Adaptive Sliding Mode Control with Chattering Elimination for Buck Converter Driven DC Motor

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Abstract: The Adaptive Sliding Mode Control (ASMC) that combines a robust proportional derivative control law for use in Buck converter driven DC motor is presented in this paper. Based on the LYAPUNOV theory, the proportional derivative control law is designed to eliminate the chattering action of the control signal. The simplicity of the proposed scheme facilitates its implementation and the overall control scheme guarantees the global asymptotic stability in the LYAPUNOV sense if all the signals involved are uniformly bounded. Simulation studies have shown that the proposed controller shows superior tracking performance.

Key–Words: DC-DC Buck Converter, DC Motor, Sliding Mode Control (SMC), Adaptive Sliding Mode Control (ASMC), Proportional Derivative PD.

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1 Introduction

Since its introduction Sliding Mode Control (SMC) [1, 2, 3] has become one of the most popular approach to control of nonlinear systems. The main reason is its high robustness and easy design and implementation that have resulted in a big number of applications [4, 5, 6, 7, 8, 9]. The fundamental idea of SMC consists in transferring a nonlinear system to a state from which it can be easily driven to the equilibrium. All those states including the equilibrium are described by a firs order linear differential equation and create a sliding surface. Hence SMC consists from two phases firs one when the system approaches the sliding surface from the initial state (approaching phase) and the second one when the system is sliding along the sliding surface to the fina state (sliding phase). Unfortunately, due to the presence of model imprecision and disturbances, the control law has to be discontinuous across the sliding surface. Since the associated control switchings represented by a signum function in control law are imperfect undesirable chattering of control signal arises.

Theoretical research concerning SMC has been focused mainly on over coming two crucial shortcomings: the chattering of control action and unknown behavior during the approaching phase. The chattering of control action is usually reduced by introduction of a boundary layer where the signum function is approximated by the saturation function. Unfortunately the boundary layer deteriorates the tracking performance and the robustness against disturbances. The problem of precisely unknown dynamics and reduced robustness during the approaching phase which becomes more serious if the initial condition is far from the target sliding surface is typically eliminated by a time-varying sliding surface. One possibility is to move the sliding surface such that the current state of the system is not far from it.

Shifting and rotating sliding surface for second order systems was proposed in [10] with a modifica tion in [11] and an application on position control of a direct current (DC) motor described in [12]. The procedure was generalized for n^{th} order systems in [13]. Unfortunately, the sliding surface may become unstable for some periods during which the performance is decreased. In [14] an optimal switching sliding surface for hard disk drives control is under investigation. A time-varying sliding surface scheduled with respect to the reference signal is proposed in [15]. Other control techniques proposed in the literature include passivity based control by [16], fuzzy logic based control by [17], neural network (NN) based control by [18], combined neuro-fuzzy control by [19], hierarchical control by [20], $\mathcal{H}_\infty$ based control by [21], active disturbance rejection and flatness-base control by [22] and backstepping control by [23].

Furthermore, many Adaptive Sliding Mode (ASMC) techniques have been used to reduce the chattering phenomenon (see [24, 25, 26, 27, 28, 29, 30, 31]). In fact, the controller conception does not need complete information about the uncertainty and perturbation bounds due to the dynamic gains adaptation. These gains increase automatically resulting in dangerous oscillations because of a too large switching control.

In this paper a switching sliding surface using rotation and shift for SMC of Buck converter driven DC motor is proposed. Secondly, the dynamic model of the Buck converter with motor is presented. In Section 3 the basic principles of SMC are summed up. In Section 4 the main idea of the paper consisting in improving SMC by shifting and rotating the sliding surface and the proposed are introduced. The concluding remarks are summarized in Section 5.

2 Dynamic Model of the Buck Converter with Motor

The cascaded combination of buck type dc-dc power converter and permanent magnet dc motor (cf. [20]) is shown in figur 1. It consists of a DC input voltage source (E), a controlled switch (S_w), a diode (D), a filte inductor (L), filte capacitor (C), and a DC motor (M). the mathematical model (see.[22]) of the above system can be described as:

$$\begin{cases} L\frac{di}{dt} &= -V_C + E\mu \\ C\frac{dV}{dt} &= i_L - i_M \\ L_M\frac{di_M}{dt} &= V_C - R_M i_M - K_e\omega \\ J\frac{d\omega}{dt} &= K_M i_M - f_v\omega - T_l \end{cases} \tag{1}$$

Where i is the converter input current, V_C is the converter output voltage, i_M is the DC motor armature circuit current and ω is the angular velocity of the motor shaft, which may be subject to a constant. K_M is torque constant and K_e is back electromotive force constant(EMF). but unknown, torque load T_l. The control input is represented by the variable μ which takes 1 for the ON state of the switch and 0 for the OFF state.

The capacitor current C is approximately 0, because the value of capacitor is of hundreds of micro Farads (μF) (cf.[32]), from seconde equation of the system (1); $i = i_M$. Under this condition; adding the firs an third expression of the equation system (1), we have:

$$(L + L_M)\frac{di}{dt} = E\mu - R_M i - K_e\omega \tag{2}$$

And expression four is :

$$J\frac{d\omega}{dt} = K_M i - f_v\omega - T_l \tag{3}$$

And

$$\frac{dT_l}{dt} = 0 \tag{4}$$

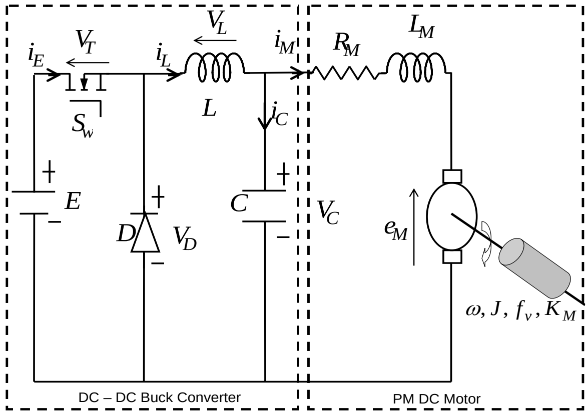

Figure 1: Cascaded buck converter DC motor combination.

We need to transform equations (2) and (3) to the triangular form. One will introduce the change of variable according to:

$$\begin{cases} x_1 &= i \\ x_2 &= -\frac{R_M}{L+L_M}i - \frac{K_e}{L+L_M}\omega + \frac{E}{L+L_M}\mu \end{cases} \tag{5}$$

Where

$$\omega = \frac{E\mu - (L + L_M)x_2 - R_M i}{K_e} \tag{6}$$

Using the transformation (5) and a time derivative of these states, we can rewrite from equations (2) and (3), a following model :

$$\begin{cases} \dot{x}_1 &= x_2 \\ \dot{x}_2 &= -\frac{1}{J(L+L_M)}\left(K_e K_M + R_M f_v\right)x_1 \\ &- \left(\frac{R_M}{L+L_M} + \frac{f_v}{J}\right)x_2 + \frac{f_v}{J(L+L_M)}E\mu \\ &+ \frac{K_e}{J(L+L_M)}T_l \end{cases} \tag{7}$$

Model (7) can be rewritten under the following compact form:

$$\begin{cases} \dot{x}_1 = x_2, \\ \dot{x}_2 = f(x) + g(x)\mu + d, \\ y = x_1. \end{cases} \quad (8)$$

where

$$f(x) = -\frac{1}{J(L+L_M)}(K_eK_M + R_Mf_v)x_1 - \left(\frac{R_M}{L+L_M} + \frac{f_v}{J}\right)x_2$$

$$g(x) = \frac{f_vE}{J(L+L_M)} \;;\; d = \frac{K_e}{J(L+L_M)}T_l.$$

And

$$x = \begin{bmatrix} x_1 & x_2 \end{bmatrix}^T$$

3 Conventional Sliding Mode Control

Consider a general class of nonlinear systems in the form (8); where x is the state vector of the system which is assumed to be available for measurement, E and y is the input and output of the system, respectively, d is unmeasurable bounded external disturbance to have upper bound D, that is $|d| \leq D < \infty$. The nonlinear system (8) is controllable and the input gain $g(x) \neq 0$. Without loss of generality, we assume that $g(x) > 0$.

To generate a direct relation between the output y and the input E, we derive the output y:

$$\dot{y} = \dot{x}_1 = x_2$$

Since $\dot{y}$ is not directly linked to the input E, we must derive another time and we obtain:

$$\begin{aligned} \ddot{y} &= \dot{x}_2 \\ &= -\frac{1}{J(L+L_M)}(K_eK_M + R_Mf_v)x_1 \\ &- \left(\frac{R_M}{L+L_M} + \frac{f_v}{J}\right)x_2 + \frac{f_v}{J(L+L_M)}E\mu \\ &+ \frac{K_e}{J(L+L_M)}T_l \\ &= f(x) + g(x)\mu + d. \end{aligned}$$

The relative degree of the system output is $r = 2$.

The control objective is to design a control law for the state x to track a desired reference state trajectory $x_d = x_{1d}$ in the presence of model uncertainties and external disturbances. The tracking error is given by the following relationship:

$$e = x_1 - x_{1d} \quad (9)$$

Consider the sliding surface σ, define by.

$$\sigma = \dot{e} + k_0 e \quad (10)$$

Where $K = \begin{bmatrix} k_0 & 1 \end{bmatrix}^T$ is the vector of the coefficient of a HURWITZ polynomial $K(p) = p + k_0$ where p is the LAPLACE operator. If it is equal to zero the state of the system is on the sliding surface. For the zero initial conditione $e(0) = 0$, the tracking problem $x = x_d$ can be considered as keeping the error state vector on the sliding surface $\sigma = 0$ for all time. A sufficien condition to guarantee that the trajectory of the error vector e will approach to the sliding surface in finit time is to choose the control strategy such that:

$$\frac{1}{2}\frac{d\sigma^2}{dt} \leq -\eta\,|\sigma|\,. \quad (11)$$

with η is positive constant.

The system is controlled in such a way that it always moves toward the sliding surface and hits it. The sliding process includes two phases:

- The firs one is the approaching phase $\sigma \neq 0$
- The second one is the sliding phase $\sigma = 0$.

The sliding mode control guarantees the convergence of the nominal system to the equilibrium point, i.e.$\lim_{t\mapsto\infty}(e(t)) = 0$.

Consider the control problem of the nonlinear system (8). The Sliding Mode Control (SMC) control law (12) satisfie the sliding condition (11):

$$\begin{aligned} U^* &= g^{-1}(x)\left(-f(x) + \dot{x}_d - k_0 e - \eta\,\text{sign}(\sigma)\right) \quad (12) \\ &= \frac{J(L+L_M)}{f_vE}\left(\frac{1}{J(L+L_M)}(K_eK_M + R_Mf_v)x_1 - \left(\frac{R_M}{L+L_M} + \frac{f_v}{J}\right)x_2 + \dot{x}_d - k_0 e - \eta\,\text{sign}(\sigma)\right) \end{aligned}$$

where $\eta > 0$ and sign denotes the signum function; where

$$\text{sign}(\sigma) = \begin{cases} -1 & if \quad \sigma < 0; \\ 0 & if \quad \sigma = 0; \\ 1 & if \quad \sigma > 0. \end{cases}$$

Let the LYAPUNOV function candidate define as

$$V(e) = \frac{1}{2}\sigma^2 \tag{13}$$

Differentiating (13) with respect to time, $\dot{V}$ along the system trajectory as

$$\begin{aligned} V(e) &= \sigma.\dot{\sigma} \\ &= \sigma.\left(\ddot{e} + k_0\dot{e}\right) \\ &= \sigma.\left(\ddot{x}_1 - \ddot{x}_{1d} + k_0\dot{e}\right) \\ &= \sigma.\left(f(x) + g(x)\mu + d - \ddot{x}_{1d} + k_0\dot{e}\right) \\ &\leq -\eta\left|\sigma\right| \end{aligned} \tag{14}$$

Hence the sliding mode control input U^* guarantees the sliding condition of equation (11). It can be noted that in order to satisfy the sliding condition, a hitting control term U_{sw} is added in the overall control action. i.e. $U^* = U_{eq} + U_{sw}$. Where

$$\begin{aligned} U_{eq} &= g^{-1}(x)\left(-f(x) + \dot{x}_d - k_0 e\right) \qquad (15) \\ &= \frac{J(L+L_M)}{f_v E}\left(\frac{1}{J(L+L_M)}\left(K_e K_M \right.\right. \\ &\quad \left. +R_M f_v\right) x_1 - \left(\frac{R_M}{L+L_M} + \frac{f_v}{J}\right) x_2 + \dot{x}_d \\ &\quad -k_0 e) \\ U_{sw} &= -g^{-1}(x).\eta\text{sign}(\sigma) \qquad (16) \\ &= -\frac{J(L+L_M)}{f_v E}.\eta\text{sign}(\sigma) \end{aligned}$$

The block diagram of the conventional sliding mode control is presented in figur 2. Besides, it is also to be made clear the control law U^* given in equation (12) is a continuous time domain signal which is further discretized using Pulse Width Modulation (PWM) method with an appropriate section of carrier wave signal. Therefore the switching signal is ultimately given to the power semiconductor switch S_w.

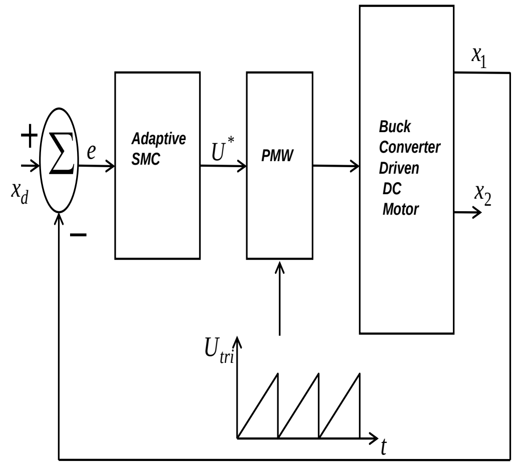

Figure 2: Block diagram of the conventional sliding mode control.

The SMC is characterized by its precision and robustness to parametric variations and external disturbances. Yet, the main disadvantage of this kind of control lies in the difficult of choosing the sliding surface parameters, ie the robustness of the system during the reaching phase and the chattering phenomenon.

In what follows, the ASMC where the surface is moving was introduced. This control technique is based on the choice of a variant time linear sliding surface that conveniently adapts to arbitrary initial conditions and allows to obtain better tracking performances. To attenuate the chattering phenomenon, an adaptive derivative proportional term was incorporated in the global control law.

4 Adaptive sliding mode control

4.1 Adaptive sliding mode controller design

The discontinuous term in equation (16) is replaced by an adaptive Proportional Derivative PD term.

$$U_{sw} = -g^{-1}(x).U_{PD} \tag{17}$$

This adaptive term can be written as follows:

$$\begin{aligned} U_{PD} &= k_P.\sigma + k_D.\frac{d\sigma}{dt} \\ &= \begin{bmatrix} k_P & k_D \end{bmatrix}.\begin{bmatrix} \sigma \\ \frac{d\sigma}{dt} \end{bmatrix} \\ &= \theta.\Psi(\sigma) = \hat{\rho}(\sigma, \hat{\theta}) \end{aligned} \tag{18}$$

Where $\theta = \begin{bmatrix} k_P & k_D \end{bmatrix}$, $\Psi(\sigma) = \begin{bmatrix} \sigma & \frac{d\sigma}{dt} \end{bmatrix}^T$; with k_P and k_D are the adjusted gains. The sliding moving surfaces σ define in expression (10).

The global control law is written as follows:

$$U_{ad} = g^{-1}(x)\left(-f(x) + \dot{x}_d - \hat{\zeta}e - \hat{\rho}(\sigma,\hat{\theta})\right) \quad (19)$$

The parameter vectors are computed using the following adaptation laws:

$$\begin{aligned} \dot{\hat{\zeta}} &= \gamma_\zeta \sigma.e & (20)\\ &= \gamma_\zeta\left(\dot{e} + \zeta e\right).e \\ \dot{\hat{\rho}}(\sigma,\hat{\theta}) &= \gamma_\theta \sigma.\Psi(\sigma) & (21)\\ &= \begin{bmatrix} \gamma_{P\theta} & \gamma_{D\theta} \end{bmatrix}.\left(\dot{e} + \zeta e\right).\begin{bmatrix} \sigma \\ \frac{d\sigma}{dt} \end{bmatrix} \end{aligned}$$

With $\gamma_\zeta > 0$, $\gamma_{P\theta} > 0$ and $\gamma_{D\theta} > 0$ are the adaption gains.

The adaptive control schema (see Figure. 3) is described by the following block diagram:

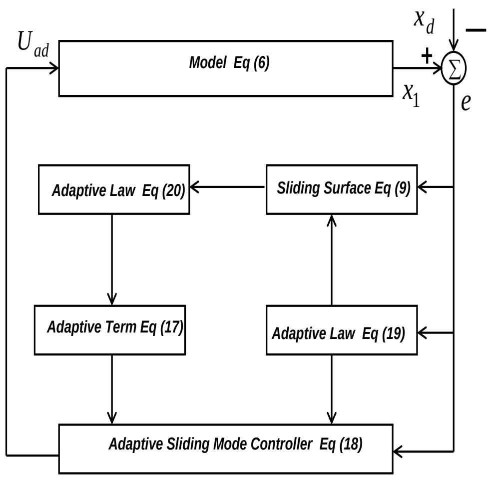

Figure 3: Block diagram of adaptive sliding mode control with moving surface.

4.2 Stability analysis

Defin the optimal parameters vector:

$$\theta^* = \arg\min_{\theta\in w}\left(\sup_{\sigma\in\mathbb{R}}\left\|\eta\text{sign}(\sigma) - \hat{\rho}(\sigma,\hat{\theta})\right\|\right) \quad (22)$$

Where w are constraint sets for θ. Defin the minimum approximation error.

$$\Omega = \left\{\theta \in \mathbb{R}^2 : \|\theta\| < M_\theta\right\} \quad (23)$$

where M_θ is a predefine parameter.

w is the minimum of the approximation errors define as:

$$w = \eta\text{sign}(\sigma) - \hat{\rho}(\sigma,\hat{\theta}) \quad (24)$$

Differentiating the moving sliding surface vector σ with respect to time, given by equation (10), and using the control law E in equation (19), the time derivative of the vector σ can be described as follows:

$$\begin{aligned} \dot{\sigma} &= -\hat{\zeta}e - \hat{\rho}(\sigma,\hat{\theta})) + \zeta e + d \\ &= \tilde{\zeta}e - \hat{\theta}.\Psi(\sigma) + d + w - \eta\text{sign}(\sigma) + \theta^*.\Psi(\sigma) \\ &= \tilde{\zeta}e + \tilde{\theta}.\Psi(\sigma) + w - \eta\text{sign}(\sigma) + d & (25) \end{aligned}$$

where $\tilde{\zeta} = \zeta - \hat{\zeta}$ and $\tilde{\theta} = \theta^* - \hat{\theta}$.

The LYAPUNOV function candidate is chosen as follows:

$$V = \frac{1}{2}\sigma^2 + \frac{1}{2\gamma_\theta}\tilde{\theta}^T\tilde{\theta} + \frac{1}{2\gamma_\zeta}\tilde{\zeta}^2 \quad (26)$$

The time derivative of V along the error trajectory (26) is:

$$\dot{V} = \sigma\dot{\sigma} + \frac{1}{\gamma_\theta}\tilde{\theta}\dot{\tilde{\theta}} + \frac{1}{\gamma_\zeta}\tilde{\zeta}\dot{\tilde{\zeta}} \quad (27)$$

Replacing the time derivative of σ given by expression (25) in equation (27), we get:

$$\begin{aligned} \dot{V} &= \sigma\left(\tilde{\zeta}e + \tilde{\theta}.\Psi(\sigma) + w - \eta\text{sign}(\sigma) + d\right) \\ &\quad + \frac{1}{\gamma_\theta}\tilde{\theta}\dot{\tilde{\theta}} + \frac{1}{\gamma_\zeta}\tilde{\zeta}\dot{\tilde{\zeta}} \\ &= \sigma\tilde{\zeta}e + \sigma\tilde{\theta}.\Psi(\sigma) + \sigma(w + d) - \sigma\eta\text{sign}(\sigma) \\ &\quad + \frac{1}{\gamma_\theta}\tilde{\theta}\dot{\tilde{\theta}} + \frac{1}{\gamma_\zeta}\tilde{\zeta}\dot{\tilde{\zeta}} & (28) \end{aligned}$$

We have $\dot{\tilde{\zeta}} = -\dot{\hat{\zeta}}$ and $\dot{\tilde{\theta}} = -\dot{\hat{\theta}}$. Hence, we can rewrite (28) as:

$$\begin{aligned} \dot{V} &= \tilde{\zeta}\left(\sigma e - \frac{1}{\gamma_\zeta}\dot{\hat{\zeta}}\right) + \tilde{\theta}\left(\sigma.\Psi(\sigma) - \frac{1}{\gamma_\theta}\dot{\hat{\theta}}\right) \\ &\quad + \sigma(w + d) - \sigma\eta\text{sign}(\sigma) & (29) \end{aligned}$$

By replacing (20) and (21) in (29), we obtain:

$$\begin{aligned} \dot{V} &= \sigma(w + d) - \sigma\eta\text{sign}(\sigma) \\ &\leq |\sigma|(w + d) - |\sigma|\eta\text{sign}(\sigma) \\ &\leq |\sigma|(w + d - \eta\text{sign}(\sigma)) & (30) \end{aligned}$$

We can conclude that:

$$\dot{V} \leq 0 \;\; if \;\; \forall\eta \geq w + d. \quad (31)$$

Because $V \leq 0$ is negative semi-definit i.e. $V(t) \leq V(0)$ implying that $\sigma; \tilde{\theta}; \tilde{\zeta}$ are bounded. From (25), it can be concluded that $\dot{\sigma}$ is bounded.

Integrating equation (30) from zero to t, it yields:

$$V(0) \geq V(t) + \eta \int_0^\infty |\sigma(\tau)|\, d\tau \qquad (32)$$

Since $V(0)$ is bounded and $V(t)$ is non increasing and bounded, it can be concluded that $\lim_{t\mapsto\infty} \int_0^\infty |\sigma(\tau)|\, d\tau$ is bounded. Then, $\lim_{t\mapsto\infty} \int_0^\infty |\sigma(\tau)|\, d\tau$ is bounded and $\dot{\sigma}$ is also bounded, based on BARBALAT's lemma, $\sigma(t)$ will converges asymptotically to zero and $\lim_{t\mapsto\infty} \int_0^\infty \sigma(t) = 0$. Then, e converges to zero asymptotically.

The stability result is verifie if all parameters involved in equation (27) are bounded. To ensure the boundedness of this parameters, the adaptive laws (20) and (21) can be modifie using the projection algorithm in [30, 31]. The modifie adaptive laws are given as follows.

System parameters	Rating
Power converter rating P	$120W$
Power switch MOSFET S_w	$IRFP460(500V/18A)$
Power diode D	$6A4$ MIC
Switching frequency f_s	$20kHz$
Supply DC voltage E	$26V$
PMDC motor rating	$\frac{1}{3}HP, 24V, 1500rpm$
Nominal load torque T_L	$0.012Nm$
Armature resistance R_M	2.5Ω
Armature inductance L_M	$176mH$
Viscous friction coefficien f_v	$2.401e^{-6}Nms$
Moment of inertia J	$3.9108e^{-6}kgm^2$
Buck inductor L	$20mH, 10A$
DC capacitor C	$220\mu F, 450V$
Back EMF constant K_e	$0.0145Vs/rad$
Torque constant K_M	$0.0105Nm/A$
Nominal reference speed ω_r	$500rpm$

Table 1: Specification of cascaded buck dcdc converter DC-motor combination.

For $\hat{\theta}$, we use:

$$\dot{\hat{\theta}} = \begin{cases} \gamma_\theta \sigma.\Psi(\sigma) & \text{if } \left(\left\|\hat{\theta}\right\| < M_\theta\right) \text{ or } \left(\left\|\hat{\theta}\right\| = M_\theta \text{ if } \sigma\hat{\theta}\Psi(\sigma) \geq 0\right) \\ P_\theta\left\{\gamma_\theta \sigma.\Psi(\sigma)\right\} & \text{if } \left(\left\|\hat{\theta}\right\| = M_\theta\right) \text{ and } \left(\sigma\hat{\theta}\Psi(\sigma) > 0\right) \end{cases} \qquad (33)$$

For $\hat{\zeta}$, we use:

$$\dot{\hat{\zeta}} = \begin{cases} \gamma_\zeta \sigma.e & \text{if } \left(\left\|\hat{\zeta}\right\| < M_\zeta\right) \text{ or } \left(\left\|\hat{\zeta}\right\| = M_\zeta \text{ if } \gamma_\zeta \sigma.e \geq 0\right) \\ P_\zeta\left\{\gamma_\zeta \sigma.e\right\} & \text{if } \left(\left\|\hat{\zeta}\right\| = M_\zeta\right) \text{ and } (\gamma_\zeta \sigma.e > 0) \end{cases} \qquad (34)$$

where M_θ and M_ζ are the design parameters that specify the allowable bounds of $\hat{\theta}$ and $\hat{\zeta}$.

The projection operator $P_\theta\{*\}$ and $P_\zeta\{*\}$ are define as:

$$P_\theta\left\{\gamma_\theta \sigma.\Psi(\sigma)\right\} = \gamma_\theta \sigma.\Psi(\sigma) - \gamma_\theta \sigma.\frac{\theta\theta^T\Psi(\sigma)}{\|\theta\|^2} \qquad (35)$$

$$P_\zeta\left\{\gamma_\zeta \sigma.e\right\} = \gamma_\zeta \sigma.e - \gamma_\zeta \sigma.\frac{\zeta^2 e}{\|\zeta\|^2} \qquad (36)$$

5 Simulation and Results

The parameters associated with the proposed controller are given in table 1.

The stabilizing gains of the controller are chosen suitably to obtain a satisfactory response. The gains selected are $k_0 = 2$ and $\eta = 477$; where the adaptive gain initialize are $\gamma_\zeta = 6$, $\gamma_{P\theta} = 50$ and $\gamma_{D\theta} = 100$. The system of buck converter fed PMDC-motor is studied for the following case studies:

Case study I: A step change in T_L from $0.025Nm$ to $0.1575Nm$.

Case study II: A step change in i from $0.392A$ to $1.512A$.

Case study III: A step change in i from $1.512A$ to $0.392A$.

The results obtained by using the proposed controller are compared against the results of conventional adaptive control technique. An exact knowledge of the time varying load torque is essential for effective tracking and control of angular velocity of the dc motor. The figure 4 and 5 shows the results containing current armature i and angular velocity ω for sudden load torque variation from nominal value of $0.025Nm$ to $0.1575Nm$ for both the conventional SMC scheme and the proposed ASMC scheme. The conventional SMC scheme results pick $1.046A$ in start with chattering, It is not for ASMC, but speed grits are similar. We notice, in the other cases change of the current, the two commands fl w exactly the desired trajectory.

The armature current error and the histogram with GAUSSIAN distribution for SMC are shown by figur 6 when the error means is equal $26 \times 10^{-4}A$ and the variance is 2×10^{-3}. But for ASMC are shown by figur 7. The error means is equal $18 \times 10^{-4}A$ and the variance is 1.1×10^{-3}.

The evolution of the correction gain ζ and ρ are given by the figure 8 and 9.

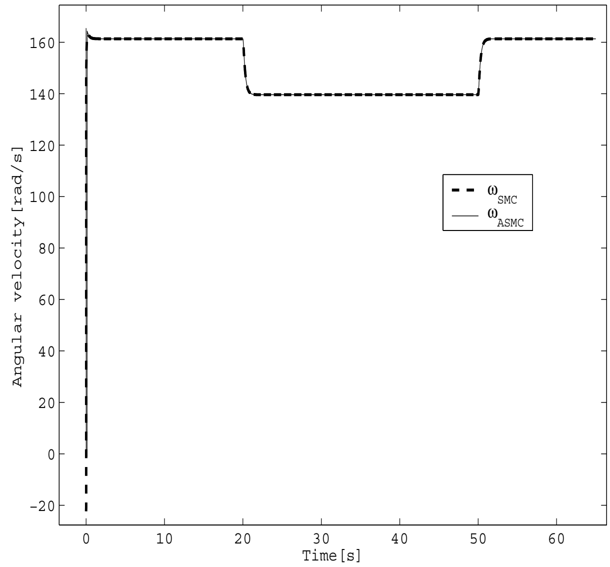

Figure 5: Angular velocity.

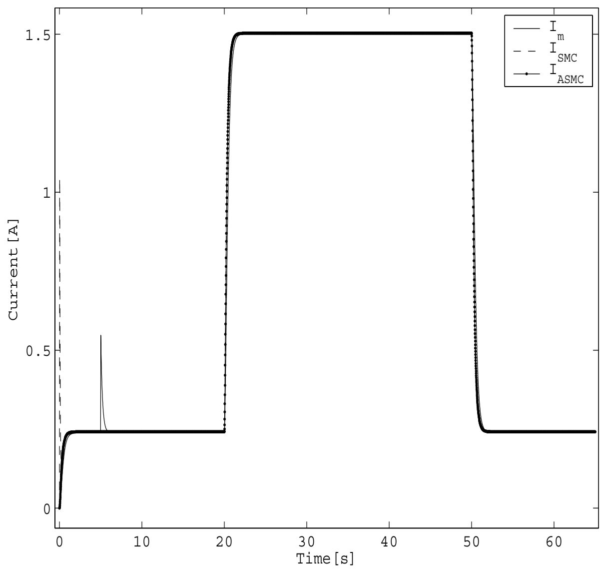

Figure 4: Armature current.

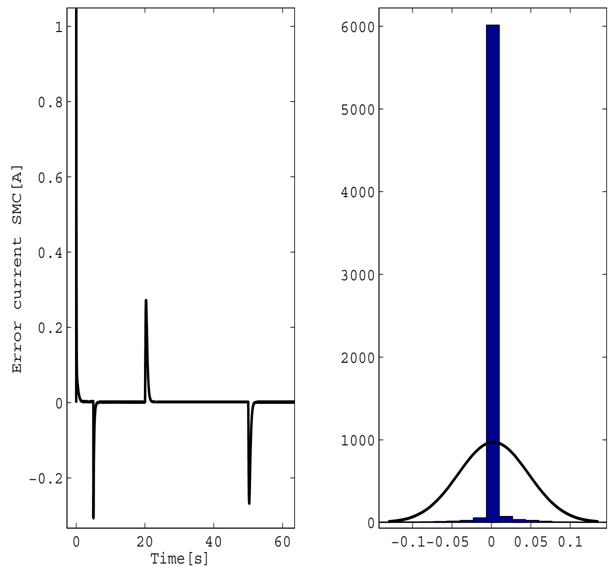

Figure 6: Armature current error with SMC.

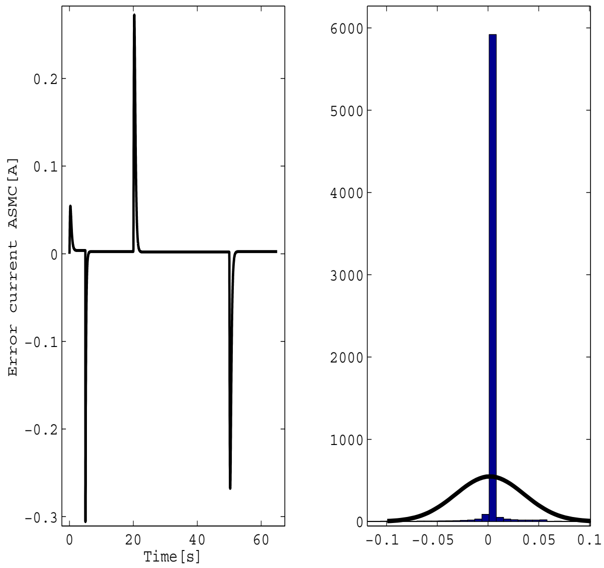

Figure 7: Armature current error with ASMC.

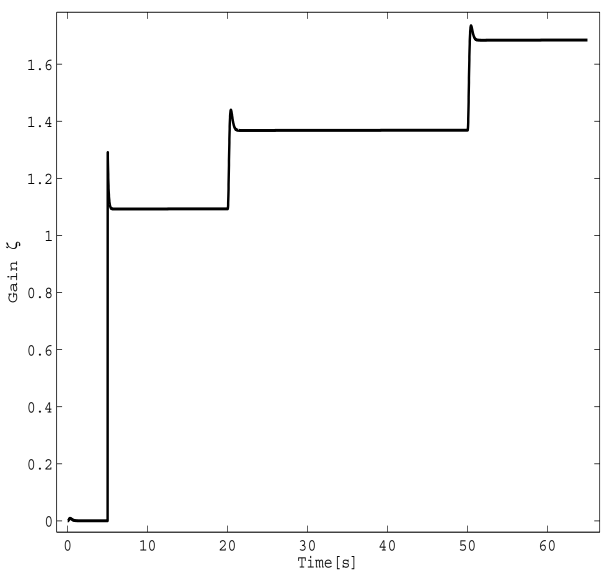

Figure 8: Gain ζ.

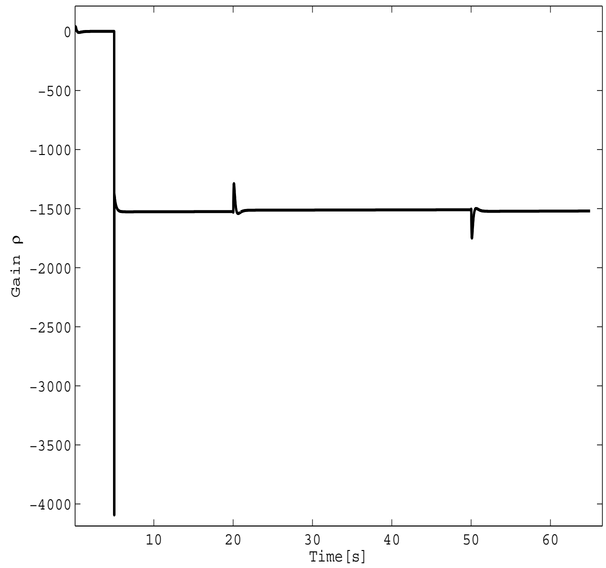

Figure 9: Gain ρ.

6 Conclusions

In this study, an adaptive sliding mode control was proposed for a Buck converter driven DC motor was presented. The proposed switching strategy consisting of rotation and shift guarantees stability of sliding surface at any time. In comparison with classical sliding mode control the presented control improves tracking performance of the controlled system during the approaching phase, especially in the presence of a torque load.

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