Design and Implementation of a Fuzzy Controller for a High Performance Induction Motor Drive

Chang-Ming Liaw and Jin-Blaio Wang

Abstract—A limit-cycle controlled induction motor drive with a fuzzy controller is designed and implemented. The torque and flux of the proposed drive system are regulated by the limit-cycle control technique. It follows that very quick torque response can be achieved. Since the dynamic model of this type of drive system is not easy to obtain, a fuzzy controller is developed and used in the speed control feedback loop to obtain good dynamic rotor speed response. The fuzzy algorithms in the proposed controller are systematically found from the intuition and experience about the motor drive systems. The experimental results indicate that good dynamic speed performance can be achieved by the proposed controller. Moreover, since the rotor parameters are not needed in the implementation of the drive system, and due to the inherent feature of high adaptive capability possessed by the fuzzy controller, the performance of the controlled drive system is rather insensitive to the parameter and operating condition changes.

I. INTRODUCTION

Today, the field-oriented control method [1]–[7] has made possible the application of induction motor drives in high performance industrial applications where only dc motor drives were previously available. In a perfect field-oriented induction motor, the coupling between the d and q axes is eliminated, hence, high performance of drive can be obtained. However, the decoupling characteristic of the field-oriented induction motor drive depends on the motor parameter changes [1], [2], [7]. Thus without applying more sophisticated control techniques, such as, adaptive control, variable structure system control, robust control, etc., very good performance with parameter insensitive property still cannot be achieved.

In this paper, a quick response induction motor drive using a fuzzy controller is presented. In the proposed drive system, a ROM table [8] is used to select the prescribed optimal switching pattern for the PWM inverter. Selection of switching pattern is based on limit-cycle control technique, such that under the preset constant flux condition, a very fast torque response can be achieved with minimized switching frequency. Although this drive system has the novel feature of possessing very fast torque response, good dynamic speed response is not easy to achieve. The chief reason is that accurate dynamic drive model is difficult to obtain, and hence, the controller design based on mathematical derivation is quite difficult to perform. To overcome this difficulty, a speed controller based on the fuzzy control algorithms [9]–[13] is developed and implemented, such that good dynamic responses both in the speed following the regulation characteristics can be achieved. The fuzzy algorithms adopted in the proposed controller are systematically found according to intuition and experience about the motor drive systems. Having testing the effectiveness of the proposed fuzzy controller by computer simulation, the hardware of the drive system is implemented, and the software realization of the fuzzy control algorithms is carried out using a PC-386 personal computer. The experimental results indicate that good dynamic speed response can be achieved by the proposed controller. Moreover, since the rotor parameters are not needed in the implementation of the drive system, and due to the inherent feature of highly adaptive capability possessed by the fuzzy controller, the performance of the controlled drive system is quite robust and insensitive to the parameter and operating condition changes.
TABLE I
THE REDUNDANCY CHARACTERISTICS OF THE SWITCHING VOLTAGE VECTORS

<table>
<thead>
<tr>
<th>Group</th>
<th>dq voltage vector groups</th>
<th>Terminal voltage vectors</th>
<th>Switching modes</th>
</tr>
</thead>
<tbody>
<tr>
<td>ZG</td>
<td>( V_0 )</td>
<td>( (0, 0, 0) )</td>
<td>( {(0, 0), (0, 0), (0, 0); (0, 0), (0, 0), (1, 1); (0, 0), (1, 1), (0, 0); \ldots } )</td>
</tr>
<tr>
<td></td>
<td>( (E, E, E) )</td>
<td>( {(0, 0), (1, 0), (1, 0) } )</td>
<td></td>
</tr>
<tr>
<td></td>
<td>( (-E, -E, -E) )</td>
<td>( {(0, 1), (0, 1), (0, 1) } )</td>
<td></td>
</tr>
</tbody>
</table>

| IG    | \( V_1 \)               | \( (E, 0, 0) \)          | \( \{(1, 0), (0, 0), (1, 0); \ldots \} \) |
|       | \( (0, -E, -E) \)        | \( \{(1, 0), (1, 0), (1, 0); \ldots \} \) |
|       | \( V_2 \)               | \( (0, 0, -E) \)         | \( \{(0, 0), (0, 0), (1, 0); \ldots \} \) |
|       | \( (E, E, 0) \)          | \( \{(0, 1), (0, 1), (0, 0); \ldots \} \) |
|       | \( V_3 \)               | \( (0, E, 0) \)          | \( \{(0, 0), (0, 1), (0, 0); \ldots \} \) |
|       | \( (-E, -E, 0) \)        | \( \{(0, 1), (0, 1), (0, 1); \ldots \} \) |
| LG    | \( V_4 \)               | \( (-E, 0, 0) \)         | \( \{(0, 0), (0, 0), (1, 0); \ldots \} \) |
|       | \( (0, E, E) \)          | \( \{(0, 1), (1, 0), (1, 1); \ldots \} \) |
|       | \( V_5 \)               | \( (0, 0, 0) \)          | \( \{(0, 0), (0, 0), (1, 0); \ldots \} \) |
|       | \( (-E, -E, 0) \)        | \( \{(1, 0), (0, 0), (1, 0); \ldots \} \) |
|       | \( V_6 \)               | \( (0, -E, 0) \)         | \( \{(0, 0), (0, 1), (0, 0); \ldots \} \) |
|       | \( (E, 0, E) \)          | \( \{(1, 0), (0, 1), (0, 1); \ldots \} \) |

Complex form can be expressed by

\[
V_\theta = V_{d\theta} + jV_{q\theta} = \sqrt{\frac{2}{3}} \left[ \left( V_a - \frac{V_b - V_c}{2} \right) + j \frac{\sqrt{3}}{2} V_b - \frac{\sqrt{3}}{2} V_c \right].
\]

Although \( 2^6 = 64 \) different switching modes are produced by the operations of inverter switches, only \( 27 \) \( (V_a, V_b, V_c) \) terminal voltage combinations and \( 19 \) distinct \( d-q \) equivalent voltage vector classes can be generated, since some switching modes yield the same terminal voltage vectors, and some terminal voltage vectors result in the same \( d-q \) voltage vector. According to their amplitudes, these \( d-q \) voltage vector classes \( V_9 - V_{18} \) can be divided into three groups, namely the zero group (ZG, \( V_0 \)), the intermediate group (IG, \( V_1 - V_6 \)) and the large group (LG, \( V_7 - V_{18} \)). Each distinct \( d-q \) voltage vector class is composed of a different number of terminal voltage vectors and switching modes. Table I lists these redundancy characteristics for reference. The different terminal voltage vectors in each equivalence \( d-q \) voltage vector class can be selected to minimize the zero phase sequence current in the motor. Also, the different switching modes in each terminal voltage vector can be used to equalize the switching frequencies of two inverters. Fig. 2 shows the combined three \( d-q \) voltage vector groups and the 12 switching areas, where the \( i \)th area is identified by using a area index \( \theta_i \), which is defined as

\[
\theta_i: (i - 1)(\pi/6) < \alpha < i(\pi/6), \quad i = 1, 2, \cdots, 12
\]

where \( \alpha \triangleq \tan^{-1}(\phi_{d\theta}/\phi_{q\theta}) \), \( \phi_{d\theta} \), and \( \phi_{q\theta} \) denote the \( d \) and \( q \)
Fig. 1. The configuration of power circuit used in the proposed drive system.

Fig. 2. The equivalent $d-q$ voltage vector groups and switching areas.

components of flux linkage that is calculated by

$$\phi_\alpha = \phi_{0\alpha} + j\phi_{0\beta} = \int (V_s - R_s i_s) \, dt + \phi_{0\alpha} \tag{3}$$

where $R_s$ is the stator resistance, and $\phi_{0\alpha}$ is the initial value of $\phi_\alpha$ at the beginning of the switching interval. It is worth mentioning that some problems arise from the integration operation at very low velocity regions. The reason for this is that the operation of integrators cannot be done perfectly at near zero velocity because of the small induced electromotive force in the motor, so that control of $\phi_\alpha$ might be unstable when $R_s$ deviates from the correct value. These problems can be negligible at relatively high velocities such as above 2 Hz. However, another calculation of $\phi_\alpha$ must be employed below 2 Hz [8].

The objectives of switching mode selection are 1) giving quick torque response under constant flux level; 2) reducing zero phase sequence current; and 3) minimizing and equalizing the switching frequencies of inverters. The algorithms [8] for switching mode selection are described as follows.

A. Flux Control

The stator voltage vector $V_s$ in (3) supplied by the inverter is constant at each switching interval. If $R_s$ is neglected, the flux linkage at a particular switching interval can be expressed as

$$\phi_\alpha = V_s t + \phi_{0\alpha} \tag{4}$$

The stator flux linkage is regulated by the limit-cycle control technique such that its amplitude is forced to track the setting value $|\phi_\alpha|^*$ within a hysteresis band $|\Delta \phi_\alpha|$:

$$|\phi_\alpha|^* - |\Delta \phi_\alpha| < |\phi_\alpha| < (|\phi_\alpha|^* + |\Delta \phi_\alpha|) \tag{5}$$

The switching condition of $|\phi_\alpha|$ can be identified by a flux index $\phi_i$, $i = 0, 1; (\phi_0, \phi_1) = (0, 1)$. The status "0" and status "1" denote that the flux linkage needs to be increased and decreased, respectively. The flux index can be generated using a hysteresis comparator shown in Fig. 3(a).

B. Torque Control

In order to force the generated torque $T$ to track the command $T^*$ within a desired hysteresis band with minimum response time, a four-loop hysteresis five-level comparator shown in Fig. 3(b) is used for choosing the $d-q$ voltage vector. In general, at low speed or a steady-state operating condition, the voltage vectors in IG and ZG are selected, i.e., operating in loop 2. Now if the command $T^*$ is increased letting $|T^* - T| > \Delta T/2$, this will cause the control
to switch from loop 2 to loop 1, and furthermore, the switching stays in LG. When the generated torque increases to let \( |T^* - T| < \Delta T_1/2 \) or \( \Delta T_2 > \Delta T_1 \), the switching will change to loop 2. However, if \( |T^* - T| < (\Delta T_2/2) \), due to command change or disturbance, the voltage vector will be selected back from loop 2 to loop 1. The present torque can be calculated by

\[
T = \phi_{dq} q_{i_{\alpha}} - \phi_{dq} q_{i_{\beta}}.
\]

(6)

Based on the previous analyses, the torque index \( t_i (i = 0, 1, 2, 3, 4) \) used to identify the present torque condition can be obtained from Fig. 3(b) as

\[
t_0, t_1, t_2, t_3, t_4 = (-2, -1, 0, 1, 2) \quad \Delta
\]

(CCW LG, CCW IG, ZG, CW IG, CW LG)

(7)

where CW and CCW denote clockwise and counter clockwise respectively.

C. Zero Phase Sequence Current Reduction

At any instant, using the measured variables, the area index \( \theta_i \) is determined first. Then the torque index \( t_i \) and the flux index \( \phi_i \) are obtained from the comparators of Fig. 3(b) and (a). According to these three indexes, the \( dq \) voltage vectors, and hence, the switching modes at all conditions can be determined. However, since the inductor is open-delta connected, the neutral points of the motor are not isolated and zero phase sequence current \( i_0 \) exists. It is known that the change of \( i_0 \) is dependent on the magnitude and polarity of \( V_i \triangleq V_a + V_b + V_c \) developed by the inverter. By supposing that \( i_0 = i_x + i_y + i_z \) is calculated from the measured phase current, the hysteresis loop with three-level comparator shown in Fig. 3(c) can be used to yield the current index \( I_i \), which is defined as

\[
\begin{align*}
I_0 &= I_0 = 1, & \text{if } i_0 > \Delta i_0 \\
I_1 &= I_1 = 0, & \text{if } -\Delta i_0 < i_0 < \Delta i_0 \\
I_2 &= I_2 = -1, & \text{if } i_0 < -\Delta i_0.
\end{align*}
\]

(8)

According to the identified current index \( I_i \), and using the equivalence set concept of Table 1, the terminal voltage vector is changed such that \( i_0 \) is reduced, but the \( d-q \) equivalence voltage vector is not altered.

D. Switching Frequency Coordination of the Inverter

Having determined the terminal voltage vectors at every instant using the limit-cycle control algorithms for torque, flux, and zero phase sequence current, the switching modes can be determined. However, the switching frequencies of the two inverters may not be in balance. Table 1 shows when the terminal voltage \( V_x (x = a, b, c) \) equals to \( E \) or \(-E \), the switching mode \((S_{1x}, S_{2x})\) is uniquely determined as \((1, 0)\) and \((0, 1)\), respectively. However, for the case of \( V_x = 0 \), the switching modes \((S_{1x}, S_{2x}) = (0, 0)\) and \((1, 1)\) can be selected alternately to equalize the switching frequencies of the two inverters.

E. Implementation of the Switching Mode Generation Mechanism

The configuration of the whole drive system is shown in Fig. 4, in which block 1 is the power circuit; block 2 shows the torque and flux estimating circuits; the switching mode generating scheme is shown in blocks 3 and 4; and block 5 denotes the proposed fuzzy controller, which is realized digitally using a PC-386 personal computer. According to the above analyses, the switching mode pattern, without considering the frequency coordination, is programmed and stored in a ROM circuit. The 10-b address of the ROM is constructed by using the torque index (3 b), flux index (1 b), zero current index (2 b), and area index (4 b). The outputs from the ROM circuit are processed by the equalizing circuits to yield the actual switching signals for the inverter transistors.

III. THE PROPOSED FUZZY CONTROL ALGORITHMS FOR MOTOR DRIVES

Generally, the requirements for a high performance motor drive system are: 1) fast tracking of set point changes without overshoot; 2) the maximum speed dip and the restore time due to step load change must be kept small as much as possible; and 3) the steady-state errors both in the command tracking and load regulation cases must be zero. In order to achieve these requirements, the output feedback loops for controlled variables of the drive system must be added. Since the dynamic drive system model is not necessary for the fuzzy controller design, and the performance of the fuzzy controller is insensitive to the parameter changes, it is very suitable in this application. Theoretic bases of the fuzzy set theory have been introduced in many literatures [9]–[13]. However, unfortunately, there is no mature guidance for fuzzy control algorithm determination. In the following, the development of the fuzzy controller for the motor drive system is described in detail.

A. Dynamic Signal Analysis

For convenience of incorporating the intuition and experience into the fuzzy control algorithms, the behavior of the dynamic motor speed response is first investigated. The speed error and the speed error change of the drive system are defined as

\[
\begin{align*}
\epsilon(k) &\triangleq \omega^*(k) - \omega(k) \\
\Delta \epsilon(k) &\triangleq \epsilon(k) - \epsilon(k-1)
\end{align*}
\]

(9)

(10)

where \( \omega^*(k) \) is the speed command in kth sampling interval, \( \omega(k) \) is the speed response in kth sampling interval, \( \epsilon(k) \) is the speed error in kth sampling interval, and \( \Delta \epsilon(k) \) is the speed error change in kth sampling interval.
The general waveforms of the drive rotor speed response and the command are drawn in Fig. 5(a). According to the magnitude of $e$ and the sign of $\Delta e$, the response plane is roughly divided into four areas. The index used for identifying the response area is defined as

$$a_1: e > 0 \text{ and } \Delta e < 0, \quad a_2: e < 0 \text{ and } \Delta e < 0$$

$$a_3: e < 0 \text{ and } \Delta e > 0, \quad a_4: e > 0 \text{ and } \Delta e > 0.$$  

(11)

For further increases in the resolution of the behavior representation, the response around the set point and the extremes in Fig. 5(a) are emphasized in Fig. 5(b) and (c), respectively. The crossover index $c_i$ for identifying the slope of the response across the set point is defined as

$$c_1: (e > 0 \rightarrow e < 0) \text{ and } \Delta e \ll 0$$

$$c_2: (e > 0 \rightarrow e < 0) \text{ and } \Delta e \ll 0$$

$$c_3: (e > 0 \rightarrow e < 0) \text{ and } \Delta e \ll 0$$

$$c_4: (e < 0 \rightarrow e > 0) \text{ and } \Delta e > 0$$

$$c_5: (e < 0 \rightarrow e > 0) \text{ and } \Delta e \gg 0$$

$$c_6: (e < 0 \rightarrow e > 0) \text{ and } \Delta e \gg 0.$$  

(12)

Also, the magnitude index for representing the extent of overshoot and undershoot is defined as

$$m_1: \Delta e \approx 0 \text{ and } e \ll 0$$

$$m_2: \Delta e \approx 0 \text{ and } e \ll 0$$

$$m_3: \Delta e \approx 0 \text{ and } e < 0$$

$$m_4: \Delta e \approx 0 \text{ and } e > 0$$

$$m_5: \Delta e \approx 0 \text{ and } e \gg 0$$

$$m_6: \Delta e \approx 0 \text{ and } e \gg 0.$$  

(13)

**B. Linguistic Control Rules**

The three types of indexes previously mentioned can be combined and shown in the state plane of Table II for reference, where the qualitative statements are quantized by using the linguistic set defined as

$$\{\text{NB, NM, NS, ZE, PS, PM, PB}\}$$

(14)

where N is negative, P is positive, B is big, M is medium, S is small, and ZE is zero.

The linguistic control rules defined according to Fig. 5 and Table II are listed in Table III. The conditional rules are implied in the table, see for example, the element of the first row and seventh column that implies

$$\text{IF } e \text{ is PB and } \Delta e \text{ is NB THEN the control input is PB.}$$

(15)

**C. Membership Functions**

Having defined the fuzzy linguistic control rules, the membership functions corresponding to each element in the linguistic set must be defined. Depending on particular applications, many types of membership functions can be defined. For simplicity, the trapezoidal functions shown in Fig. 6 are proposed, where the universes of discourse of the error and error change are $-1.2$ V (the scaling factor is $1$ V to $360$ rpm) to $1.2$ V ($-6, -5, \cdots, 5, 6$) and $-0.3$ to $0.3$ V ($-6, -5, \cdots, 5, 6$), respectively. The quantizations of $e$ and

$\Delta e$ are shown in Table IV. The control input is also quantized into thirteen levels. It is obvious that the membership functions shown in Fig. 6 can also be mathematically expressed as:
\[
\begin{align*}
\text{ZE : } f &= \begin{cases} 
0, & 6 \geq x > 2 \\
-2/3(x - 0.3) + 1, & 2 \geq x > 0.5 \\
2/3(x + 0.5) + 1, & -1/2 \geq x > -2 \\
0, & -2 \geq x > -6 \\
2/3(x - 1.5) + 1, & 1.5 \geq x > 0
\end{cases} \\
\text{PS : } f &= \begin{cases} 
1, & 2.5 \geq x > 1.5 \\
-2/3(x + 2.5) + 1, & 4 \geq x > 2.5 \\
0, & 6 \geq x > 4 \\
0, & 2 \geq x > -6 \\
2/3(x - 3.5) + 1, & 3.5 \geq x > 2
\end{cases} \\
\text{PM : } f &= \begin{cases} 
1, & 4.5 \geq x > 3.5 \\
-2/3(x - 4.5) + 1, & 6 \geq x > 4.5 \\
0, & x > 6 \\
0, & 4 \geq x > -6 \\
2/3(x - 5.5) + 1, & 5.5 \geq x > 4 \\
1, & 6 \geq x > 5.5 \\
0, & 0 \leq x \leq 6 \\
-2/3(x + 1.5) + 1, & -1.5 \leq x \leq 0
\end{cases} \\
\text{NS : } f &= \begin{cases} 
1, & -2.5 \leq x \leq -1.5 \\
2/3(x + 2.5) + 1, & -4 \leq x \leq -2.5 \\
0, & -6 \leq x \leq -4 \\
0, & -2 \leq x \leq 6 \\
-2/3(x + 3.5) + 1, & -3.5 \leq x \leq -2 \\
1, & -4.5 \leq x \leq -3.5 \\
2/3(x + 4.5) + 1, & -6 \leq x \leq -4.5 \\
0, & x \leq -6 \\
0, & -4 \leq x \leq 6
\end{cases} \\
\text{NM : } f &= \begin{cases} 
-2/3(x + 5.5) + 1, & -5.5 \leq x \leq -4 \\
1, & -6 \leq x \leq -5.5
\end{cases} \\
\text{NB : } f &= \begin{cases} 
-2/3(x + 5.5) + 1, & -5.5 \leq x \leq -4 \\
1, & -6 \leq x \leq -5.5
\end{cases}
\end{align*}
\]

\[\text{(16)}\]

**IV. DESIGN AND IMPLEMENTATION OF THE FUZZY CONTROLLER**

Generally, the fuzzy controller that is based on the decision table of Table V, cannot lead to excellent performances both in the transient and static periods. Since the quantization level is too coarse, the overshoot and hunting around the set point in the static period may result. To solve this problem, the fine decision tables are usually used to replace the coarse table when the error falls within a preset limit. However, this increases the complexity of the system and thus the process time is increased. To solve this problem, a fuzzy controller for the motor drives is proposed in Fig. 7, in which an integral controller is used to replace the fine tables. When the error falls within a specified region (i.e., \(|e| < \epsilon\)), the fuzzy controller is disconnected (but the final control input contributed by the fuzzy controller is memorized by an integrator), and the integral controller remains for eliminating the error in steady state. Since the integral action must be dominated in the static period, and the overshoot in the tracking case must be avoided, a simple integral control having the ability of error adaptation is proposed by setting

\[K_I = K_{Is} - K_{Ip} \cdot |e|.\]  \[\text{(17)}\]

The parameters of \(K_{Is}\) and \(K_{Ip}\) are chosen as 15 and 20, respectively, in this case.

The parameters \(G_s\) and \(G_e\) in Fig. 7 are used to adjust the sensitivity of the quantization of error and error change. Although these parameters can be tuned according to the particular applications, they are all set to unity here for simplicity. On the other hand, contrary to that of the integral controller, the design of \(G_s\) is emphasized in the improvement of transient response speed. It follows that the output gain \(G_s\) is set as

\[G_s = K_{Is} + K_{Ip} \cdot |e|.\]  \[\text{(18)}\]

**TABLE IV**

**QUANTIZED ERROR AND ERROR CHANGE**

<table>
<thead>
<tr>
<th>error e (mV)</th>
<th>error change (\Delta e) (mV)</th>
<th>quantized level</th>
</tr>
</thead>
<tbody>
<tr>
<td>-1200</td>
<td>-300</td>
<td>-6</td>
</tr>
<tr>
<td>-1000</td>
<td>-250</td>
<td>-5</td>
</tr>
<tr>
<td>-800</td>
<td>-200</td>
<td>-4</td>
</tr>
<tr>
<td>-600</td>
<td>-150</td>
<td>-3</td>
</tr>
<tr>
<td>-400</td>
<td>-100</td>
<td>-2</td>
</tr>
<tr>
<td>-200</td>
<td>-50</td>
<td>-1</td>
</tr>
<tr>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>200</td>
<td>50</td>
<td>1</td>
</tr>
<tr>
<td>400</td>
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<td>250</td>
<td>5</td>
</tr>
<tr>
<td>1200</td>
<td>300</td>
<td>6</td>
</tr>
</tbody>
</table>
with $K_{op} = 5$, $K_{ol} = 0.5$. It is obvious from (18) that the output gain can be kept high in the transient period and then automatically reduced when it enters the static period. Accordingly, the output gain also has the error adaptive capability.

Before implementing the proposed controller, the computer simulation is first carried out. Since the dynamic model of this proposed drive system is rather difficult to find by the analytical derivation, the stochastic modeling technique developed in [14] is applied to estimate the continuous model of the proposed drive system. With the white noise superimposed on-line to a constant torque command equal to 3 V (corresponding to 37.5% rated torque), the rotor speed signal and the noise input signal are recorded, filtered, and sampled by a data acquisition system. The continuous model shown in Fig. 7 estimated from the sampled data is

\[ H_p(s) = \frac{1.1781}{s + 1.7167}. \]  

(19)

Without the proposed controller and from this estimated model, the dynamic speed responses of the drive system to a unit-step torque command and to a unit-step load torque disturbance show that (the scaling is 1-V torque signal to 0.5 N-m/rad, 1-V speed signal=360 r/min):

1) response time $t_r = 1.3$ s;
2) steady-state error of tracking is 0.314 V (113 r/min);
3) maximum dip $\Delta \omega_{dlm} = 0.686$ V (246 r/min); and
4) steady-state error of regulation $= -0.686$ V ($-247$ r/min).

In order to improve the response, the proposed controller is applied. Fig. 8(a) and 8(b) shows the unit-step speed tracking and unit-step load regulation responses. The effectiveness of the proposed controller can be observed by comparing the results shown in Fig. 8 and that of above. The dynamic rotor speed responses obtained by the conventional PI controller with $K_p = 5$, $K_i = 5$ are also shown in Fig. 8(a) and (b). The results indicate that the proposed controller gives better performances both in the tracking and the regulation characteristics. For testing the robustness of the proposed controller, Fig. 8(c) shows the rotor speed response when the model of (19) is suddenly changed to

\[ H_p(s) = \frac{1.8}{s + 1.3} \]  

(20)

at $t = 2$ s. The result shows that the response is insensitive to the parameter changes.

V. EXPERIMENTAL RESULTS

To further test the effectiveness of the proposed drive system, the hardware implementation of the drive system and the software realization of the proposed fuzzy controller using a PC-386 personal computer are performed. The motor used in this drive system is a 3-phase 60-Hz open-Δ-connected 4-pole 220-V 1-Hp induction motor. Fig. 9(a) shows the dynamic speed responses due to a step speed command ($\Delta \omega = 180$ r/min) applied when the motor was operated at ($\omega_{vo} = 360$ r/min, $R_L = 18.7$ $\Omega$) and
Fig. 8. The simulation results. (a) Unit-step tracking response. (b) Unit-step load regulation characteristic. (c) Unit-step tracking response with plant model changed from \( H_p(s) \) to \( H_p(s) \) at \( t = 2 \) s.

\( \omega_{\text{r}} = 720 \, \text{r/min}, R_{\text{c}} = 18.7 \, \Omega \), respectively. The results indicate that good speed following responses are obtained. As to the load regulation characteristics, a permanent magnet dc generator with switched resistors is used as the dynamic load of the proposed drive system. Fig. 9(b) shows the dynamic speed responses due to step load changes at two different operating conditions. Good speed regulation characteristics are also observed from the results in Fig. 9(b). Moreover, the control performances of the proposed controller are rather insensitive to the operating condition changes.

**VI. CONCLUSION**

The control system design and implementation of a high performance induction motor drive have been presented in this paper. Based on the limit-cycle control technique, very fast torque dynamic response under constant flux condition is achieved. To further obtain good dynamic speed responses both in the following and load disturbance regulation characteristics, a fuzzy controller and a systematic design procedure for forming the fuzzy algorithms are proposed. The algorithms are systematically constructed based on the experience about the motor drive system. Having tested the performance of the proposed controller by simulations, the implementations of the drive system and the proposed controller are performed. The experimental results show that good dynamic speed responses in both the command following and load regulation characteristics are achieved. Moreover, the performances are rather insensitive to the operating condition changes.

**APPENDIX**

**THE ROUTING FOR CONSTRUCTING THE LOOKUP DECISION TABLE**

begin
off line calculated decision table
for \( i_1 = 1 \ldots m_1 \) do
for \( i_2 = 1 \ldots m_2 \) do
begin
sum := 0;
weight sum := 0;
for each linguistic rules \( j = 1 \ldots n_3 \) do
begin
\( a[j] := \min \{ R_{11}(i1), R_{12}(i2) \}; \)
\( b[j] := w_j * a[j]; \)
sum := sum + a[j];
weightsum := weightsum + b[j];
end
end
applied the center of gravity method
table \([i1][i2] := \text{weightsum}/\text{sum};\)
end

where
\( i_1, i_2 \) indexes denote the quantization levels of error \( e \) and error change \( \Delta e \).
\( R_{11}, R_{12} \) membership functions,
\( w_j \) element in the universe discourse.

**REFERENCES**


