SLIDING MODE CONTROLLER DESIGN OF INDUCTION MOTOR BASED ON SPACE-VECTOR PULSEWIDTH MODULATION METHOD

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ABSTRACT. In this paper, an integrated sliding mode controller (SMC) based on space-vector pulsewidth modulation method is proposed to achieve high-performance speed control of an induction motor. Using a field-oriented control principle, a flux SMC is first established to achieve fast direct flux control and then a speed SMC is presented to enhance speed control by the direct torque method. Both SMCs are designed in light of predetermined load disturbance and parameter uncertainties. Finally, the effectiveness of the proposed control scheme under the load disturbance and parameter uncertainties is verified by simulation results and comparison with the direct torque control (DTC) and proportional-integral direct torque control (PIC) approaches.

Keywords: Induction motor, Sliding mode controller, Direct torque control

1. Introduction. The squirrel-cage induction motor has been widely used in many control systems and industrial automatic systems. It has many advantages, such as simple structure, firmness and low maintenance cost. In many applications of the induction motor, high performance speed control is the fundamental issue. However, induction motors are difficult to control for several reasons: their dynamics are intrinsically nonlinear and multivariable; not all of the state variables and not all of outputs to be controlled may be available for feedback; there are critical parameters (for instance, load torque, stator and rotor resistances) which may considerably vary during operations [1]. In recent years, various high-performance control strategies have been developed for the induction motor [1-9]. A well-known method in AC servo drive systems is field-oriented control (FOC) [1,2]. With field-oriented techniques, the decoupling of the induction motor’s flux and torque can be controlled separately as an excited DC motor. But FOC requires accurate and complex calculation of the decoupling, so it is difficult to operate and easily influenced by load disturbance and parameter uncertainties.

The conventional direct torque control (DTC) strategy is useful for overcoming these disadvantages of FOC [3,4]. The principle of DTC is to use the respective hysteresis band comparators of torque and stator flux to determine errors, and then to select the stator voltage vector according to the switching table to control the torque and stator flux directly. DTC doesn’t require complex decoupling calculation and is easy to implement due to its simple structure. However, DTC has some drawbacks during induction motor operation, such as large ripples in torque and flux at low speeds. In order to reduce ripples in torque and flux, a discrete space vector modulation (DSVM) has been proposed [5] by means of a new switching table. Since its sampling period is subdivided, this new switching table will be more complex, with the result that its DSVM takes more time to calculate and requires a greater sampling time than that of the DTC does. As
an alternative, space vector modulation (SVM) is incorporated with DTC for induction motor drives, as described in [6-9], to provide higher control resolution and help improve the drive’s behavior by a constant inverter switching frequency. For example, Lascu et al. [7] and Lai and Chen [8] proposed a simple proportional-integral controller design by a SVM-based DTC approach to improve torque and flux control performance and reduce the total harmonic distortion of the induction motor drive. However, this simple controller design based on SVM and DTC techniques is sensitive to parameter variations and load disturbance.

Aside from these considerations, a sliding mode controller (SMC) is a nonlinear, high-speed switching, feedback control strategy that provides an effective and robust approach for controlling nonlinear plants [10-12]. Since power converters for ac drives are inherent switching devices, it is worth considering the SMC as a solution for generating discontinuous control laws. Recently, several solutions that integrate SMC and DTC principles have been proposed by [13-17] for the high-performance control of induction motor drives. For example, Barambones and Garrido [13] attempted to improve the speed control performance of an indirect field oriented induction motor drive system by using a sliding mode controller under the influences of load uncertainties and disturbances. Reddy et al. [14] proposed an integral switching surface sliding mode speed controller cooperated with hybrid space vector PWM methods for direct torque controlled induction motor to reduce inverter switching losses and the steady state ripple in torque and flux. However, the SMC easily produces a chattering phenomenon due to its discontinuous switching control. The SMC’s high-speed switching command may detrimentally excite the system’s unmodelled dynamics. One way to solve this problem is to use a boundary layer [10, 14]. Furthermore, some research has been carried out for the design of speed sensorless control schemes [7,17]. In these schemes the speed is estimated according to the measurement of stator voltages and current. However, the estimation is usually complex and heavily dependent on machine parameters.

This paper proposes a hybrid sliding mode flux and torque control using a SVM-based DTC approach to achieve high-performance speed control under the influence of load disturbance and parameter uncertainties. This control scheme applies the two separate speed and flux closed-loop controller designs in synchronously rotating reference frames; the control efforts of these two designs are then directly fed through SVM to produce an adequate switching signal for the inverter by inversely transforming the control command from synchronously rotating reference frames to stationary reference frames. This strategy maintains the advantage of a simple approach and dramatically reduces the ripples in torque and flux. This also makes hybrid sliding mode controllers more attractive for induction motor control applications. Finally, the effectiveness of the proposed control scheme under load disturbance and parameter uncertainties is verified by simulation results and comparison with a conventional (DTC) approach and three proportional-integral controllers using a SVM-based DTC approach.

2. Mathematical Model of Induction Motor. A squirrel-cage induction motor is adopted in this paper. Based on three-phase to two-phase (d-q axis) transformation, the dynamics of the induction motor can be described by the following equations in synchronously rotating reference frames, denoted by the superscript $e$ [18]:

Stator and rotor voltage equations are

$$v_{ds}^e = r_s i_{ds}^e - \omega_e \lambda_{qs}^e + \dot{\lambda}_{ds}^e$$  \hspace{1cm} (1)

$$v_{qs}^e = r_s i_{qs}^e + \omega_e \lambda_{ds}^e + \dot{\lambda}_{qs}^e$$  \hspace{1cm} (2)

$$0 = r_r i_{dr}^e - (\omega_e - \omega_r) \lambda_{qr}^e + \dot{\lambda}_{dr}^e$$  \hspace{1cm} (3)
SLIDING MODE CONTROLLER DESIGN OF INDUCTION MOTOR

0 = r_r i_{qr}^e + (\omega_e - \omega_r)\lambda_d r^e + \dot{\lambda}_q r

and the mechanical equations are represented as

\[ T_e = \frac{3P}{4}(\lambda_{ds}^e i_{qs}^e - \lambda_{qs}^e i_{ds}^e) \]  

\[ T_e = J \frac{d}{dt}\omega_{rm} + B\omega_{rm} + T_L \]

In order to reduce control law dependence on the motor parameters, the DTC approach usually adopts the stationary frame, the \( d \)-axis of which is aligned in the direction of the stator winding of phase \( A \); the mathematical equations of the induction motor can be rewritten as follows by assuming \( \omega_e = 0 \). Therefore, stator and rotor voltage equations are

\[ v_{ds} = r_s i_{ds} + \dot{\lambda}_{ds} \]  

\[ v_{qs} = r_s i_{qs} + \dot{\lambda}_{qs} \]  

\[ 0 = r_r i_{dr} + \omega_r \lambda_{qr} + \dot{\lambda}_{dr} \]  

\[ 0 = r_r i_{qr} - \omega_r \lambda_{dr} + \dot{\lambda}_{qr} \]

and the mechanical equations are represented as

\[ T_e = \frac{3P}{4}(\lambda_{ds}^e i_{qs}^e - \lambda_{qs}^e i_{ds}^e) \]  

\[ T_e = J \frac{d}{dt}\omega_{rm} + B\omega_{rm} + T_L \]

According to Eq.(7) and (8), the stationary stator flux can be calculated by

\[ \lambda_{ds} = \int (v_{ds} - r_s i_{ds})dt \]  

\[ \lambda_{qs} = \int (v_{qs} - r_s i_{qs})dt \]

and the resultant stator flux is shown in Figure 1 and described by

\[ \lambda_s = \sqrt{(\lambda_{ds})^2 + (\lambda_{qs})^2} \]  

\[ \theta_e = \tan^{-1}\left(\frac{\lambda_{qs}}{\lambda_{ds}}\right) \]

Furthermore, based on the field-orientation control principle [13, 16], the resultant stator flux linkage, \( \lambda_s \), is aligned in the direction of the \( d \)-axis of the synchronously rotating reference frames, i.e., \( \lambda_{ds}^e = \lambda_s \) and \( \lambda_{qs}^e = 0 \), to reduce the number of variables by one. Therefore, Eq.(1) can be simplified as follows:

\[ \dot{\lambda}_s = -r_s i_{ds}^e + v_{ds}^e \]  

Similarly, Eqs. (2) and (5) can be rewritten as

\[ v_{qs}^e = r_s i_{qs}^e + \omega_e \lambda_s \]

\[ T_e = \frac{3P}{4}\lambda_s i_{qs}^e \]
Substituting Eqs. (18) and (19) into the electromechanical dynamic equation of Eq.(6) yields the following compound speed dynamics equation:

$$\dot{\omega}_{rm} + \left( \frac{B}{J} + \frac{k_T P \lambda_s}{2 J r_s} \right) \omega_{rm} + \frac{T_L}{J} = \frac{k_T}{J r_s} v_{qs}^e$$  (20)

where \( k_T = \left( \frac{3P}{4} \right) \lambda_s \) and \( \omega_e \) is approximated by \( P \omega_{rm}/2 \). It is observed that \( v_{ds}^e \) and \( v_{qs}^e \) can control the induction motor’s flux and speed dynamics, respectively. Thus, they will be directly applied to the controller design in the next section.

3. **Sliding Mode Controller Design.** Figure 2 presents a block diagram of speed control by two sliding mode direct torque controlled drives, based on the SVM technique. Speed control of the induction motor is realized by the combined control strategies of direct flux and torque SMC according to the field-orientation control principle [15, 18]. That is, using Eqs. (17) and (20), the controller designs are derived from the dynamic equations of the induction motor in the synchronously rotating reference frames. Then, the control inputs of \( v_{ds}^e \) and \( v_{qs}^e \) will be transformed to stationary reference frames for application of the SVM technique. The detailed controller design is described in the following subsection.

3.1. **Flux sliding mode controller design.** Based on flux dynamic Eq.(17), define the flux tracking error to be the difference between the desired resultant stator flux \( (\lambda_s^*) \) and the actual resultant stator flux \( (\lambda_s) \), as below:

$$e_{\lambda} = \lambda_s^* - \lambda_s$$  (21)

Taking the derivative of the previous equation with respect to time yields

$$\dot{e}_{\lambda} = -\dot{\lambda}_s = r_s i_{ds}^e - v_{ds}^e$$  (22)

because the command of the desired resultant stator flux is constant. Considering the stator resistance uncertainty, Eq.(22) can be rewritten in the following form:

$$\dot{e}_{\lambda} = r_s n_{ds} i_{ds}^e - u_{\lambda} + d(t)$$  (23)
where \( u_\lambda = v^e_{ds} \), \( d(t) = \Delta r_s i^e_{ds} \), \( r_{sn} \) is the nominal value of the stator resistance, and \( \Delta r_s \) is the uncertainty of stator resistance with respect to \( r_{sn} \).

Next, a flux sliding function is defined in the following integral form:

\[
s_\lambda(t) = e_\lambda(t) + \int_0^t k_\lambda e_\lambda(\tau) d\tau (24)
\]

where \( k_\lambda \) is the design constant gain and the sliding surface is

\[
s_\lambda(t) = e_\lambda(t) + \int_0^t k_\lambda e_\lambda(\tau) d\tau = 0 (25)
\]

Then, taking the derivative of the sliding surface with respect to time yields

\[
\dot{s}_\lambda = \dot{e}_\lambda + k_\lambda e_\lambda = 0 (26)
\]

Substituting Eq.(23) into Eq.(26) gives

\[
\dot{s}_\lambda = k_\lambda e_\lambda + r_{sn} i^e_{ds} - u_\lambda + d(t) = 0 (27)
\]

The ideal control input can then be obtained by directly canceling the undesired term and embedding the desired dynamics, as below:

\[
u_\lambda = k_\lambda e_\lambda + r_{sn} i^e_{ds} + d(t) (28)
\]

which leads to \( \dot{s}_\lambda = 0 \). That is, flux trajectory tracking will be onto the sliding surface \( s_\lambda = 0 \). However, due to the parameter trajectories uncertainties, the sliding mode flux controller should be designed as:

\[
u_\lambda = \ddot{u}_\lambda + \beta_\lambda sgn(s_\lambda) (29)
\]

where \( \ddot{u}_\lambda = k_\lambda e_\lambda + r_{sn} i^e_{ds} \) is the best approximation control input of Eq.(28) according to the known parameters, \( \beta_\lambda \) is the switching gain, and \( sgn(\cdot) \) is the sign function. Based on the flux controller in Eq.(29), the design of the sliding mode direct flux control can be proposed by the following theorem.

**Theorem 3.1.** Consider the flux dynamic equation of an induction motor given by Eq.(23). Then, if the gain \( k_\lambda \) is chosen to be strictly positive and the gain \( \beta_\lambda \) is designed to meet the criterion of \( \beta_\lambda \geq |d(t)| + \eta_\lambda \), the flux control law in Eq.(29) can enforce the flux dynamic onto the flux sliding surface so that the flux tracking error \( e_\lambda = \lambda^*_s - \lambda_s \) tends to zero as time tends to infinity.

**Proof:** Define the Lyapunov function candidate for the direct flux control to be:

\[
V = \frac{1}{2} s^2_\lambda (30)
\]
Its time derivative is calculated as:

$$\dot{V}_\lambda = s_\lambda \dot{s}_\lambda = s_\lambda [k_\lambda e_\lambda + r_{sn} \dot{i}_{qs}^* - u_\lambda + d(t)]$$  \hspace{1cm} (31)

Substituting Eq.(29) into (31) induces:

$$\dot{V}_\lambda = s_\lambda [d(t) - \beta_\lambda sgn(s_\lambda)] \leq -[\beta_\lambda - |d(t)|]|s_\lambda| \leq -\eta_\lambda |s_\lambda|$$  \hspace{1cm} (32)

where $\eta_\lambda$ is the designated positive constant. If the gain $\beta_\lambda$ is designed to meet the following criterion

$$\beta_\lambda \geq |d(t)| + \eta_\lambda$$  \hspace{1cm} (33)

the $\dot{V}_\lambda$ is negative definite. When $V_\lambda$ is clearly positive definite and $\dot{V}_\lambda$ is negative definite, $s_\lambda(t)$ asymptotically converges to zero; i.e., $s_\lambda(t) \to 0$ as $t \to \infty$. This means that the flux trajectory starting off the sliding surface $s_\lambda(t) = 0$ must reach it in a finite time which is controlled by the designate parameter $\eta_\lambda$, and then will remain on this flux sliding surface, which is called the flux sliding mode. When the flux sliding mode occurs on the sliding surface in Eq.(25), $s_\lambda(t) = \dot{s}_\lambda(t) = 0$. Therefore, the relationship between $s_\lambda(t)$ and $e_\lambda$ in Eq.(26), considering $s_\lambda(t) \to 0$ as $t \to \infty$, gives $e_\lambda(t) \to 0$ as $t \to \infty$, i.e., asymptotic flux trajectory tracking if the gain $k_\lambda$ is chosen to be strictly positive. This completes the proof.

Furthermore, in order to reduce the chattering performance, the boundary layer approach of the flux control law Eq.(29) can be directly obtained as below:

$$u_\lambda = k_\lambda e_\lambda + r_{sn} \dot{i}_{qs}^* + \beta_\lambda sat(s_\lambda/\phi_\lambda)$$  \hspace{1cm} (34)

where sat($\cdot$) is the saturation function [10] and $\phi_\lambda$ is the boundary layer thickness.

3.2. Sliding mode speed controller design. The mechanical dynamic of Eq.(20) can be rewritten in the following form:

$$\dot{\omega}_rm = -a\omega_{rm} - f + bv_{qs}^e$$  \hspace{1cm} (35)

where $a = B/J + (3P^2/\lambda_s^2)/(8 \cdot J \cdot r_s)$; $f = T_L/J$; and $b = (3P \cdot \lambda_s)/(4 \cdot J \cdot r_s)$. Taking into consideration of load disturbances and parameter uncertainties, the previous equation can be expressed as:

$$\dot{\omega}_rm = -(a_n + \Delta a)\omega_{rm} - (f_n + \Delta f) + bu_\omega$$  \hspace{1cm} (36)

where $u_\omega = v_{qs}^e$; $a_n$ and $f_n$ represent the known nominal values of the parameters $a$ and $f$ with the individual uncertainties of $\Delta a$ and $\Delta f$, respectively; $b$ is the control gain, and its value is within the region of $b_{\min} \leq b \leq b_{\max}$. Define the speed tracking error as:

$$e_\omega = \omega_{rm}^* - \omega_{rm}$$  \hspace{1cm} (37)

In addition, define an integral speed sliding function as:

$$s_\omega(t) = e_\omega(t) + \int_0^t k_\omega e_\omega(\tau)d\tau$$  \hspace{1cm} (38)

where $k_\omega$ is the designed constant gain. Then, let the speed sliding surface to be

$$s_\omega(t) = e_\omega(t) + \int_0^t k_\omega e_\omega(\tau)d\tau = 0$$  \hspace{1cm} (39)

Taking the derivative of the above sliding surface with respect to time yields

$$\dot{s}_\omega = \dot{e}_\omega + k_\omega e_\omega = 0$$  \hspace{1cm} (40)

Substituting Eq.(36) into Eq.(40) gives

$$\dot{s}_\omega = \dot{\omega}_{rm}^* - (a_n + \Delta a)\omega_{rm} + (f_n + \Delta f) - bu_\omega + k_\omega e_\omega = 0$$  \hspace{1cm} (41)
An ideal input control law is specified in Eq.(41) by canceling the unexpected terms and embedding the desired terms:

$$bu_o = \omega_{rm}^* + (a_n + \Delta a)\omega_{rm} + (f_n + \Delta f) + k_o e_o$$  

(42)

A best approximation input control without uncertainties is expressed as:

$$b\tilde{u}_o = \dot{\omega}_o = \omega_{rm}^* + a_n\omega_{rm} + f_n + k_o e_o$$

(43)

Then, the sliding mode speed controller is designated as:

$$\hat{b}u_o = \dot{\omega}_o + \beta sgn(s_o)$$  

(44)

where $\hat{b} = \sqrt{b_{min} \cdot b_{max}}$, $\beta$ is the switching gain, and $sgn(\cdot)$ is the sign function. Based on the speed controller in Eq.(44), the design of the sliding mode speed controller can be proposed by the following theorem.

**Theorem 3.2.** Consider the speed dynamic equation of the induction motor given by Eq.(36). Then, if the gain $k_o$ is chosen to be strictly positive and the gain $\beta_o$ is designed to meet the criterion of $\beta_o \geq \alpha |\Delta a||\omega_{rm}| + \alpha|\Delta f| + (\alpha - 1)|\dot{U}_o| + \alpha n_o$, the speed control law Eq.(44) can enforce the speed tracking error dynamic onto the speed sliding surface so that the speed tracking error $e_o = \omega_{rm}^* - \omega_{rm}$ tends to zero as time tends to infinity.

**Proof:** Define the Lyapunov function candidate for the speed control to be $V_o = (s_o)^2/2$ with its time derivative calculated as:

$$\dot{V}_o = s_o[\dot{\omega}_{rm}^* + (a_n + \Delta a)\omega_{rm} + (f_n + \Delta f) - bu_o + k_o e_o]$$

(45)

Substituting Eqs. (43) and (44) into Eq.(45) leads to:

$$\dot{V}_o \leq -b\tilde{b}^{-1} \beta_o - |\Delta a||\omega_{rm}| - |\Delta f| - (1 - b\tilde{b}^{-1})|\dot{U}_o||s_o| \leq -\eta_o |s_o|$$

(46)

if $\beta_o$ is chosen to meet the following condition:

$$\beta_o \geq \frac{\hat{b}}{b} |\Delta a||\omega_{rm}| + \frac{\hat{b}}{b} |\Delta f| + |(\frac{\hat{b}}{b} - 1)|\dot{U}_o| + \frac{\hat{b}}{b} \eta_o$$

(47)

where $\alpha = \sqrt{b_{max}/b_{min}} > 1$. Assume that the conditions of $(1/\alpha) \leq (\hat{b}/b) \leq \alpha$ and $(1/\alpha - 1) \leq (\hat{b}/b - 1) \leq (\alpha - 1)$ are both satisfied. Thus, Eq.(47) can be rewritten as:

$$\beta_o \geq \alpha |\Delta a||\omega_{rm}| + \alpha|\Delta f| + (\alpha - 1)|\dot{U}_o| + \alpha n_o$$

(48)

so that $\dot{V}_o$ is guaranteed to be negative definite. Like the sliding mode flux control, when $V_o$ is clearly positive definite and $V_o$ is negative definite, $s_o(t)$ asymptotically converges to zero; i.e., $s_o(t) \to 0$ as $t \to \infty$. Furthermore, when the sliding mode is used for speed control, $s_o(t) \to 0$ as $t \to \infty$ will induce $e_o(t) \to 0$ as $t \to \infty$, i.e., asymptotic speed trajectory tracking if the gain $k_o$ is chosen to be strictly positive. This completes the proof.

Finally, the speed sliding mode controller is summarized as:

$$u_o = \tilde{b}^{-1}[\dot{\omega}_{rm}^* + a_n\omega_{rm} + f_n + k_o e_o + \beta_o sgn(s_o)]$$

(49)

In order to reduce the chattering phenomenon, the new control input according to the boundary layer approach can be directly presented as:

$$u_o = \tilde{b}^{-1}[\dot{\omega}_{rm}^* + a_n\omega_{rm} + f_n + k_o e_o + \beta_o sat(s_o/\phi_o)]$$

(50)

where $sat(\cdot)$ is the saturation function and $\phi_o$ is the boundary layer thickness for the speed control. Finally, the control inputs of $v_{ds}^e$ and $v_{qs}^e$ obtained by the control law of
Eqs. (29) and (44), respectively, need to be transformed to stationary reference frames by Eq.(51) for application of the SVM technique, as shown in Figures 1 and 2.

\[
\begin{bmatrix}
    v_{ds} \\
    v_{qs}
\end{bmatrix} = \begin{bmatrix}
    \cos (-\theta_e) & \sin (-\theta_e) \\
    -\sin (-\theta_e) & \cos (-\theta_e)
\end{bmatrix} \begin{bmatrix}
    v_{ds}^e \\
    v_{qs}^e
\end{bmatrix}
\]

4. **Simulation Results.** In order to evaluate the performance of the proposed controller designs mentioned above, some different cases of load disturbance and parameter uncertainties are considered in the simulation studies; we also compare the proposed designs with conventional DTC and PIC controllers. Consider a squirrel-cage induction motor of 1 horsepower whose parameters are given in Table 1. The PI controller gain of the conventional direct torque controllers are designed to be \(K_{pw} = 1.5819\) and \(K_{iw} = 129.6\). Also, the three PI controller gains based on SVM-based DTC approach are chosen to be \((K_{P\omega} = 1.5819, K_{I\omega} = 129.6)\), \((K_{P\lambda} = 192, K_{I\lambda} = 9612)\), and \((K_{PT} = 3.1667, K_{TT} = 6076.8)\) for speed, flux, and torque control, respectively. The gains of the proposed SMC controllers are designed to be \(k_\omega = 0.5, k_\lambda = 2\); both reaching time parameters \((\eta_\lambda, \eta_\omega)\) of speed and flux control are set as 75. Furthermore, for simulation studies, the system parameter uncertainties of \(\Delta r_s, \Delta a,\) and \(\Delta f\) are considered to vary within \(\pm20\%\) from the corresponding parameter nominal value.

<table>
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<th>Description</th>
<th>Value</th>
<th>Description</th>
<th>Value</th>
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</tr>
<tr>
<td>Rated current</td>
<td>3.1 A</td>
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<td>Rotor resistance</td>
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<td>Viscous friction coefficient</td>
<td>0.00825 N-m/(rad/sec)</td>
</tr>
</tbody>
</table>

Table 1. Parameters and data of the squirrel-cage induction motor

Figure 3 shows the simulation results for speed control of 1600 rpm by the proposed SMC using SVM-based DTC approach, according to a cycloidal speed trajectory [17]. From Figure 3(a) and Figure 3(b), it can be observed that the sliding mode controller has an excellent speed response and its speed tracking error in steady-state is nearly zero. Figure 3(c) shows that a required torque is generated for increasing the speed during the lunching period of induction motor operation, but reduced to small value for the constant speed operation. From Figure 3(d), it can also be found that the stator flux linkage quickly reaches the flux command circle in the counter-clockwise direction from the zero value. Furthermore, Figure 4 also demonstrates the simulation results of speed control by comparing with the DTC and PIC approaches. It can be observed that the DTC and PIC approaches present a large overshoot response, but the proposed SMC controller provides a very smooth and stable speed response without the overshoot phenomenon. The conventional DTC approach obviously produces a torque ripple. However, the PIC and proposed SMC approaches based on the SVM technique can largely diminish the torque ripple.

Figure 5 shows the speed error and torque response for a low speed control of 35 rpm without parameter uncertainties. It is noted that the conventional DTC approach brings a large speed tracking error and obviously induces the speed and torque ripple effects. Among these three approaches, the proposed SMC provides an excellent speed and torque error response. Figure 6 shows the speed tracking response for the sinusoidal wave trajectory of \(5\sin(10t)\). Both DTC and PIC approaches produce poor speed tracking control performance. As for the proposed SMC approach, it has a slight speed tracking
error only initially and then leads to excellent speed tracking control. To evaluate the speed-tracking performance for the parameter uncertainties, simulation studies for low speed control with the stator resistance increased by 100% at $t = 0.5$ sec are shown in Figure 7. Both DTC and PIC approaches lead to a poor speed tracking response because stator resistance plays an important role for low speed control. However, the proposed SMC approach not only maintains a consistent speed tracking performance but also reduces the torque ripple even when stator resistance is suddenly changed.

Figure 3. Simulation results for sliding mode speed control of 1600 rpm

Figure 8 shows the simulation results for a speed control of 400 rpm with load disturbance changes of 2 N-m at $t = 0.5$ sec. Note that the speed tracking error for the DTC and PIC approaches will suddenly increase by about 8 rpm when the load disturbance is applied. By contrast, the proposed SMC controller simply increases the speed tracking error of 2 rpm and then quickly compensates for this disturbance to recover the desired speed tracking performance. Moreover, the SMC approach significantly attenuates the torque ripple effect.

5. Conclusions. To enhance the control performance of an induction motor, this paper presents a hybrid sliding mode controller design based on the DTC and SVM methods to achieve high-performance speed control. Simulation studies demonstrated that the proposed SMC controller, as compared with the DTC and PIC approaches, can produce...
a better response in terms of flux, torque, and speed control for different speed commands, load disturbances, and parameter uncertainties. Clearly, the proposed controller can largely reduce the ripple effect that is usually produced by the conventional DTC approach. By contrast, the controlled effects of the DTC and PIC approaches are easily influenced by system uncertainties. In our future work, the proposed sliding mode controller will be implemented in a real control system with digital signal processing (DSP) chips.

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Figure 7. Speed-tracking error for speed control of 35 rpm with stator resistance increased by 100% at t=0.5 sec.

Figure 8. Speed-tracking error for speed control of 400 rpm with load disturbance changed to 2 N-m at t=0.5 sec.


