International Review on Modelling and Simulations (I.RE.MO.S.), Vol. 6, n. 3
June 2013

A New State-dependent of Sliding Mode Control for Three-Phase Induction Motor Drives

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Abstract – *Chattering is known as a main obstacle in realizing sliding mode control. In this study, new robust and chattering suppression of sliding mode control method is investigated. First, an integral sliding mode control is developed. Then, a simple smooth function and switching gain with state-dependent based on fast sigmoid function is introduced. This method allows chattering reduction, as well as maintaining the robust characteristics of sliding mode control. The effectiveness of the proposed algorithm is demonstrated using simulation speed control of indirect field-oriented three-phase induction motor drives with regard to external load disturbances.*

***Keywords**: chattering, fast sigmoid function, induction motor, state-dependent sliding mode*

Nomenclature

B	*Friction coefficient, Nm/(rad/sec)*
$e(t)$	*Speed error, rpm*
d	*lumped uncertainties*
i_{ds}, i_{qs}	*d and q axis stator currents, A*
i_{dr}, i_{qr}	*Rotor current in d and q axis, A*
J	*Inertia, kg-m^2*
K_T	*torque constant*
L_{ls},	*Stator-leakage inductance, H*
L_{lr}	*Stator-referred rotor-leakage inductance, H*
L_m	*Magnetizing inductance, H*
L_s	*Stator selfinductance,H*
L_r	*Stator-referred rotor selfinductance,H*
R_s	*Stator resistance, Ω*
R_r	*Stator-referred rotor-phase resistance, Ω*
T_r	*Rotor time constant*
V_{ds}, V_{qs},	*d and q axis stator voltage, V*
ω_r	*Rotor speed, rpm*
ω_r^*	*Rotor speed reference, rpm*
φ_{qs}, φ_{ds}	*Stator flux linkage in q and d axis, V-s*
φ_{qr}, φ_{dr}	*Rotor flux linkage in q and d axis, V-s*
β	*switching gain*

I. Introduction

Sliding mode control has successfully implement in wide area applications such as in mobile robot [1], DC motor control [2], power supplies [3], AC motors [4-7] and many other applications. In AC motor drives, sliding mode control is implemented as outer loop in speed controller [8], as inner loop in current controller [5] as speed observer in sensorless methods [9-11] and other applications [12]. Sliding mode is a nonlinear control method that alters the dynamics of a nonlinear system by application of a high frequency switching control. Researchers in the field of sliding mode control are striving to overcome the problem associated with variable structure control, that so-called chattering phenomenon. Chattering is the high-frequency oscillations of the controller output, brought about by the high speed switching necessary for the establishment of a sliding mode. These oscillations are caused by the high-frequencies of a sliding mode controller exiting unmodeled dynamics in the closed loop system. Unmodelled dynamics may refer to sensors, actuator data processor neglected in the principles modelling process since they are generally significantly faster than the main system dynamics. However since ideal sliding mode systems are infinitely fast, all system dynamics should be considered in the control design. In practical implementation, chattering is highly undesirable because it may excite unmodeled high-frequency plant dynamics, and this can result in unforeseen instability [13]. This harmful phenomenon often leads to undesirable result such as low control accuracy, high wear of moving mechanical parts and in power electronics, high frequency can lead to high losses [14]. Without proper treatment in the control design, chattering can be a major obstacle in implementing sliding mode control in wide range applications.

In addition, chattering level is influenced by the value of the switching gain. Switching gain is employed in sliding mode as upper bound of uncertainties. As the system discover from a large uncertainty, or unknown uncertainties, the higher value of switching gain must be applied. That means, the magnitude of chattering is proportional to switching gain which is also represents as an upper bound of uncertainties of the system. This upper bound of uncertainties which is required in conventional sliding mode is difficult to obtain precisely. If the bound is selected too large, the hitting control law will result in serious chattering phenomenon. On the other hand, if the bound is selected too small, the stability conditions may not be satisfied. Thus, the idea in choosing the adequate switching gain is by reducing the value of switching gain in order to decrease the amplitude of chattering at the same time preserving the existence of sliding mode control. To support this idea, a state-dependent gain method is proposed in this study. This method is originated from [14] which is use the switching gain is depends on the absolute error of the systems and from [15] which proposed a switching gain depend on absolute sliding variable. In this research, the proposed state-dependent switching gain is depending on the absolute value of sigmoid function.

In order to reduce this high frequency oscillation, the discontinuous function is replaced with a smooth function. In [16] hyperbolic tangent function and saturation function is used to alleviate the discontinuous function and applied to position servo systems, in [17, 18] modified hyperbolic tangent function is designed with self-tuning law algorithm, and the widely used technique is utilized boundary layer [19, 20]. Different method of boundary layer technique has been used in literature; however, the intention is to limit the use of discontinuous function during the operation system. The larger the boundary layer width, the smoother the control signal. Even though the boundary layer design is to alleviate the chattering phenomenon, however it no longer drives the system state to the origin, but have a small residual set around the origin [14]. The size of residual set is determined by the width of boundary layer; the larger width of boundary layer, the larger set of residual set. As consequences, there exit a design conflict between requirement of smoothness of control signal an on control accuracy. For smoothness of the control signal, a large boundary layer width is preferred, but for better control accuracy, a small boundary layer is preferred. Therefore this study proposes that the boundary layer width be proportional to the modulus of the system state, sigmoid function. This design automatically adjust the boundary layer width based on the system condition, and will be more capable when dealing with unexpected uncertainties.

II. Indirect Field-Oriented Controlled of Three-phase Induction Motor

To regulate these induction motor in high performance application, one of the most popular technique is indirect field oriented control method [21]. It allows, by means a co-ordinate transformation, to decouple the electromagnetic torque control from the rotor flux, and hence manage induction motor controlled as separately exited DC motor. Fig. 1 shows the block diagram of the drives system in real application. The control is divided into two control loops; inner current loop and outer speed control loop. The three-phase squirrel cage induction motor in synchronously rotating reference frame can be represent in mathematical form [22] as (1) –(8):

$$V_{qs} = R_s i_{qs} + \frac{d\varphi_{qs}}{dt} + \omega_e \varphi_{ds} \quad (1)$$

$$V_{ds} = R_s i_{ds} + \frac{d\varphi_{ds}}{dt} - \omega_e \varphi_{qs} \quad (2)$$

$$V_{qr} = R_r i_{qr} + \frac{d\varphi_{qr}}{dt} + (\omega_e - \omega_r)\varphi_{dr} \quad (3)$$

$$V_{dr} = R_r i_{dr} + \frac{d\varphi_{dr}}{dt} + (\omega_e - \omega_r)\varphi_{qr} \quad (4)$$

where , and the flux equation:

$$\varphi_{qs} = L_{Ls} i_{qs} + L_m (i_{qs} + i_{qr}) \quad (5)$$

$$\varphi_{qr} = L_{lr} i_{qr} + L_m (i_{qs} + i_{qr}) \quad (6)$$

$$\varphi_{ds} = L_{ls} i_{ds} + L_m (i_{ds} + i_{dr}) \quad (7)$$

$$\varphi_{dr} = L_{lr} i_{dr} + L_m (i_{ds} + i_{dr}) \quad (8)$$

where V_{qs}, V_{ds} are the applied voltages to the stator, i_{ds}, i_{qs}, i_{dr}, i_{qr} are the corresponding d and q axis stator current and rotor currents. φ_{qs}, φ_{qr}, φ_{ds}, φ_{dr}, are the stator and rotor flux component, Rs, Rr are the stator and rotor resistances, L_{ls}, L_{lr} denotes stator and rotor inductances, whereas L_m is the mutual inductance. The electromagnetic torque equation is:

$$T_e = \frac{3}{2}\frac{P}{2}\frac{L_m}{L_r}(\varphi_{dr} i_{qs} - \varphi_{qr} i_{ds}) \quad (9)$$

where *P*, denote the pole number of the motor. If the vector control is fulfilled, the q-axis component of the rotor field φ_{qr} would be zero. Then the electromagnetic torque is controlled only by *q*-axis stator current and becomes:

$$T_e = \frac{3}{2}\frac{P}{2}\frac{L_m}{L_r}(\varphi_{dr} i_{qs}) \quad (10)$$

The rotor flux quantities are estimate using computational rotor time constant, rotor angular velocity and stator current as in (11).

$$\frac{d\theta_e}{dt} = \frac{1}{Tr}\frac{i_{qs}}{i_{ds}} + \omega_r \quad (11)$$

The rotor speed ω_r is compared to rotor speed command ω_r^* and the resulting error is processed in the sliding mode speed controller. The sliding mode speed controller will generate stator q-axis current reference i_{qs}^*. Both reference current in d-axis and q-axis is compared to the feedback from the motor current through Clark and Park Transformation. From the respective error the voltage command signal is generated through PI current controller and converted to two phase voltage through Inverse Park Transformation and fed to Space Vector PWM which generates switching signal for Voltage Source Inverter (VSI). These in turn, control the stator winding current of induction motor, so controlling the speed of the motor. The transformation between stationary *a-b-c* frame, stationary *α-β* frame and synchronously rotating *d-q* frame are describes as the following equation [22].

- Clarke; convert stationary a-b-c frame to stationary α-β frame.

$$\begin{bmatrix} I_\alpha \\ I_\beta \end{bmatrix} = \begin{bmatrix} \frac{2}{3} & \frac{-1}{3} & \frac{-1}{3} \\ 0 & \frac{1}{\sqrt{3}} & \frac{-1}{\sqrt{3}} \end{bmatrix} \begin{bmatrix} I_a \\ I_b \\ I_c \end{bmatrix} \quad (12)$$

- Clarke^{-1}; convert stationary *α-β* frame to stationary *a-b-c* frame.

$$\begin{bmatrix} I_a \\ I_b \\ I_b \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ \frac{-1}{2} & \frac{\sqrt{3}}{2} \\ \frac{-1}{2} & \frac{-\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} I_\alpha \\ I_\beta \end{bmatrix} \qquad (13)$$

- Park; convert stationary *α-β* frame to rotating *d-q* frame.

$$\begin{bmatrix} I_d \\ I_q \end{bmatrix} = \begin{bmatrix} \cos\theta_e & \sin\theta_e \\ -\sin\theta_e & \cos\theta_e \end{bmatrix} \begin{bmatrix} I_\alpha \\ I_\beta \end{bmatrix} \qquad (14)$$

- Park^{-1}; convert rotating *d-q* frame to stationary *α-β* frame.

$$\begin{bmatrix} I_\alpha \\ I_\beta \end{bmatrix} = \begin{bmatrix} \cos\theta_e & -\sin\theta_e \\ \sin\theta_e & \cos\theta_e \end{bmatrix} \begin{bmatrix} I_d \\ I_q \end{bmatrix} \qquad (15)$$

III. Sliding Mode Speed Controller Design

With the advantage of integral sliding mode control, and the practice explain in electric drives systems in [14], this section will derive the sliding mode speed control for induction motor drives. Based on complete indirect field orientation, sliding mode control with integral sliding surface is developed. Under the complete field oriented control, the mechanical equation of three-phase induction motor can be equivalently described as:

$$\mathrm{T_e} = K_T i_{qs} \qquad (12)$$

Where, K_T is the torque constant and defined as follows:

$$K_T = \frac{3}{2}\frac{4}{2}\frac{L_m}{L_r}\varphi_{dr} \qquad (13)$$

Whereas, the mechanical equation of an induction motor can be written as:

$$T_e = J\dot{\omega} + B\omega_m + T_L \qquad (14)$$

Where, J and B are the inertia constant of the induction motor and viscous friction coefficient respectively; T_L is external load; ω_m is the rotor mechanical speed and Te denotes the generated torque of an induction motor. Using (12) into (14), one can obtain;

$$\dot{\omega}_m(t) = (a + \Delta a)\omega_m(t) + (b + \Delta b)i_{qs} + f \qquad (15)$$

Where, $a = -B/J$, $b = K_T/J$ $f = T_L/J$ and , $\Delta a = \Delta B/J$, $\Delta b = \Delta K_t / J$. The tracking speed error is defined as:

$$e(t) = \omega_m(t) - \omega_m^*(t) \qquad (16)$$

where, ω_m^* is a rotor speed reference. Taking derivative of Equation (16) with respect to time yields:

$$\dot{e}(t) = ae(t) + b(u_{qs}(t) + d) \qquad (17)$$

where, d is called lumped uncertainties, defined as:

$$d = \frac{\Delta a}{b}\omega_m(t) + \frac{a}{b}i_{qs} + \frac{f}{b} \qquad (18)$$

and

$$u_{qs}(t) = i_{qs}(t) + \frac{a}{b}\omega_m^* \qquad (19)$$

The sliding variable *S(t)* can be defined with integral component as [23]:

$$S(t) = e(t) - \int (a + bK)e(\tau)d\tau \qquad (20)$$

where, K is a linear feedback gain. When the sliding mode occurs on the sliding surface, then $S(t) = \dot{S}(t) = 0$ and therefore the dynamical behaviour of the tracking problem in Equation (20) is equivalently governed by the following:

$$\dot{e}(t) = (a + bK)e(t) \qquad (21)$$

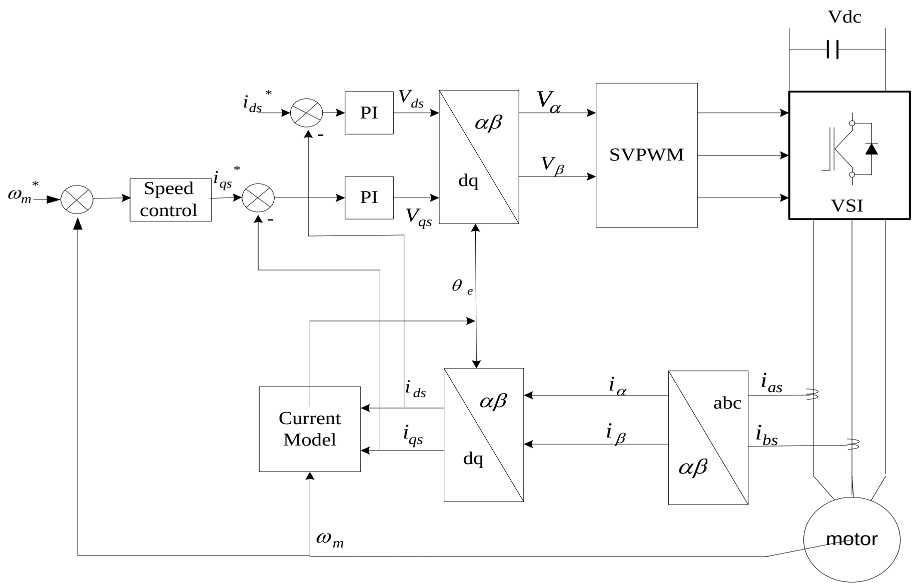


Fig. 1: Overall block diagram for indirect field oriented controlled of induction motor drives

Where, $(a+bK)$ is designed to be strictly negative. Based on the sliding surface (20), and the following assumption,

$$\beta \geq |d(t)| \tag{22}$$

The variable structure controller is design as:

$$u_{qs} = Ke(t) - \beta\,\text{sgn}(S) \tag{23}$$

where β is a switching gain, S is the sliding variable and sgn(.) is the sign function defined as:

$$\text{sgn}(S(t)) = \begin{cases} 1 & if S(t) > 0 \\ -1 & if S(t) < 0 \end{cases} \tag{24}$$

Finally the torque current command or q-axis stator current reference $i^{*}_{qs}(t)$ can be obtained by directly substituting equation (23) into (19).

$$i^{*}_{qs} = Ke(t) - \beta.\text{sgn}(S(t)) - \frac{a}{b}\omega^{*}_{m} \tag{25}$$

Therefore, the sliding mode controller resolves the speed tracking problem for the induction motor, with bounded uncertainties in parameter variation and load disturbances.

IV. State-dependent Sliding Mode Control

Integral sliding mode control has been developed in previous section. The control law is depend on the discontinuous control, signum function which leads to chattering. In this section the development of state-dependent sliding mode control will be explained. Base on equation (25) the signum function is replace by fast sigmoid function as:

$$sgm(\rho', S) = \frac{S}{\rho' + |S|} \tag{26}$$

With this technique, the discontinuous function is eliminated, and the sigmoid function takes place for the whole operation. Where, ρ' is a state-dependent small positive constant the thickness of the boundary layer. The boundary layer is obtained from the proposed state-dependent variable the sigmoid function as:

$$\rho' = \rho_1[(1 - |sgm(\rho', S)|) + \delta_1] \tag{27}$$

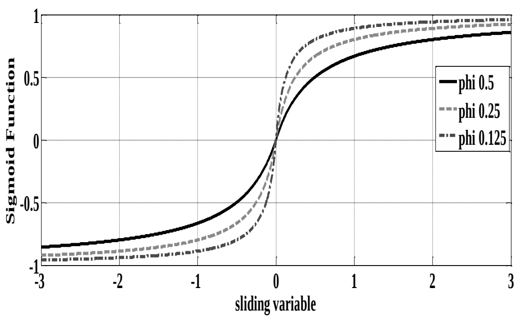


Fig. 2 variable sigmoid function versus sliding variable

Where, δ_1 is sufficiently small and ρ' is positive constant used to adjust the tuning rate of the sigmoid function. This design automatically adjust the boundary layer width based on the system condition, and will be more capable when dealing with unexpected uncertainties. Fig. 2 shows the sigmoid function versus sliding variable function. When sliding mode occur, the sliding variable is set to the origin $S(t)=0$. When there is uncertainties occur, $S(t)$ will be change to a certain value before its drives back to the origin. As from the figure, higher value of sliding variable, gives high value of absolute $sgm(\rho' S)$, but within a sigmoid shape and value from zero to one.

Therefore, when the uncertainties are large, ρ' will produce a small boundary layer for control accuracy and better tracking performance. If there is no uncertainty at the system, sliding variable will stay at the origin as well as the sigmoid function; ρ' will produce constant maximum setting boundary layer. The state-dependent switching gain is design, to be proportional to the absolute sigmoid function as:

$$\beta' = \beta_1(|sgm(\rho', S)| + \delta_2) \tag{28}$$

Where, $\beta_1\delta_2$ is a constant, which should be enough to force the sliding mode to occur. That means, if there are uncertainties in the systems, β' will be increases, and in normal operation, β' will stay at adequate switching gain value. With this state-depending switching technique, the switching gain is depend on the uncertainties of the systems, and there is no overestimated of the boundary layer in the whole operating condition, thus chattering will be reduce. Finally the torque current command or q-axis stator current reference i_{qs}^{*} for the proposed sliding mode controller can be obtained as:

$$i_{qs}^{*} = Ke(t) - \beta'.sgm(\rho' S) - \frac{a}{b}\omega^{*}_{m} \tag{29}$$

With the proposed sliding mode controller, the width of boundary layer and the switching gain are tuned to cause the tracking error to approach zero. Therefore, the β' is exhibit a varying switching gain depend on uncertainties of the system and ρ' exhibit in varying boundary layer in sigmoid function which effectively eliminate input chattering and better tracking performance. With the proposed sliding mode control, no complex algorithm such as exponential function or hyperbolic function involved, so it is easy to implement in fixed point digital signal processor.

V. Results State-dependent Sliding Mode Control

The performance of the proposed sliding mode control is investigated using simulation in MATLAB/SIMULINK. The induction motor use for this simulation is 1.5KW, 1400rpm. The parameter of the motor are, R_s=4.6 Ω R_r=5.66, L_s=0.3153H, L_r=0.3153H, L_m=0.3H and J=0.004kgm^2. The stator q-axis current reference is limit to 5A. The sliding mode controller parameters are: K=-0.2, β=4.0, $\beta_1\delta_2$=2.0, δ_1=0.01, ρ_1=0.5 and the PI speed controller parameters are: K_p=0.5, K_i=0.95 All the parameters are chosen to achieve superior transient control performance and to get the similar performance in term of percentage of overshoot and settling time in rated speed. Therefore, the notion comparison made will be fair and equitable. For simulation results, Fig. 3 shows the results of the speed response and associate control effort (torque current command i_{qs}*), of PI speed control, sliding mode control and state-dependent sliding mode control respectively, in no load conditions. From the figure, good tracking responses are obtained for the entire three controllers during transient and steady state condition. High chattering occurs in control effort of sliding mode controller and significantly reduced by the proposed state-dependent sliding mode control method.

Fig. 4 shows the behaviour of the speed response and associated control effort in load rejection condition. A half rated load, 5.0Nm is applied to the motor during steady state condition in 900rpm. Robust control performance in load rejection behaviour is obvious in sliding controller, followed by state-dependent sliding mode controller. The proposed state-dependent varying sliding mode controller is suppressing the chattering effect completely. Based on the results obtained, the proposed state-dependent sliding mode control can provide robust performance as well as reducing the chattering effect. By means of simulation examples, it has been shown that the proposed control scheme performs reasonably well in practice, and that the speed tracking objective is achieved under load disturbances.

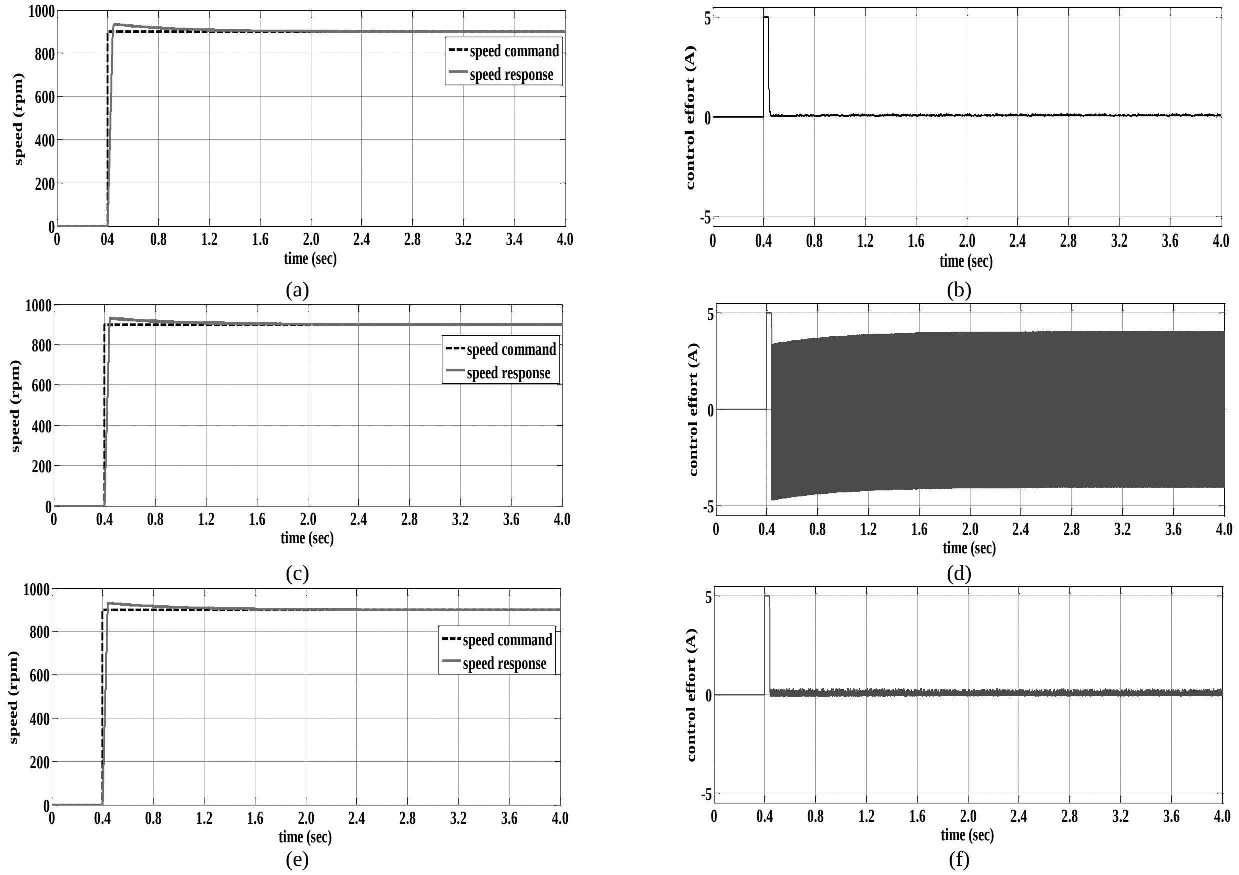
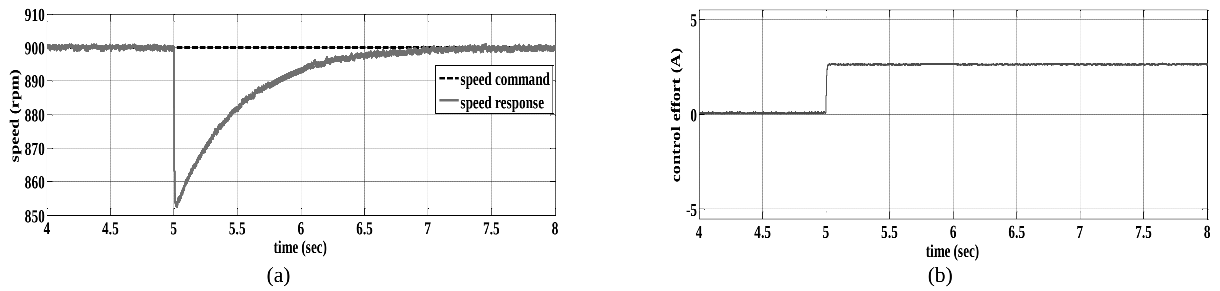


Fig. 3.Speed response and control effort, (a-b) PI controller (c-d) Sliding mode control (e-f) State-dependent Sliding Mode Control



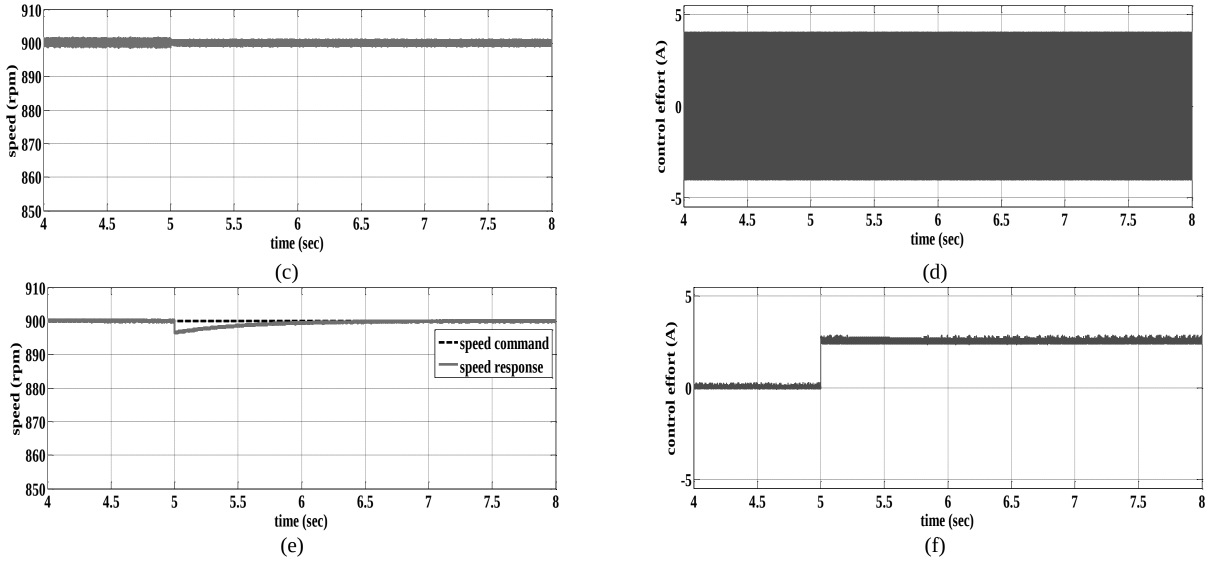


Fig. 4.Speed response and control effort during load disturbances (a-b) PI controller (c-d) Sliding mode control (e-f) State-dependent Sliding Mode Control

Next, is comparison in load rejection behaviour in sliding mode control with state-dependent sliding mode control and constant boundary layer of sliding mode control. Two constants boundary layer are selected, first is 0.5 and second is 0.05 (the maximum and ten times of minimum value in state-dependent boundary layer in this simulation). Fig. 5 shows the result when load 5.0Nm and 2.5Nm is applied in the systems. From the figure, a small boundary layer gives a better tracking performance, however a chattering effect not completely eliminate. In contrast, large boundary layer can reduce chattering effect but gives poor tracking performance. From these results, it is clearly seen that state-dependent boundary layer gives almost similar characteristic for both cases. This is due to state-dependent boundary layer adaption.

Fig. 6 shows the effect of changing speed and load changes at the drives system to parameter ρ' and switching gain β'. First, the motor is run at 900rpm from stand still. Then, 2.5Nm load is applied to the motor at t=3sec. A minimum value of ρ' is obtained during transient to get good tracking performances and increasing to maximum setting of ρ' as speed go to steady state. When load 2.5Nm is applied to the motor, parameter ρ' reduces to a certain value get through the uncertainty and to gives good tracking performances. The parameter β' is increasing during transient and reduce at adequate value during steady state condition and increasing when the load is applied. Therefore the proposed sliding mode controller parameter is changed according to uncertainties of the systems.

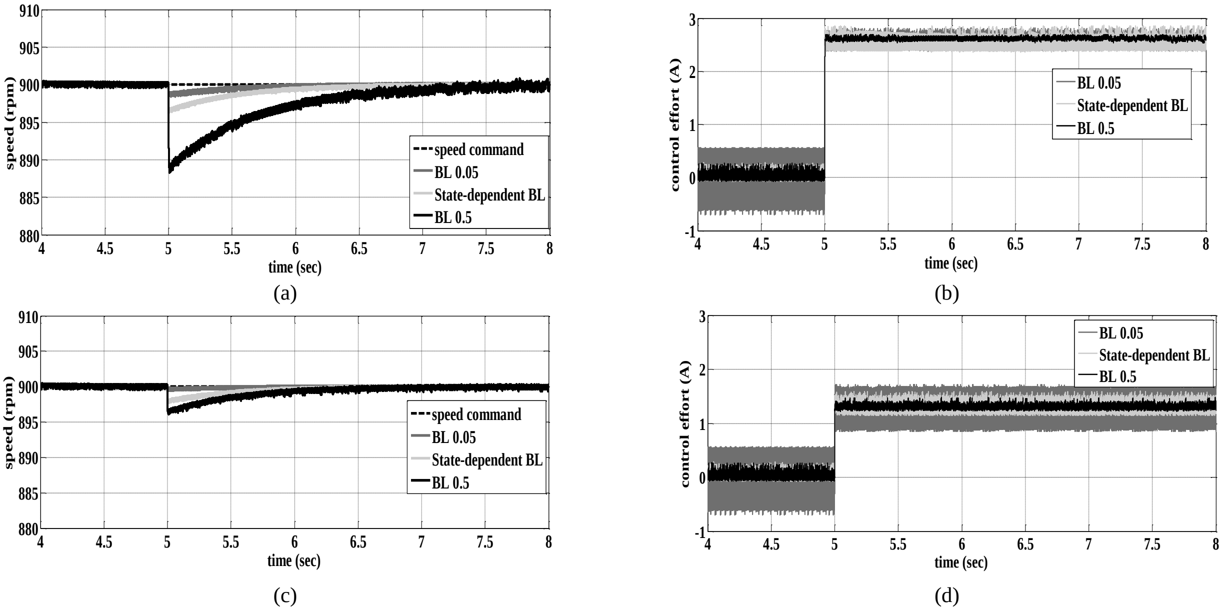


Fig. 5 Speed response and control effort during load disturbances (a-b) with 5.0 Nm (c-d) with 2.5Nm

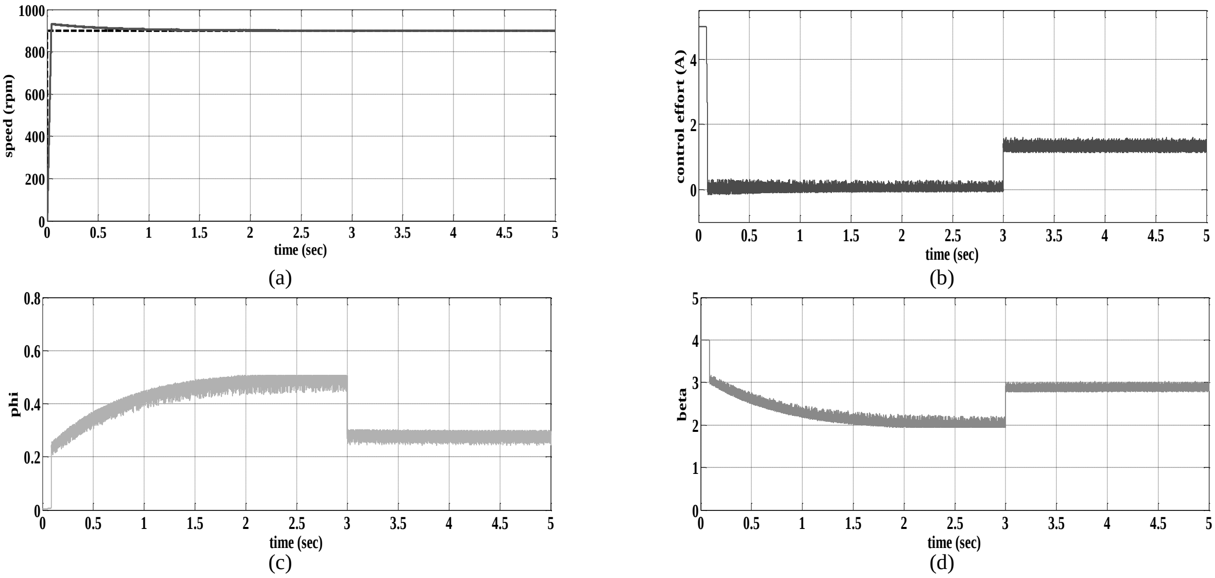


Fig. 6 The effect of speed and load changes to parameter ρ' and β' in state-dependent sliding mode control
(a) Speed response (b) control effort (c) parameter ρ' (d) parameter β'

VI. Conclusion

A new state-dependent varying boundary layer and state-dependent switching gain is proposed for speed control induction motor drive. First, an integral sliding mode, which is theoretically robust with regard to plant uncertainties, is designed. Then a smooth fast sigmoid function is replaced the discontinuous function in sliding mode control law with state-dependent boundary layer is introduced. To overcome over estimation of switching gain in the whole operating condition of sliding mode control, a state-dependent switching gain based on sigmoid function is introduces. Simulation results have shown that the proposed sliding mode controller is robust with regard to load disturbances as well as reduce the chattering phenomenon.

Acknowledgment

The authors wish to thanks Universiti Teknikal Malaysia Melaka for financing the research under the Grant ERGS/2011/FKE/TK02/1-E0001

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