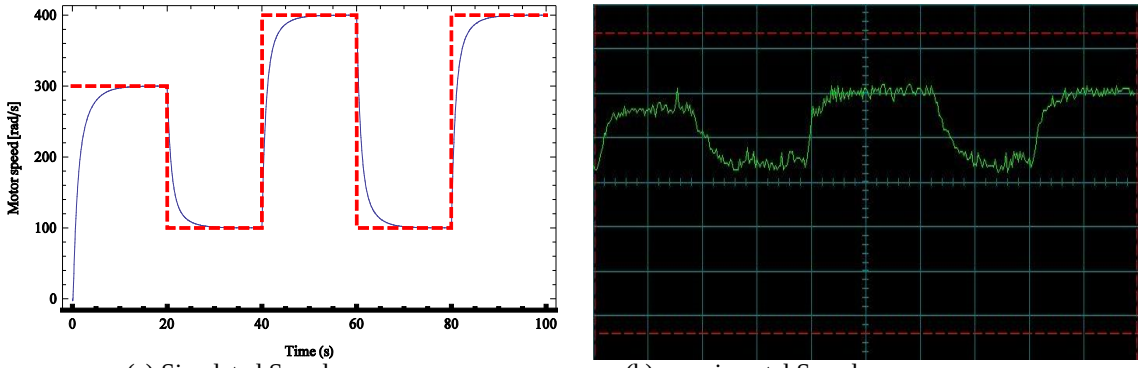


Figure 5: Experimental Block Diagram.

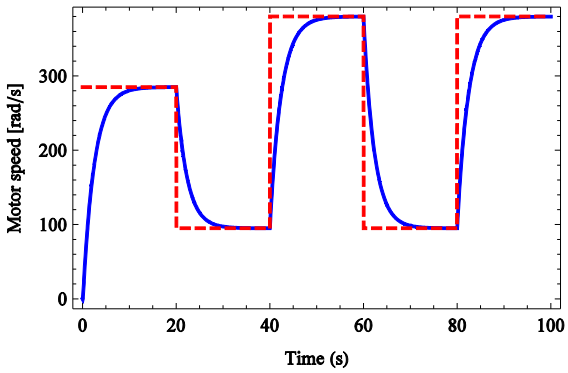
Figs. (6-a), and (6-b) show the simulated and exprimental speed responses of the system for usual vector control, respectively. The dashed line shows the reference speed in Fig. (6-a). As seen in Fig. (6-b), the speed of the motor follows the reference speed, but its dynamical quality is not good. Figs. (7-a) and (7-b) show the simulated and exprimental 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.



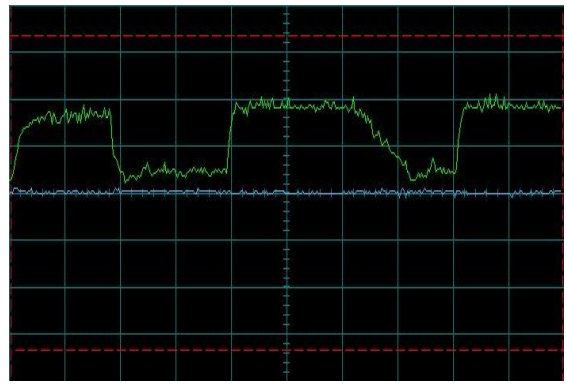
(a) Simulated Speed response. (b) experimental Speed response.

Figure 6: Speed response of vector control.

[4] Temperature Coefficient



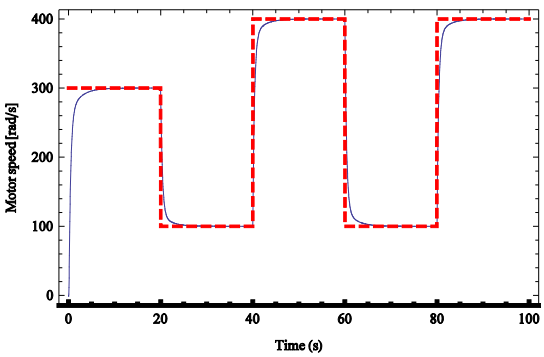
(a) Simulated Speed response.



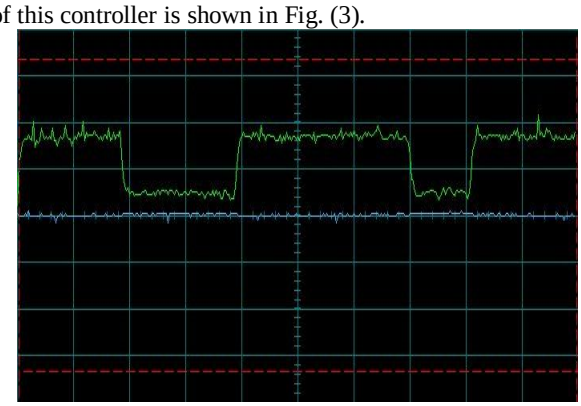
(b) experimental Speed response.

Figure 7: Speed response of sliding mode control of vector control.

Figs. (8-a) and (8-b) show the simulated and exprimental 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).



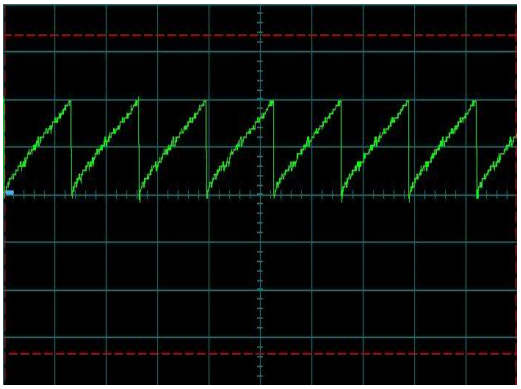
(a) Simulated Speed response.



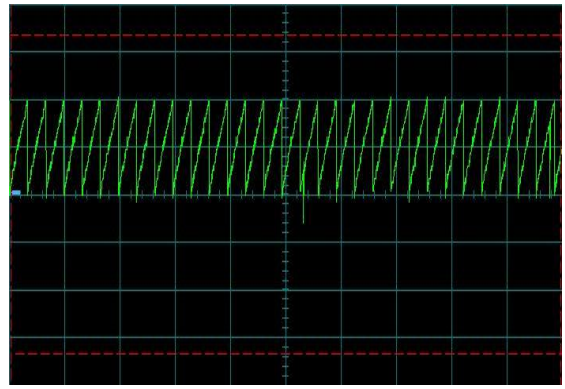
(b) experimental Speed response.

Figure 8: Speed response of adaptive-fuzzy sliding mode vector control.

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.



(a) rotor flux vector phase angle for ω_{ref} = 100 rad/s



(a) rotor flux vector phase angle for ω_{ref} = 400 rad/s

Figure 9: rotor flux vector angle.

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.

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