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New Methodology for Asynchronous Motor the Adaptive-Sliding-Mode-Control Capable of High Performance Regulation

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Abstract

A new methodology for the design of adaptive sliding mode control (ASMC) for Asynchronous motor control will be presented in this paper. The sliding mode control (SMC) has become one of the most active branches of control theory that has found successful applications in a variety of engineering systems, such an the electrical motors. The new Adaptive sliding mode control method is compared to other existing techniques. The pros and cons of ASMC controller will be demonstrated by intensive simulation results. It will be shown that the presented controller is with fast tracking capability, less steady state error, and robust to load disturbance.

Keywords: *Adaptive sliding mode control, Asynchronous motor, Controller*

1. Introduction

Asynchronous Machine is complex electromechanical devices which are used in most industrial applications for converting electrical power to mechanical form. The Asynchronous Machine can be energized from adjustable-speed ac drives or from constant frequency sinusoidal power supplies.

Nonlinear control theory has been applied to the control of Asynchronous Motors (AM) in much research [1]-[6], such as input–output linearization strategy and nonlinear state feedback control [7], since the Asynchronous motor is basically a nonlinearly coupling system [8]. For instance, [9] and [10] presented an adaptive feedback linearization tracking controller for the Asynchronous motor. Furthermore, [11] utilized the feedback linearization approach to design a controller to achieve input–output decoupling, high dynamic performance, and high power efficiency. For example, a feedback-linearizable system obtained by an integrator addition is presented in [12]. However, the parametric deviation will significantly affect the dynamic performance and the stability for practical implementation. Therefore, for the nonlinear feedback control of Asynchronous motors, many studies have also presented the compensators for the influence of the variation of motor parameters. In these studies, an adaptation law to the nonlinear sliding-mode control has been studied, the sliding-mode control can offer many advantages, [13]-[15], such as invariance condition, insensitivity or robustness, and fast dynamic response.

Sliding mode control (SMC), as an effective robust control strategy, has been successfully applied to a wide variety of complex systems and engineering [16], including uncertain systems [17], time-delay systems [18], stochastic systems [19], singular systems [20] and Markovian jump systems [21]. The good performance of the proposed scheme has been tested using a realistic numerical simulation. The steady-state and the transient behavior have been investigated. In both cases (with ASMC and with), the results obtained emphasize the effectiveness of the proposed drive system.

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2. Asynchronous Machine

The asynchronous machine d-q or dynamic equivalent circuit is shown in Figure 1. One of the most popular asynchronous motor models derived from this equivalent circuit is detailed in [22]. According to his model, the modeling equations in flux linkage form are as follows:

a. The stator voltage Equation

$$\begin{cases} u_{sd} = R_s i_{sd} + \dfrac{d\varphi_{sd}}{dt} - \omega_s \varphi_{sq} \\ u_{sq} = R_s i_{sq} + \dfrac{d\varphi_{sq}}{dt} + \omega_s \varphi_{sd} \end{cases} \tag{1}$$

b. The rotor voltage Equations

$$\begin{cases} u_{rd} = 0 = R_r i_{rd} + \dfrac{d\varphi_{rd}}{dt} - (\omega_s - \omega)\varphi_{rq} \\ u_{rq} = 0 = R_r i_{rq} + \dfrac{d\varphi_{rq}}{dt} + (\omega_s - \omega)\varphi_{rd} \end{cases} \tag{2}$$

c. The stator flux Equation

$$\begin{cases} \varphi_{sd} = L_s i_{sd} + L_m i_{rd} \\ \varphi_{sq} = L_s i_{sq} + L_m i_{rq} \end{cases} \tag{3}$$

d. The rotor flux Equation

$$\begin{cases} \varphi_{rd} = L_r i_{rd} + L_m i_{sd} \\ \varphi_{rq} = L_r i_{rq} + L_m i_{sq} \end{cases} \tag{4}$$

e. The mechanical Equation

$$T_e = \frac{PL_m}{3}\left(i_{rd} i_{sq} - i_{rq} i_{sd}\right). \tag{5}$$

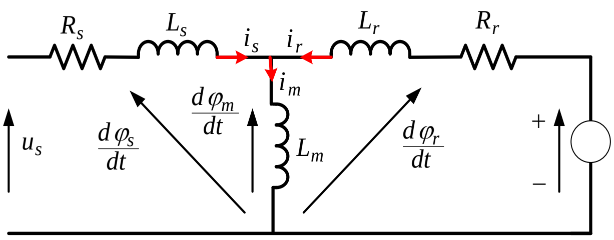

Figure 1. Equivalent Circuit of an Asynchronous Motor

For the development to follow, the mathematical model of asynchronous motor referred to a stationary reference frame, which is denoted by the superscript “dq,” and with d-axis attached on the stator winding of phase “A,” is presented as follows [22].

2.1. Sliding Mode Control

A Sliding mode control (SMC) following a desired linear reference model is designed in the following way. Rewriting the Asynchronous Motors model shown in Eq. (6), in a compact form as:

$$\dot{\chi} = G(\chi) + F_1 u_{sa} + F_2 u_{sb}. \tag{6}$$

where

$$\chi = \begin{bmatrix} i_{sa} & i_{sa} & \varphi_{sa} & \varphi_{sb} \end{bmatrix}^T, F_1 = \begin{bmatrix} \frac{1}{L_s\sigma} & 0 & 1 & 0 \end{bmatrix}^T,$$
$$F_2 = \begin{bmatrix} 0 & \frac{1}{L_s\sigma} & 0 & 1 \end{bmatrix}^T. \tag{7}$$

In order to achieve fast torque response as well as operate in the flux weakened region and maximize the power efficiency for the Asynchronous Motors drive, the torque (T) and the norm of the stator flux linkage $\left(\varphi_{sa}{}^2, \varphi_{sb}{}^2\right)$ are assumed to be the system outputs. Hence, on the basis of the input–output feedback linearization technique, the following variables are introduced;

$$\eta_1(\chi) = \frac{3P}{2}(\varphi_{sa} i_{sb} - \varphi_{sb} i_{sa}),\ \eta_2(\chi) = \varphi_{sa}{}^2 + \varphi_{sb}{}^2. \tag{8}$$

Using Equations (6) and (7), the system model shown in Equation (9) is modified to:

$$\begin{bmatrix} \dot{\eta}_1 \\ \dot{\eta}_2 \end{bmatrix} = \begin{bmatrix} L_f \eta_2 \\ L_f \eta_1 \end{bmatrix} + \begin{bmatrix} 2\varphi_{sa} & 2\varphi_{sb} \\ L_{F_1}\eta_1 & L_{F_2}\eta_2 \end{bmatrix} \begin{bmatrix} u_{sa} \\ u_{sb} \end{bmatrix} \tag{9}$$

where the following notation is used derivative of a function $\eta(\chi) : \Re^n \to \Re.$ along a vector: $G(\chi) = \left(G_1(\chi), G_2(\chi), ..G_n(\chi)\right)$.

$$\begin{cases} L_f \eta(\chi) = \sum_{i=1}^{n} \frac{\partial \eta}{\partial \chi} G_i(\chi) \\ L_f^i \eta = L_f \left(L_f^{i-1} \eta\right) \end{cases}. \tag{10}$$

Using the above notation, one can obtain that:

$$L_f \eta_1(\chi) = \frac{-3P}{2} \varphi_{sb} \left\{ -\left(\frac{R_s}{L_s\sigma} + \frac{R_r}{L_r\sigma} \right) i_{sa} - \omega_r i_{sb} + \frac{\omega_r}{L_s\sigma} \varphi_{sb} \right\} +$$
$$\frac{3P}{2} \varphi_{sa} \left\{ -\left(\frac{R_s}{L_s\sigma} + \frac{R_r}{L_r\sigma} \right) i_{sb} - \omega_r i_{sa} + \frac{\omega_r}{L_s\sigma} \varphi_{sa} \right\} \tag{11}$$

$$L_f \eta_2(\chi) = -2\rho R_s (\varphi_{sa} i_{sa} + \varphi_{sb} i_{sb}) \tag{12}$$

Furthermore,

$$L_{F_1}\eta_1(\chi) = -\frac{3P}{2\sigma L_s}\varphi_{sb} + \frac{3P}{2}i_{sb}$$
$$L_{F_2}\eta_1(\chi) = \frac{3P}{2\sigma L_s}\varphi_{sa} - \frac{3P}{2}i_{sa} \quad (13)$$

Moreover, from Equation (11), the following control inputs are defined:

$$\begin{bmatrix} u_{sa} \\ u_{sb} \end{bmatrix} = \begin{bmatrix} 2\varphi_{sa}u_{sa} + 2\varphi_{sb}u_{sb} \\ L_{F_1}\eta_1 u_{sa} + L_{F_1}\eta_1 u_{sb} \end{bmatrix} \quad (14)$$

Linking Equations (11) and (14) gives :

$$\begin{bmatrix} \dot{\eta}_1 \\ \dot{\eta}_2 \end{bmatrix} = \begin{bmatrix} L_f\eta_2 \\ L_f\eta_1 \end{bmatrix} + \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}\begin{bmatrix} v_{sa} \\ v_{sb} \end{bmatrix} \quad (15)$$

Defining a desired second order linear reference model as: $\dot{\eta}_m = A_m\eta_m + B_m u_{ref}$

$$\begin{bmatrix} \dot{\eta}_{m_1} \\ \dot{\eta}_{m_2} \end{bmatrix} = \begin{bmatrix} -a_{m_1} & 0 \\ 0 & -a_{m_2} \end{bmatrix}\begin{bmatrix} \eta_{m_1} \\ \eta_{m_2} \end{bmatrix} + \begin{bmatrix} a_{m_3} & 0 \\ 0 & a_{m_4} \end{bmatrix}\begin{bmatrix} \varphi_s^{r2} \\ T_e^r \end{bmatrix} . \quad (16)$$

With a_m as design positive constants. At this step, the aim is to design a nonlinear SM controller that will be capable of following the above linear reference model. From Eq. (15) and (16), the error dynamic between the plant and reference model is obtained as: $e_Z = [\eta_1 - \eta_{m1}, \eta_2 - \eta_{m2}]^T = [e_{Z1}, e_{Z2}]^T$ and $\dot{e}_Z = A + B\bar{V}$. Where

$$A(\chi) = \begin{bmatrix} L_f\eta_2 \\ L_f\eta_1 \end{bmatrix}, B(\chi) = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \quad (17)$$

$$\bar{V} = \begin{bmatrix} \bar{v}_{sa} \\ \bar{v}_{sb} \end{bmatrix} = \begin{bmatrix} \hat{v}_{sa} + a_{m_1}\eta_{m_1} - a_{m_3}\varphi_s^{r2} \\ \hat{v}_{sb} + a_{m_2}\eta_{m_2} - a_{m_4}T_e^r \end{bmatrix} \quad (18)$$

Based on Equation (11), two independent SM switching functions are defined:

$$S(e_z) = \Lambda e_z(\chi) \quad (19)$$

where $\Lambda \in \Re^{2\times 2}$ is a constant linear matrix so that the inverse of $\Lambda B(\chi)$ exists for all χ. It will be proved that the following nonlinear controller is able to guarantee that the SM reaches condition,

$$\bar{V} = -(\Lambda B)^{-1}\left[\Lambda A(\chi) + Qsig(S) + KS\right] \tag{20}$$

where $sig()$ is the sign function and:

$$Q = \begin{bmatrix} q_1 & 0 \\ 0 & q_2 \end{bmatrix}, K = \begin{bmatrix} k_1 & 0 \\ 0 & k_2 \end{bmatrix}, k_2 > 0, q_2 > 0 \tag{21}$$

The switching surface dynamics are chosen as:

$$\dot{S} = \Lambda \dot{\eta}_z = \Lambda A + \Lambda B\bar{V} = -\left(Qsig(S) + KS\right) \tag{22}$$

Then,

$$\dot{S}_i = -\left(q_i sig(S_i) + k_i S_i\right)_{i=1,2} \tag{23}$$

Or, Equation (24) guarantees the SM reaches the condition.

$$\begin{aligned} S_i \dot{S}_i &= -\left(q_i S_i sig(S_i) + k_i S_i^{\,2}\right) \\ &= -\left(q_i \left|S_i\right| + k_i S_i^{\,2}\right) < 0 \end{aligned} \tag{24}$$

2.2. A New Methodology of Adaptive-Sliding Mode Control

Consider Equation (17) with some errors in the stator and rotor resistances $(\Delta R_s, \Delta R_r)$

$\dot{\eta}_z = \left[A + \Delta A\right] + B\bar{V}$ Where $\Delta A = \left[\varepsilon_1 d_1, \varepsilon_2 d_1\right]^T$ With,

$$\Delta G = \begin{bmatrix} -\left(\dfrac{\Delta R_s}{L_s\sigma} + \dfrac{\Delta R_r}{L_r\sigma}\right) i_{sa} + \dfrac{\Delta R_r}{L_s L_r \sigma}\varphi_{sa} \\ -\left(\dfrac{\Delta R_s}{L_s\sigma} + \dfrac{\Delta R_r}{L_r\sigma}\right) i_{sb} + \dfrac{\Delta R_r}{L_s L_r \sigma}\varphi_{sb} \\ \Delta R_s i_{sa} \\ \Delta R_s i_{sb} \end{bmatrix}, \Delta A = \begin{bmatrix} L_{\Delta G}\eta_2 \\ L_{\Delta G}\eta_1 \end{bmatrix} \tag{25}$$

$$\begin{aligned} L_{F_1}\eta_1(\chi) &= \left[\Delta R_s i_{sa}\right]\left[-\varphi_{sa} i_{sb} - \varphi_{sb} i_{sa}\right] \\ L_{F_2}\eta_1(\chi) &= \left[\frac{3P}{2}\left(\frac{\Delta R_s}{L_s\sigma} + \frac{\Delta R_r}{L_r\sigma}\right)\right]\left[\varphi_{sb} i_{sa} - \varphi_{sa} i_{sb}\right] \end{aligned} \tag{26}$$

Where $\varepsilon_1 = 2\Delta R_s$, and $\varepsilon_2 = \dfrac{3P}{2}\left(\dfrac{\Delta R_S}{L_S\sigma} + \dfrac{\Delta R_r}{L_r\sigma}\right)$

Since R_s, R_r slowly vary with temperature, therefore, one can assume that $\left|\varepsilon_i\right|$ are unknown bounded uncertainties. Candidate the following Lyapunov function.

$$V = \frac{1}{2}\left\{e_{\eta_1}^2 + e_{\eta_2}^2 + \frac{1}{\mu_1}\tilde{\varepsilon}_{\eta_1}^2 + \frac{1}{\mu_2}\tilde{\varepsilon}_{\eta_2}^2\right\} \tag{27}$$

where $\tilde{\varepsilon} = \hat{\varepsilon} - \varepsilon$ and $\hat{\varepsilon}$ is an estimate of ε_i and $\mu_1, \mu_2 > 0$ are adaptive input–output controller gains. Taking the derivative of V with respect to time (t) and then using Eq. (27), yields

$$\begin{aligned} \dot{V} &= -k_1 e_{\eta_1}^2 - k_2 e_{\eta_2}^2 \\ &= \tilde{\varepsilon}_1 \left\{ -e_{\eta_1} d_1 + \frac{1}{\mu_1} \dot{\hat{\varepsilon}}_{\eta_1} \right\} + \tilde{\varepsilon}_2 \left\{ -e_{\eta_2} d_2 + \frac{1}{\mu_2} \dot{\hat{\varepsilon}}_{\eta_2} \right\} + e_{\eta_1} \left\{ L_G h_2 + \bar{v}_{sa} + \hat{\varepsilon}_1 d_1 + k_1 e_{\eta_1} \right\} + e_{\eta_2} \left\{ L_G h_1 + \bar{v}_{sb} + \hat{\varepsilon}_2 d_2 + k_2 e_{\eta_2} \right\} \end{aligned} \quad (28)$$

From Equation (27), if the estimation laws and inputs are chosen as;

$$\begin{cases} \varepsilon_1 = \mu_1 e_{\eta_1} d_1 \\ \dot{\hat{\varepsilon}}_2 = \mu_2 e_{\eta 2} d_2 \\ \bar{v}_{sa} = -L_G h_2 - \hat{\varepsilon}_1 d_1 - k_1 e_{\eta_1} - p_1 \operatorname{sgn}(e_{\eta_1}) \\ \bar{v}_{sa} = -L_G h_1 - \hat{\varepsilon}_2 d_2 - k_2 e_{\eta_2} - p_2 \operatorname{sgn}(e_{\eta 2}) \end{cases} \quad (29)$$

Then, Equation (28) is reduced to:

$$\dot{V} = -k_1 e_{\eta_1}^2 - k_2 e_{\eta_2}^2 \leq 0 \quad (30)$$

Equation (31) guarantees that the state errors ez1, ez2 asymptotically converge to zero if the design parameters k1, k2, are chosen to be positive constants.

$$\Pi(t) = k_1 e_{\eta_1}^2 + k_2 e_{\eta_2}^2 \geq 0 \quad (31)$$

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3. Simulation Results

After concluding the training process the simulations were conducted using the Simulink-Matlab. The control system block diagram is illustrated in Figure 2. The descriptions of the blocks present in this diagram are:

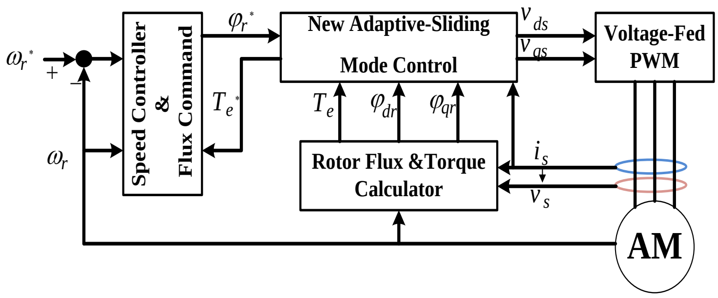

Figure 2. The Overall Simulated Block Diagram of the Proposed New Adaptive Sliding Mode Control

Motor parameters. : RS=40; RR=15.74; LS=0.103;LR=0.103; LM=0.814. It is considering, in this section that the machine parameters can change under various perturbations during the operation, this is can drive a change in Lr, Ls and Rr.

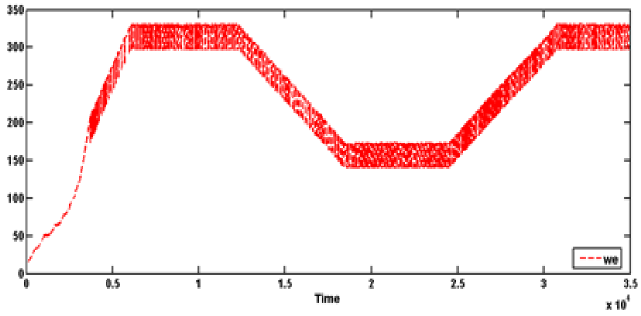

Figure 2. Speed Rotor PI Controller

Or that, one considers, in the simulation, a change inductances and in the rotor resistance. The simulations results are shown in Figure 4 to Figure 8. Comparatively with the results of the Figure 3, the Figures widely improved whereas they are superposed with the reference. One observes almost no oscillations originate of the commutation system particularly for the change of inductances.

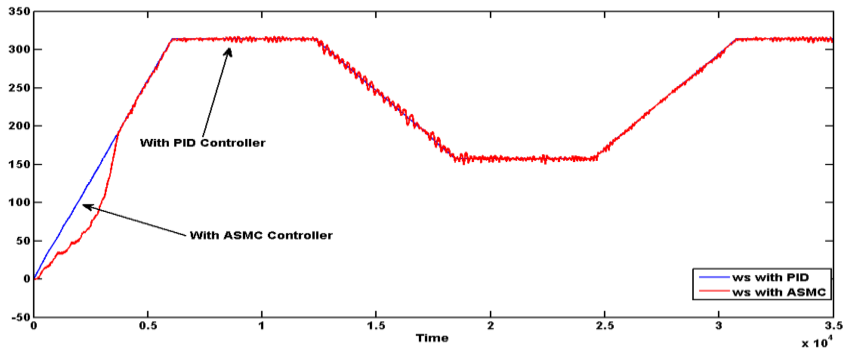

Figure 3. Speed Rotor PI Controller (Red) and ASMC Controller (Blue)

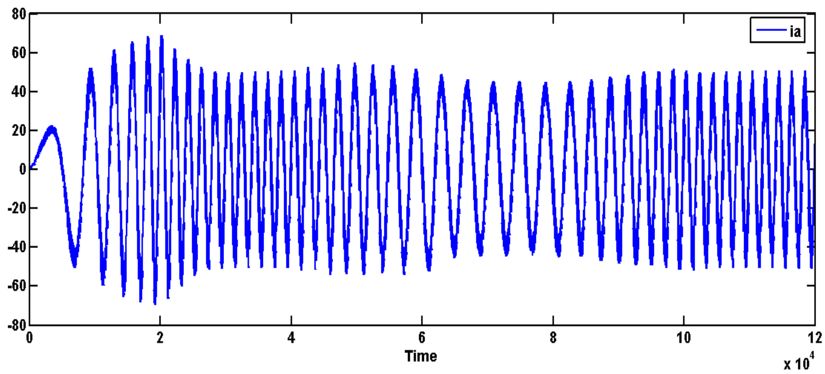

Figure 4. Current Waveforms of Phase-a

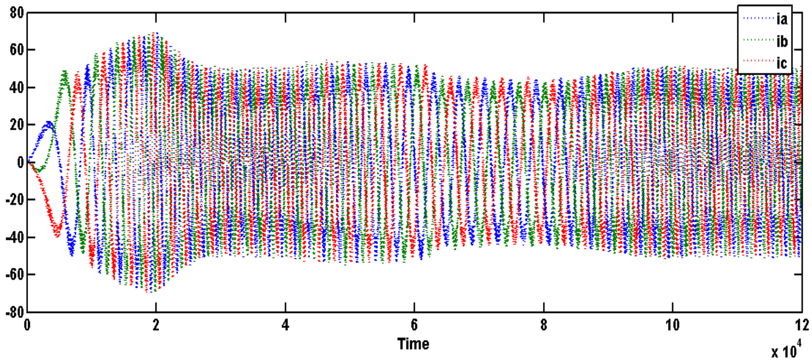

Figure 5. Current Waveforms of Three-Phase

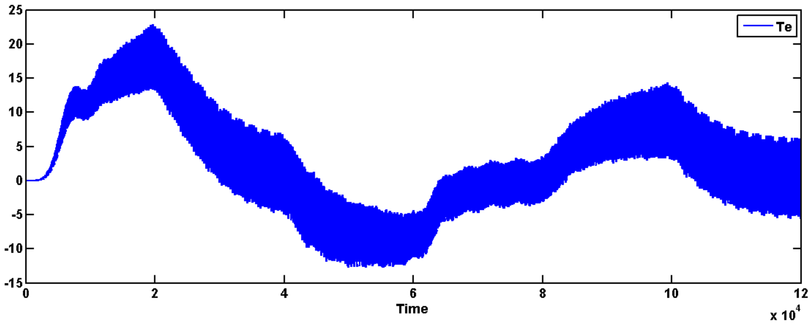

Figure 6. Torque Response with Load Torque Variation

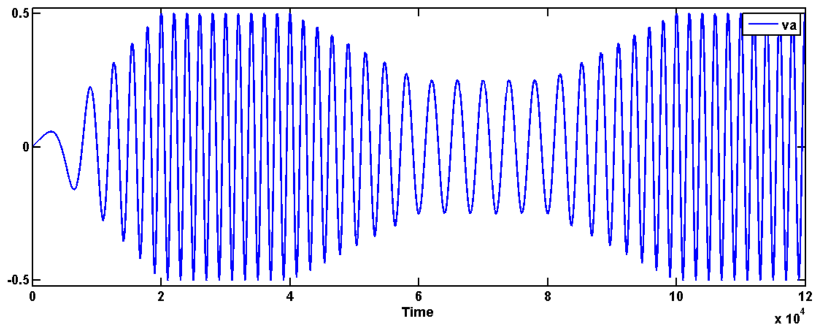

Figure 7. The Phase Voltage

However, change for the rotor resistor shows some oscillations particularly for the reactive power as shown in Figure 4 but in spite of all the result are good comparatively with the PI controller. The simulation results obtained show the robustness of a new methodology of ASMC controller address changes in machine parameters compared with PI controller.

4. Conclusion

This paper contributes to the presentation of a new methodology of adaptive-sliding mode controller asynchronous motor drive. The new ASMC controller reduces the steady state error as compared with PI-type. The Simulation results confirm that the presented controller for an asynchronous motor drive provides fast tracking capability, robust to load disturbance, and less steady state error, in very wide speed range. The resulting controller proved to have good characteristics in different operating conditions as was verified in the presented results, were performed to justify the approach to the control system synthesis.

References

[1] M Eftekhari, *et al*. *Review of induction motor testing and monitoring methods for inter-turn stator winding faults*. Electrical Engineering (ICEE), 2013 21st Iranian Conference on. 2013.
[2] DG Dorrell, *et al*. Analysis and design techniques applied to hybrid vehicle drive machines—Assessment of alternative IPM and induction motor topologies. *Industrial Electronics, IEEE Transactions on*. 2012; 59(10): 3690-3699.
[3] AVR Teja, *et al*. A new model reference adaptive controller for four quadrant vector controlled induction motor drives. *Industrial Electronics, IEEE Transactions on*. 2012; 59(10): 3757-3767.
[4] L Peretti, M Zigliotto. Automatic procedure for induction motor parameter estimation at standstill. *Electric Power Applications, IET*. 2012; 6(4): 214-224.
[5] V Verma, *et al*. Speed sensorless vector controlled induction motor drive using single current sensor. *Energy Conversion, IEEE Transactions on*. 2013; 28(4): 938-950.
[6] C Chakraborty, V Verma. Speed and Current Sensor Fault Detection and Isolation Technique for Induction Motor Drive Using Axes Transformation. 2015.
[7] M Bodson, *et al*. High-performance nonlinear feedback control of a permanent magnet stepper motor. *Control Systems Technology, IEEE Transactions on*. 1993; 1(1): 5-14.
[8] CJ Yang, *et al*. Magnetic-Field Modulation Principle and Transmission Ratio Calculation of Field-Modulated Asynchronous Magnetic Coupling. *Applied Mechanics and Materials*. 2014; 535: 48-52.
[9] R Marino, S Peresada, P Valigi. Adaptive input-output linearizing control of induction motors. *Automatic Control, IEEE Transactions on*. 1993; 38(2): 208-221.
[10] R Yazdanpanah, J Soltani, GRA Markadeh. Nonlinear torque and stator flux controller for induction motor drive based on adaptive input–output feedback linearization and sliding mode control. *Energy Conversion and management*. 2008; 49(4): 541-550.
[11] J Chiasson. A new approach to dynamic feedback linearization control of an induction motor. *Automatic Control, IEEE Transactions on*. 1998; 43(3): 391-397.
[12] R Marino, S Peresada, P Valigi. *Adaptive partial feedback linearization of induction motors*. Decision and Control, Proceedings of the 29th IEEE Conference on. IEEE. 1990.
[13] C Lascu, I Boldea, F Blaabjerg. Direct torque control of sensorless induction motor drives: a sliding-mode approach. *Industry Applications, IEEE Transactions on*. 2004; 40(2): 582-590.
[14] KK Shyu, HJ Shieh. A new switching surface sliding-mode speed control for induction motor drive systems. *Power Electronics, IEEE Transactions on*. 1996; 11(4): 660-667.
[15] M Comanescu, L Xu. Sliding-mode MRAS speed estimators for sensorless vector control of induction machine. *Industrial Electronics, IEEE Transactions on*. 2006; 53(1): 146-153.
[16] S Sastry. Nonlinear systems: analysis, stability, and control. New York. Springer. 1999; 10.
[17] Y Niu, Daniel WCH, J Lam. Robust integral sliding mode control for uncertain stochastic systems with time-varying delay. *Automatica*. 2005; 41(5): 873-880.
[18] Koshkouei, A Jafari, AS Zinober. *Sliding mode time-delay systems*. Variable Structure Systems, 1996. VSS'96. Proceedings. IEEE International Workshop on.1996.
[19] Y Niu, Daniel WCH, X Wang. Sliding mode control for Itô stochastic systems with Markovian switching. *Automatica*. 2007; 43(10): 1784-1790.
[20] L Wu, Daniel WCH. Sliding mode control of singular stochastic hybrid systems. *Automatica*. 2010; 46(4): 779-783.
[21] M Liu, Daniel WCH, Y Niu. Stabilization of Markovian jump linear system over networks with random communication delay. *Automatica*. 2009; 45(2): 416-421.
[22] Y Zhao, TA Lipo. Space vector PWM control of dual three-phase induction machine using vector space decomposition. *Industry Applications, IEEE Transactions on*. 1995; 31(5): 1100-1109.