# Implementation of a Fast Terminal Sliding Mode Controller for Direct Thrust Control Systems

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***Abstract*—This paper presents a new methodology to implement a permanent magnet linear synchronous motor. The proposed method is a combination way which is combined the fast terminal sliding mode with the direct thrust control method. The intended advantage of the proposed method is that the chattering phenomenon can be reduced and/or removed. To our knowledge, the inherent flux linkage problems of permanent magnet linear synchronous motors did not seem to be handled by the pioneer contributions. For overcoming the problems, a new estimator algorithm is used to get the accurate flux linkage value. Finally, some experimental results are met as expected.**

***Keywords*—Permanent magnet linear synchronous motor, fast terminal sliding mode, direct thrust control.**

## I. INTRODUCTION

Linear motor drives equipment directly and does not need any mechanical coupling like belts, gears, screws or crankshafts, etc. System efficiency with linear motor driving directly and its control circuits is higher than other rotating driving systems, even higher positioning precision. Therefore, permanent magnet linear synchronous motors (PMLSMs) are widely used in industry drive areas of linear transmission or precision servo control [1-3]. In recent years, scientist and technologist are trying to replace rotating motor with linear motor in numerical control machining center or more axes cooperation system because of its efficiency and controlled precision by advanced control strategies. The direct torque control has been utilized as a general formalism methodology for AC rotary motors [4-5]. It has been expanding into linear motor drives, named as direct thrust control (DTC) [6-7]. However, the technical skill in the implementation of PMLSMs does not seem to be performed yet. The reasons state both the problems of the amplitude of the actual flux linkage and the position angle of the actual flux linkage are no easy obtained. For overcoming the problems, the paper proposes a new estimator algorithm to obtain both. However, it is a well-known that PMLSM positioning accuracy is greatly affected by parameter variations and load disturbances. Furthermore, how to reduce these factors quickly and directly is a very important issue.

Sliding-mode control works on the concept of designing a control so as to drive the system states to the so called sliding manifold. One of the major advantages of sliding-mode control is that when the system states are on the sliding manifold, the closed-loop system behavior is robust to certain internal parameter variations and external disturbances [8-10]. In general, sliding manifolds are chosen to be linear hyperplane named linear sliding mode (LSM). Such hyperplane guarantees asymptotic stability of the system in the sliding mode. That is, the system states will reach the equilibrium in infinite time. Nevertheless, tracking error will no converge to zero in finite time in any case. Terminal sliding-mode (TSM) control [11] aims at designing a sliding-mode control strategy that would guarantee a finite time convergence to the origin. This is accomplished by using a nonlinear sliding line which results in finite time convergence. And also [12-14] have been used successfully. Although this method has proved successfully, there is a series lack of effective or tractable design methods to choose the suitable parameters. Comparison with LSM policy, the TSM policy does not seem to provide the same convergent performance when the system state is far away from the equilibrium. The main reason is that the convergent time of the LSM is shorter than the TSM when the system state is close the equilibrium. On the other hand, the discontinuity phase is a major contributor to the chattering behaviors.

In this paper, we propose a fast terminal sliding mode (FTSM) controller that can drive the system to reach the nonlinear switching manifold in the reaching phase. It is worthy to notice that LSM and TSM control technologies are used to improve the precision of the control and deliver a fast time response, respectively. Moreover, the chattering phenomenon can be reduced and/or removed in the reaching phase and sliding phase under our controller.

This paper is organized as follows. A modified flux linkage estimation policy is introduced in section II. A fast terminal sliding mode controller is presented in section III. The simulation results are depicted in section IV. Finally, some conclusions are given in section V.

## II. A MODIFIED FLUX LINKAGE ESTIMATION POLICY

A control policy is proposed which is insensitive of rotor parameters in this paper. And also the implementation of the control policy is completely without the complicated rotating frame transform. For convenience, a block diagram of a direct thrust control scheme for the PMLSM control system is depicted in figure 1. It is worthy to point that DTC allows a controller to control the decoupled of electromagnetic thrust and the flux magnitude, respectively. The main advantages of the DTC are as follows. 1) DTC has a simple and robust control structure; 2) DTC with a switching table provides excellent thrust dynamics. In fact, the performance of DTC strongly depends on the quality of the estimation of the actual flux and thrust. Therefore, some problems associated with the DTC of a PMLSM [15], namely, 1) drift in flux linkage estimation of flux linkage due to offset error in measurement, 2) error in the estimation of flux linkage due to variation of mover resistance. Based on the reasons, the flux estimation becomes an important task when one wants to implement a high-performance motor drive. There are two methods used for the flux estimation: one is based on measured the currents of motors, called current-based method, and the other one is based on measured voltages [16-17], called voltage-based method. In the voltage-based method, the mover flux can directly be obtained from its back *emf* via an integrator. For the mentioned above, the common method for estimating flux linkage is based on the voltage model. The voltage equation of a PMLSM in the stationary reference frame can be expressed as follow:

$$u_m = Ri_m + \frac{dF_m}{dt} \tag{1}$$

where $R$ is the mover resistance, $u_m = [u_D\ u_Q]^T$, $i_m = [i_D\ i_Q]^T$, $F_m = [F_D F_Q]^T$ is mover flux linkage vector. In a DTC scheme, the flux linkage is estimated by integrating difference between the input voltage and the voltage drop across the mover resistance given by equation (1) and can be rewritten as

$$F_m = \int (u_m - Ri_m)dt = \int e_m dt \tag{2}$$

where $e_m$ is the back *emf*. Then the flux linkage amplitude and angle can be obtained by the following equations.

$$|F_m| = \sqrt{F_D^2 + F_Q^2} \tag{3}$$

$$\theta = \tan^{-1}(F_Q / F_D) \tag{4}$$

where $|F_m|$ and $\theta$ is the flux linkage amplitude and position angle, respectively. Generally, the developed electromagnetic thrust can be described as

$$T_m = \frac{3}{2}\frac{\pi}{\tau} k_F n_p (F_D i_Q - F_Q i_D) \tag{5}$$

where $T_m$ is the output electromagnet thrust, $n_p$ is the number of pole pairs, $k_F$ is the end effect correct coefficient of linear motor thrust. In the DTC, the exact prediction of the actual thrust is very important because the control performance of PMLSM is depends on it. The end effect is a special phenomenon because the linear mover is limited in its stroke. Notice that the thrust correction coefficient of the linear motor is defined as a constant 0.9 by [6].

In the estimations of the thrust and flux, the integration of the equations (2) and (3) are the most important factors. If the motor operates in the high-speed region, the voltage should be dropped into the rotor resistance. Notice that the voltage can be neglected because $u_m >> Ri_m$. And also a value of the flux can be obtained from $u_m$ via an integrator. Moreover, the correct space voltage vector can be obtained from the switching algorithm. And it can minimize the errors of the thrust and the flux. However, the implementation of an integrator for motor flux estimation is no easy task. It is worthy to point that the DC offset in the measurements of $u_m$ and $i_m$ would eventually lead to a large drift in the estimated flux linkage. Therefore, the integrator must be reset regularly to reduce the effect of the offset error. A few compensation techniques of this offset have been reported in the literature [18-20]. However, the problem does not seem to be solved completely by the prior researches. For overcoming this drawback, there is a voltage-model-based flux estimator is proposed [20]. It is interesting to note that the modified flux linkage estimator is proposed in this paper based on [20]. The new one is with a correction coefficient ($k_\varphi$) in the flux estimate algorithm. A new modefied block diagram is shown in Fig. 2. The estimator structure consists of a low pass filter and a signal feedback block. It is worthy to notice that (5) shows the thrust can be influenced by the flux linkage. And also the estimation accuracy of the flux linkage estimated is required. Considering the estimator input signals, they are combined with a sine wave which is with a foundation frequency and its higher order harmonics. And the output signals are sine waves which are only with the foundation frequency. According to the waveform theory, the correction coefficient is determined to π/4 in this paper.

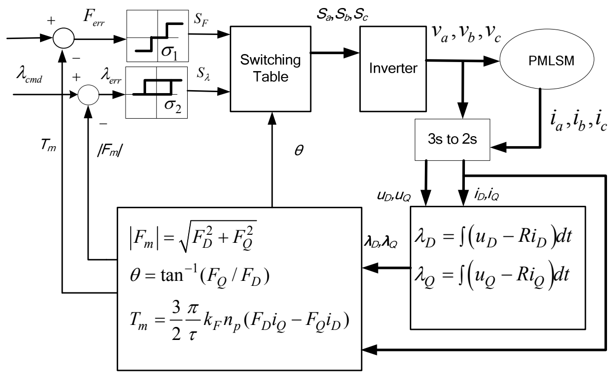


Figure 1. A PMLSM control block diagram using DTC policy.

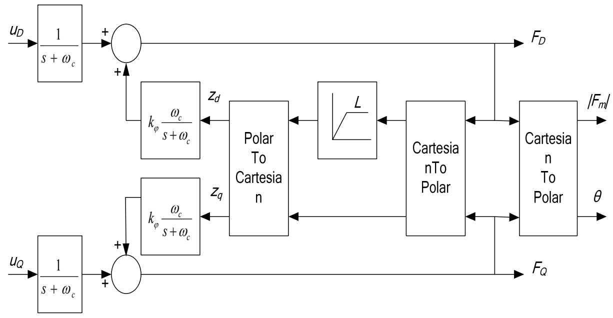


Figure 2. A voltage-base modified integrator block diagram.

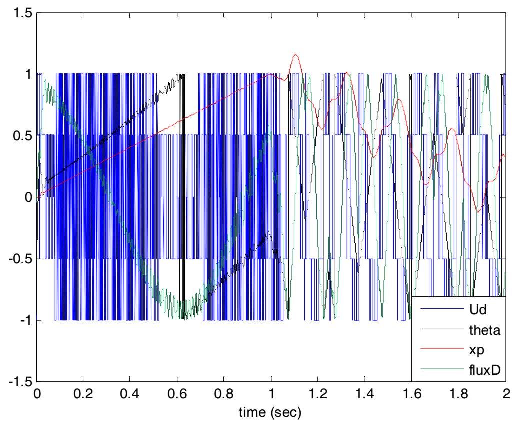


Figure 3(a). A performance of the flux estimate algorithm [20].

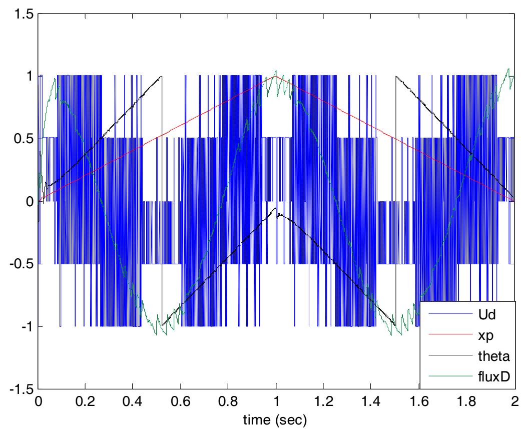


Figure 3(b). A performance of a new flux estimate algorithm.

Fig. 3(a) shows the performance of the control policy [20] is without the correction coefficient $k_\varphi$. And Fig. 3(b) is with $k_\varphi$. Notice that both the performances are under controlled by the same position command. In the two examples, a triangular reference input is required for moving 1τ (cm) with velocity 1τ/*s* (cm/sec). It is worthy to notice that the result (i.e., Fig. 3(b)) is obtained under our modified integrator. It hints the suitable voltage vectors are selected successfully from the switching table by our control algorithm. For convenience, all experimental results of the motor are normalized. For instance, the voltage vectors ($u_D$) is normalized as $U_d = u_D / 280\ V$, the angle theta = $(\theta/180)$ *degree* and the flux linkage *fluxD* = $F_D / 0.3 Wb$. Comparing the two performances (i.e., Fig. 3(a) and Fig. 3(b)), it is easy to point out that the new position angle $\theta$ is more linearity. And also the new flux linkage $F_D$ is less distortion than the old one. As a result, we can conclude that the performance under our control policy is better than the prior work [20].

However, if the motor is operated in the low speed, the mover voltage ($Ri_m$) is the same order to the input voltage ($u_m$). One should be obtained a large estimation error if the mover voltage is neglected. Accordingly, it becomes into a hard work for estimating the flux correctly. A position error should be appeared due to an incorrect value of the thrust estimation. And also the incorrect value of the thrust estimation is generated by the output of the incorrect flux. In fact, the incorrect value of the thrust estimation is due to an unsuitable space voltage vector. It hints the switching algorithm cannot run correctly. For reducing the influence of the low speed condition, the current-based method is needed in new proposed estimator algorithm. The new one can be considered as a flux estimated compensator. The block diagram with a low speed compensator is shown in Fig. 4. Notice that the flux estimated compensator is effective only in the low speed operation region. A correct value of the thrust estimation can be obtained if the motor is operating in the high-speed region. As a result, the new switching algorithm can provide proper voltage vectors for the driver of the motor. It is worthy to notice that the new flux estimator is better than the conventional one. Especially, the problems of the motor under the low operating speed region can be solved by our control policy. As a result, we can conclude that the new estimation algorithm may obtain a better performance than the prior research. And also the inherent problems of the system should be eliminated. The performances of the flux linkage response with two different control policies are depicted in Fig. 5.

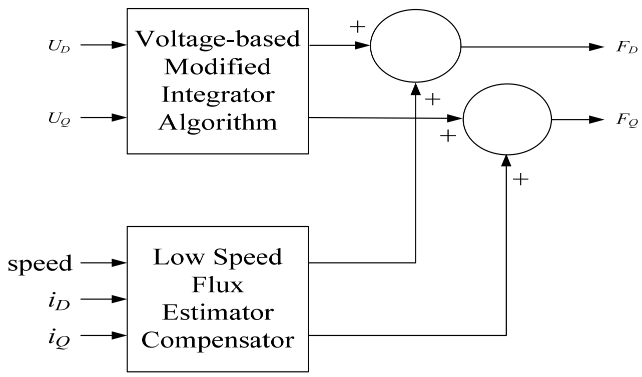


Figure 4. A flux estimator block diagram with flux compensator.

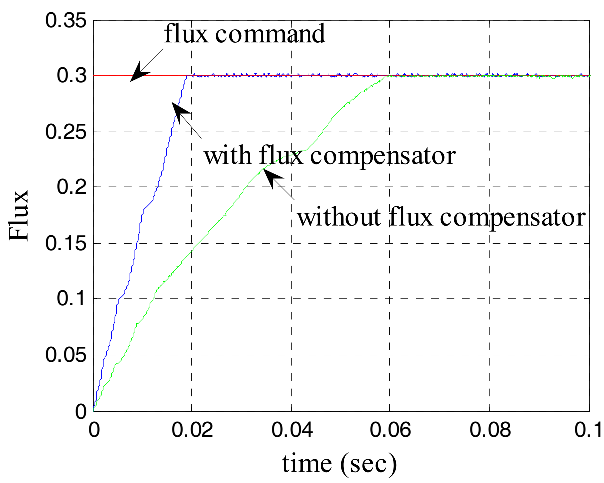


Figure 5. The flux linkage response with/without flux compensator.

## III. DESIGN OF A FAST TERMINAL SLIDING MODE CONTROLLER FOR PMLSMS

The dynamics of the PMLSM can be represented by the second-order differential equation (6).

$$F_e = M\ddot{x}_p + B\dot{x}_p + F_L \tag{6}$$

where $F_e$ is the output electromagnet thrust, $M$ is the mass of moving inertia, $B$ is the friction factor and $F_L$ is the external disturbance term. The model represented by (6) shows the mechanical motion may be perceived as a disturbance force that satisfies matching conditions. It means that the sliding mode control systems are able to reject the influence of the load nonlinearity on the mechanical motion. In addition, it is obviously that the lumped disturbance consisting of the external force acting on the system and the hysteresis can be estimated. Therefore, the application of the disturbance rejection method in the overall system design is allowed. To facilitate the derivation of the control law, (6) is written into the state-space form

$$\dot{x}_1 = x_2, \qquad x_1 = xp$$

$$\dot{x}_2 = -\frac{B}{M}x_2 + \frac{u}{M}, u = F_{cmd} \tag{7}$$

Define the position tracking error $e_1 = r_1 - x_1$, $r_1 = xpcmd$ and compute its dynamics as

$$\dot{e}_1 = \dot{r}_1 - \dot{x}_1 = r_2 - x_2 = e_2 \tag{8}$$

where $r_1$ ($r_2$) is the position (velocity) reference signal, $xp$ is the actual position signal, $e$ is the state vector $e^T = [e_1\ e_2]$, and $u_t$ is the control input. Then the error model state equation is corresponding to (7) that can be described as

$$\dot{e}_1 = e_2$$

$$\begin{aligned}\dot{e}_2 &= \ddot{r}_1 - \dot{x}_2 \\ &= \ddot{r}_1 + \frac{B}{M}x_2 - \frac{1}{M}u\end{aligned} \tag{9}$$

According to the fast terminal sliding mode concept, the recursive structure is depicted as [13]

$$\begin{aligned}s_0 &= e_1 \\ s_1 &= \dot{s}_0 + \alpha s_0 + \beta s_0^{q_0/p_0}\end{aligned} \tag{10}$$

here $\alpha$, $\beta$ and $q_0$, $p_0$ ($q_0 < p_0$), are positive odd integers. It is clearly that $s_0$ will approach zero gradually when $s_1$ reaches zero. The sliding mode control process involves the reaching phase from initial condition to sliding surface firstly and the second is driven the system states to equilibrium point in sliding phase. Based on the FTSM concepts, the fast terminal attractor is described as $s = \dot{e}_1 + \alpha e_1 + \beta e_1^{q/p} = 0$. It can be easily verified that the finite sliding time from $e(0) \neq 0$ to $e(0) = 0$ is

$$t_s = \frac{p}{\alpha(p-q)}\ln\frac{\alpha[e(0)]^{1-q/p} + \beta}{\beta} \tag{11}$$

The conventional TSM design rule is defined as $s_1 s_1\dot{s}_1 < -\varepsilon|s_1|$, which controller includes discontinuous term that will cause chattering behaviors in sliding mode control. Instead of it, the controller design in FTSM [13] is described as $\dot{s}_1 = -\phi s_1 - \gamma s_1^{q/p}$. The advantages of the change are that the controller has the following several items: 1) The reaching time $t_{s1}$ and sliding time $t_s$ can be tuned by setting parameters $p$, $q$, $\alpha$, $\beta$ and $\gamma$. 2) The control law is continuous so the chattering behaviors can be reduced. 3) It is robust to parameter uncertainty and external disturbance. The detail description is indicated as following.

Theorem 1[13]: For system (9), if we choose the following control law

$$u(t) = M[\ddot{r}_1 + \frac{B}{M}x_2 + \alpha\dot{e}_1 + \beta\frac{d}{dt}s_0^{q_0/p_0} + \phi s_1 + \gamma s_1^{q/p}] \tag{12}$$

with $s_0 = e_1$ , then the system states will reach the sliding manifold $s_1 = 0$. And also it will follow the terminal attractor $\dot{s}_1 = -\phi s_1 - \gamma s_1^{q/p}$ in finite time $t_{s1}$, where

$t_{s1} = \frac{p}{\phi(p-q)}\ln\frac{\phi[e_1(0)]^{1-\frac{q}{p}} + \gamma}{\gamma}$ with $\phi, \gamma > 0$, and $p$, $q$ are odd positive integer ($q < p$). The system will follow the recursive structure (10) to converge to the system equilibrium in finite time.

In theorem 1, for the system (9) and control law (12) that can force the system to approach the nonlinear switching manifolds in the reaching phase. Then the fast terminal sliding mode will converge its equilibrium from any arbitrary initial condition $e(0) \neq 0$ converge to $e(0) = 0$ in finite time.

The design task of a FTSM control is to determine a control parameter set such that the required performance of the controlled system can be met. It is worthy to emphasize that various methods [12-13] for this field are proposed. However, a direct and efficient control methodology has not been proposed. The authors would like to propose the control policy which is able to select the controller parameters automatically. And the selected parameters can meet the requirement performances. The requirement performances include short rise time, short settle time, low steady-state error and low tracking error. Notice that the conventional control policies usually hired trial-and-error/experiment to obtain the parameters for the requirement performances. In this paper, an intelligence algorithm is incorporated with FTSM to generate and optimize the parameters in automatic for a PMLSM control system. As a result, the authors employ the FTSM based on nonlinear control design (NCD) to demonstrate the controlled results.

## IV. Simulation Results

In this paper, an illuminated nonlinear control design (NCD) toolbox is used to obtain the vector $[\alpha, \beta, \phi, \gamma]$ for the FTSM controller. In the following simulation work, the authors want the specify of the control system is as follows: the rise-time $t_r \leqq 0.1$ sec, the settle-time $t_s \leqq 0.2$ sec, and the overshoot $\leqq 2$ %. The fixed parameter is set as $p = 3$, $q = 1$, $p_0 = 9$, and $q_0 = 5$. The authors hope the parameter vector $[\alpha, \beta, \phi, \gamma]$ is arranged between its lower and upper bound (i.e., [1, 1, 1, 1] and [50, 50, 50, 50]). Finally, the suitable parameter vector [35.24, 7.72, 24.83, 14.07] which is met the desired system criterion can be obtained.

The performance of the proposed integration algorithms is investigated by Matlab/Simulink. And the control system is shown in Fig. 6. For this simulation work, some information of the motor's data has to be involved in. The values of the PMLSM model parameter is listed in Table I. Some measured waveforms are provided here to validate the theoretical analysis. The authors use three different input signals (triangular, sine and trapezoid waves) to trace the performance. Fig. 7(a)-(c) shows the performance under a triangular-wave input signal. Fig. 8-9 shows the measured full stroke system performance using sine-wave and trapezoid-wave input signal, respectively. Based on the simulation results above, the trace of the output response is almost following the input command. Nevertheless, two brilliant performances are revealing under our controller policy. One is a fast response performance and the other one is a well trace performance.

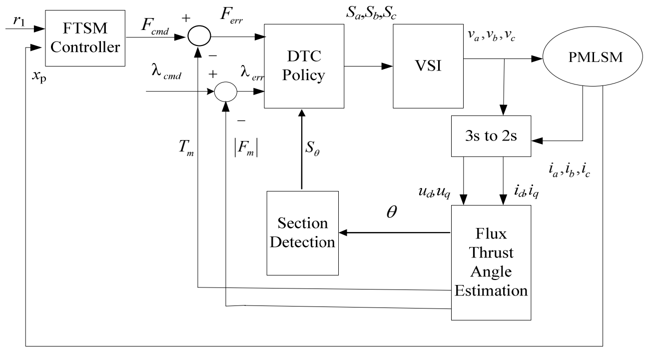


Figure 6. The block diagram of a simulated PMLSM control system.

TABLE I. The Parameters of the PMLSM.

| Parameter | Value | Parameter | Value |
|---|---|---|---|
| Mover mass ($Kg$) | 1.97 | Phase resistance ($\Omega$) | 11.8/3 |
| Friction factor ($N/(m/sec)$) | 5.2982 | Inductance ($mH$) | 4.8/3 |
| Flux linkage ($Wb$) | 0.4849 | Number of pole pairs | 1 |
| Pole pitch ($m$) | 0.06096 | Voltage ($V$) | 280 |
| Cut off frequency $\omega_c$ ($Hz$) | 100 | Flux linkage ($Wb$) | 0.3 |

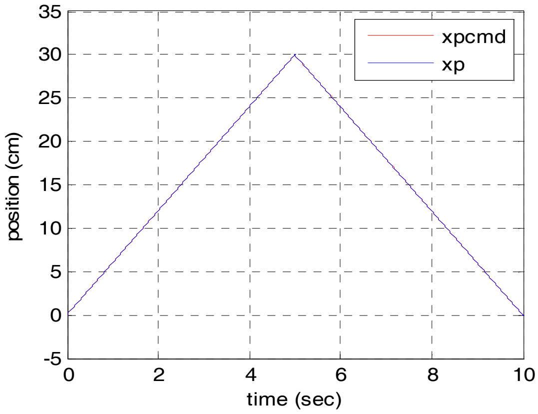


Figure 7(a). The simulation result for triangular-wave input tracking ($xp$).

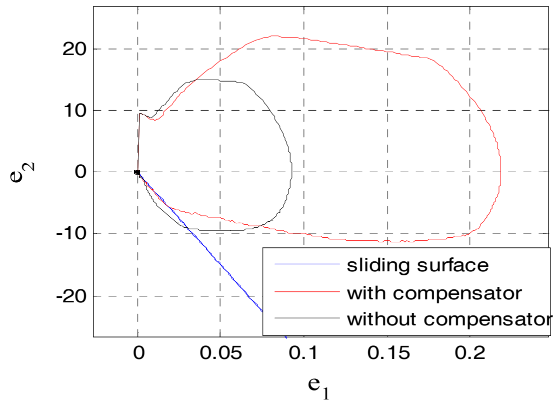


Figure 7(b). The simulation result for sliding surface.

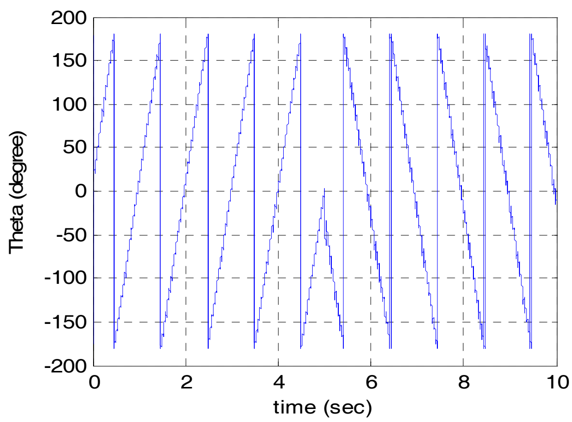


Figure 7(c). The simulation result for flux position response ($\theta$).

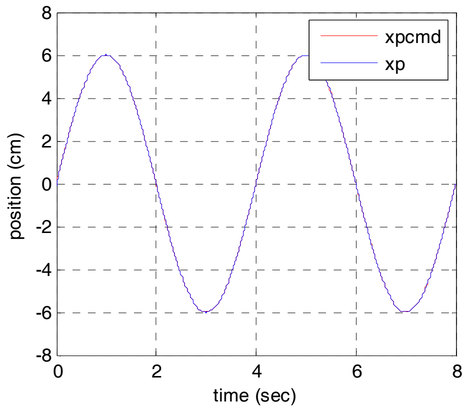


Figure 8. The simulation result for position tracking (*xp*).

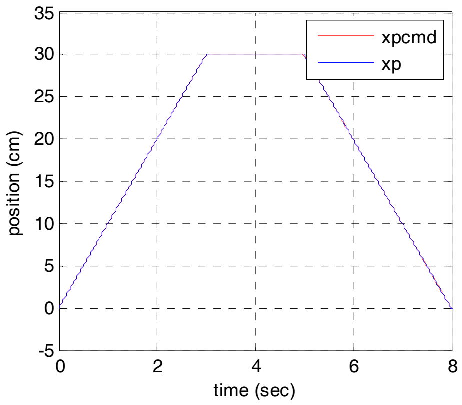


Figure 9. The simulation result for position tracking (*xp*).

## V. CONCLUSION

Based on the simulations results, we can conclude that, first, the fast terminal sliding mode controller guarantees the system can reach the sliding manifolds in finite time. Second, this paper proposes a new flux linkage estimator to obtain the actual flux magnitude and the value of its position angle. And then the direct thrust control policy can be implemented to PMLSMs. Third, the troublesome chattering modes in the conventional sliding mode control systems can be reduced. This study has successfully merged a FTSM controller and a direct thrust control policy for the motion of a PMLSM system. And the control system is robust under our proposed control policy. The complexity nature due to the inherent flux linkage problems and the imposition of underlying control policies is considered by our work. Numerical simulation results are provided to verify the effectiveness of the proposed control system.

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