Terminal sliding mode speed control method of permanent magnet synchronous linear motor based on adaptive parameter identification

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Abstract
In order to make the speed adjustment system of the permanent magnet synchronous linear motor more stable and make the convergence performance of the speed control system better, a speed regulator based on the terminal sliding mode algorithm is proposed. In addition, the permanent magnet synchronous linear motor will also be affected by temperature changes during its operation. Aiming to enhance the adaptability of linear motors to parameter changes, a terminal sliding mode speed control idea based on adaptive parameter identification is proposed. The Popov stability theory and the mathematical model of linear motor are used to construct a parameter identification system. The identification flux parameters are used as the control algorithm of the terminal sliding mode speed regulator and the update matching of speed regulator flux parameters and motor system parameters are realized. Through the above algorithm, the performance of the entire speed control system is improved. The results of simulation experiments can illustrate the effectiveness of this control method and its effect on improving the control performance.

Keywords
Terminal sliding mode control, permanent magnet synchronous linear motor, parameters identification, model reference adaptation

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Introduction
The motor described in the paper is a permanent magnet synchronous type. The only similarity between a linear motor and a rotating motor is the operating principle. There is a big difference between the two in the way of movement and the structure of the composition.1,2 The linear motor eliminates the transmission screw, biting gear, etc., converts the rotary motion into linear motion, and eliminates the mechanical transmission structure. The linear motor has a greater improvement in thrust, speed, power, etc.3–5 But linear motors also have shortcomings, such as the nonlinearity of the system, and the fracture effect caused by the specific structure. The strong coupling of the mathematical model, etc., these factors bring greater difficulty to the design of the controller. While satisfying the tracking and regulation performance, it is also necessary to solve the influence of parameter changes on the regulator.
order to further improve the effect of the speed regulator, it is necessary to solve this problem.6–8

In modern control strategies, the sliding mode variable structure control (SMC) algorithm is simple, has a fast response, and is robust to external noise interference and parameter perturbation, especially for the control of nonlinear systems. The SMC algorithm has been continuously developed by scholars in complex control applications. Ren et al.9 combined adaptively solving boundary problems with sliding mode control, and successfully applied them to the control of brushed motors. Wai and Muthusamy proposed to weaken sliding mode chattering and reduce the system’s requirements for detailed information. A fuzzy neural network genetic sliding mode control strategy is proposed. The combination of strong anti-interference and fast convergence of sliding mode is realized, and the tracking ability of the manipulator is improved.10 It can be concluded that sliding mode control has strong compatibility with other complex algorithms,11,12 which enables it to give full play to its advantages. In order to avoid the problem of sliding mode control reaching the sliding mode surface in an infinite time, and to further limit the time when the sliding modulus reaches the sliding mode surface, terminal sliding mode control has received attention.

Tran and Kang13 proposed terminal sliding mode control. It shows better control performance for nonlinear systems, but its resistance to external disturbances is low. Junejo et al.14 designed a continuous terminal sliding mode speed controller with the goal of limiting the convergence time of the tracking speed, which has improved its convergence time, anti-interference ability, and chattering suppression. However, the controller parameters are fixed, and real-time matching cannot be achieved, resulting in the controller’s weak ability to resist interference from uncertain factors. Panah et al.15 proposed the use of adaptive estimation of uncertain factors and then combined with sliding mode control to reduce tracking error and improve system accuracy, but its system convergence speed is affected and the convergence time is extended. Wang et al.16 proposed the use of model reference adaptive algorithm to identify motor parameters, and then use the identified parameters to participate in the current prediction controller. Realize the online update of the parameters in the prediction algorithm. It reduces the static error of the motor current and improves the control performance of the system. Zhang et al.17 uses MRAS to calculate motor parameters and uses simulation and experiments to prove that parameter identification can correctly converge to the motor parameter values. However, it is only used for parameter observation, and the identification parameters are not combined with the controller application. Li and Chen18 describes the relationship between temperature and motor parameters when the motor is running. The saturation of the magnetic circuit and temperature changes can have a maximum impact of about 20% on the permanent magnet.

In order to improve the regulator’s resistance to uncertain factors and obtain the desired control effect, the paper proposes a terminal sliding mode speed regulator with adaptive parameter identification. This control method combines the advantages of self-adaptive identification of motor parameters and the finite time convergence of terminal sliding mode control. In addition to making the convergence of the speed error limited in time, an adaptive parameter identification method is also used. It is applied to terminal sliding mode controllers to enable the speed controller to obtain motor parameters and realize controller parameter update. This allows the controller to cope with the influence of the motor parameter changes on the controller, reduces the influence of uncertain external factors on the system, and increases the robustness of the system. Section 4 of this paper gives simulation and experimental results to illustrate the usability of the proposed strategy.

**PMSLM mathematical modeling**

The PMSLM system is a multivariable, strongly coupled and nonlinear system.19 For the purpose of understanding the system more conveniently, first make assumptions. Assuming that the magnetic field between the primary and secondary of the linear motor is uniformly distributed, the first step is to determine the mathematical model of PMSLM. The principle structure diagram of PMSLM is shown in Figure 1.

Based on the conversion of the rotating coordinate system, there is \( L_d = L_q \) in the surface-mounted PMSLM, so set the inductance \( L_d = L_q = L_s \), and the current equation of the PMSLM is obtained from the voltage equation:

\[
\begin{align*}
\frac{d i_d}{dt} &= -\frac{R_s}{L_s} i_d + \omega i_q + \frac{u_d}{L_s} \\
\frac{d i_q}{dt} &= -\frac{R_s}{L_s} i_q - \omega i_d + \frac{u_q}{L_s} - \frac{\psi_f}{L_s} \\
\omega &= \frac{\pi v}{\tau}
\end{align*}
\]

![Figure 1. PMSLM principle structure diagram.](image-url)
where $i_d$, $i_q$-the dq axis current component, $u_d$, $u_q$-the dq axis voltage component, $R_t$-winding resistance. $\tau$ is the polar distance. $L_r$-the inductance value. $\psi_f$-the linear motor flux linkage value. $\omega$-the equivalent rotating motor angular velocity. $\nu$-the linear motor linear speed. The thrust equation of the motor as:

$$ F_e = \frac{3\pi}{2\tau} p_n \psi_f i_q + (L_d - L_q) i_d i_q $$  \hspace{1cm} (2) $$

where $p_n$ is the polar logarithm of the PMSLM. Based on the same conditions as the above inductance components, the electromagnetic thrust equation is simplified to:

$$ F_e = \frac{3\pi}{2\tau} p_n \psi_f i_q $$  \hspace{1cm} (3) $$

According to the relationship between the electromagnetic thrust and the speed of the linear motor, the motion formula of the linear motor could be obtained follow:

$$ m \frac{dv}{dt} = F_e - Bv - f $$  \hspace{1cm} (4) $$

where $m$-the mass of the mover. $f$-regarded as a system disturbance. From equations (2) and (4), we can get:

$$ \dot{v} = \frac{K_t}{m} i_q - \frac{B}{m} v - \frac{f}{m} $$  \hspace{1cm} (5) $$

in the above equation, $\frac{dv}{dt} = \dot{v}$, $K_t = \frac{3\pi}{2\tau} p_n \psi_f$. $B$-the viscous friction coefficient.

### Speed regulator design

#### Speed loop terminal sliding mode controller design

Aiming to promote the convergence of the speed regulator and the accuracy of tracking the given speed, a first-order terminal sliding mode speed controller is designed based on the terminal sliding mode theory. Taking PMSLM as the controlled object and based on TSMC theory, the high-level system is shown as follows:

$$ \begin{cases} x_1 = x_2 \\ \dot{x}_2 = x_3 \\ \vdots \\ \dot{x}_{n-1} = x_n \\ \dot{x}_n = f(x, t) + d(x, t) + b(x, t)u \end{cases} $$  \hspace{1cm} (6) $$

As shown in formula (6), $n$ represents the order of the above higher-order model, and $d(x, t)$ is an external interference factor and it is generally bounded in a linear motor. $x$ means system status.

For the high-order system shown in equation (6), when the following sliding mode surface is designed, the system will converge within a limited time, namely:

$$ s = s + c_1 \text{sgn}(x_1)|x_1|^n + \cdots + c_n \text{sgn}(x_n)|x_n|^n $$  \hspace{1cm} (7) $$

where $c_1$ and $c_n$ are constants to be selected. Among them:

$$ \begin{cases} s_1 = s, n = 1 \\ s_{i-1} = \frac{c_{i+1}}{2c_i}, \quad i = 2, 3, \ldots, n \end{cases} $$  \hspace{1cm} (8) $$

For a first-order system, $\alpha \in (1 - \epsilon, 1), \epsilon \in (0, 1)$.

Because the controller designed in this paper is a first-order terminal sliding mode algorithm, it is necessary to process and simplify equation (6) to obtain the first-order equation and integrate it into the motor model. Let the system order $n = 1$. From equation (5), the linear motor’s first-order speed differential equation is:

$$ \dot{v} = k_i \frac{d_i}{m} i_q - \frac{B}{m} v - \frac{f}{m} \left( \frac{i_q}{i_q} - \frac{i_d}{i_d} \right) $$

$$ = b_0 i_q + d(t) $$  \hspace{1cm} (9) $$

where $i_q$ is the expected output control quantity, $b = \frac{k_i}{m} d(t) = -\frac{B v}{m} - \frac{f}{m} \left( \frac{i_q}{i_q} - \frac{i_d}{i_d} \right)$. The error variable required to realize sliding mode control is described as speed error. Suppose the desired speed is $\nu_r$, and the current speed is $\nu$:

$$ \begin{cases} x_1 = \nu_r - \nu = e \\ x_2 = \dot{x}_1 = \dot{\nu}_r - \dot{\nu} = \dot{e} \end{cases} $$  \hspace{1cm} (10) $$

From equations (9) and (10), we can get:

$$ \dot{e} = \nu_r - b_0 i_q - d(t) $$  \hspace{1cm} (11) $$

According to equation (7), the design of the sliding surface is:

$$ s = \dot{e} + c \text{sgn}(e)|e|^n $$  \hspace{1cm} (12) $$

The selection control law is:

$$ \begin{cases} \dot{i}_q = b_0 \left( u_{eq} + u_n \right) \\ u_{eq} = \dot{\nu}_r + c \text{sgn}(e)|e|^n \\ u_n = \int \text{sgn}(s)d\tau \end{cases} $$  \hspace{1cm} (13) $$

where $b_0 = \frac{m}{k_i}$, $m = \frac{2km}{\pi p_n \psi_f}$. Combining equations (11) and (12), (13), the sliding mode surface could be expressed:

$$ s = c \text{sgn}(e)|e|^n + \dot{\nu}_r - \left( u_{eq} + u_n \right) - d(t) $$

$$ = -u_n - d(t) $$  \hspace{1cm} (14) $$
Suppose the Lyapunov function is $V = \frac{1}{2}x^2$, combined with equations (13) and (14), we can get:

$$\dot{V} = s\dot{s} = s(-\dot{u}_n - \dot{d}(t)) = -k|s| - \dot{d}(t)s \quad (15)$$

Slightly sorting out equation (15), we can get:

$$\dot{V} \leq - (k - |\dot{d}(t)|)|s| \leq (k - |\dot{d}(t)|)|s| \leq -\sqrt{2(k - k_d)}V^\frac{1}{2}$$

where $k_d$ is the derivative limit of the system disturbance $d(t)$. $k_d$ and $k$ are constants greater than 0, when $k > k_d, \dot{V} < 0$ according to the Lyapunov stability theorem, the system converges stably and satisfies the sliding mode surface within a limited time. Figure 2 is the principle structure of the speed regulator algorithm.

Adaptive parameter identification

The advantage of adaptive parameter identification is that the algorithm structure is simple and easy to implement. First, two models need to be established, one is a reference model, which can be obtained by equation (1), and then an adjustable model is established. The form is basically the same as the reference model, but the calculation parameters in the model are identification parameters. The current error models of the two established models are converged by the adaptive law, so that the parameters are consistent, and the identification parameters are estimated. First establish a reference model, and write equation (1) into the form of the state equation as follows:

$$\begin{bmatrix}
\frac{di_d}{dt} \\
\frac{di_q}{dt}
\end{bmatrix} = \begin{bmatrix}
\frac{-R_s\omega}{L_s} & \frac{-R_s}{L_s} \\
0 & \frac{-\omega}{L_s}
\end{bmatrix} \begin{bmatrix}i_d \\
i_q
\end{bmatrix} + \begin{bmatrix}0 \\
\frac{1}{L_s}
\end{bmatrix}u_d + \begin{bmatrix}0 \\
\omega \psi_f / L_s
\end{bmatrix}$$

(17)

The differential operator in the above equation is expressed by $d$. For the convenience of equation expression, the parameters in equation (17) are expressed as follows:

$$a = \frac{R_s}{L_s}, b = \frac{1}{L_s}, c = \frac{\psi_f}{L_s} \quad (18)$$

Equation (17) is simplified as:

$$\begin{bmatrix}
\frac{di_d}{dt} \\
\frac{di_q}{dt}
\end{bmatrix} = A^\hat{i} + Bu + \hat{C} + k(i - \hat{i}) \quad (19)$$

The current in the above formula is expressed as $i = [i_{d1} i_{d2}]^T$. The voltage state is expressed as $u = [u_d u_q]^T \cdot A = \begin{bmatrix}-\alpha & \alpha \\
-\omega - \alpha & -\omega - \alpha
\end{bmatrix}, B = \begin{bmatrix}b_0 \\
0\end{bmatrix}, C = [0 - \alpha c]$. According to equation (19), the adjustable reference model is constructed as follows:

$$\begin{bmatrix}
\frac{di_d}{dt} \\
\frac{di_q}{dt}
\end{bmatrix} = A^\hat{i} + Bu + \hat{C} + k(i - \hat{i}) \quad (20)$$

where $\hat{i} = [i_{d1} i_{d2}]^T$ in equation (20) represents the estimated current value of the reference model. Coefficient matrix $\hat{A} = \begin{bmatrix}-\alpha & \alpha \\
-\omega - \alpha & -\omega - \alpha
\end{bmatrix}, \hat{B} = \begin{bmatrix}b_0 \\
0\end{bmatrix}$. The gain coefficient matrix $\hat{C} = [0 - \alpha c], k = \begin{bmatrix}k_10 \\
k_20\end{bmatrix}$ is the gain coefficient matrix. $\alpha = \dot{R_s}/L_s, b = 1/L_s, \psi_f / L_s, \dot{L_s}, \dot{\psi_f}$ are the identification values of resistance, inductance and flux linkage respectively. Use equation (19) to subtract equation (20), and get the error mathematical model after finishing:

$$de = (A + k)e + \Delta A\hat{i} + \Delta Bu + \Delta C \quad (21)$$

where $e = i - \hat{i}$ is the error between actual current and estimated current. The coefficient difference is:

$$\Delta A = A - \hat{A}, \Delta B = B - \hat{B}, \Delta C = C - \hat{C}.$$ Let $w = -\left(\Delta A\hat{i} + \Delta Bu + \Delta C\right)$, equation (21) can be written as:

$$de = (A + k)e - w \quad (22)$$

According to Popov’s stability theory, one of the necessary conditions to make the designed model stable is the positive real matrix $\left[S - (A + k)\right]^{-1}$ of the error transfer function. According to this rule, the gain
matrix can be designed, where \( I \) is the second-order identity matrix. The design of the parameter adaptive law requires that the nonlinear time-varying part conform to the Popov integral inequality:

\[
\eta(0, t) = \int_0^t w e^T d\tau \geq - \gamma_0^2 \forall t > 0
\]

(23)

where \( \gamma_0 \) is a bounded normal number. Putting equation (22) into equation (23), we can get:

\[
\eta(0, t) = \int_0^t -e^T (\Delta A + \Delta Bu + \Delta C) dt \geq - \gamma_0^2
\]

(24)

Expand and decompose equation (24):

\[
\begin{cases}
\int_0^t (a - \hat{a}) (i_d e_d + i_q e_q) dt \geq - \gamma_1^2 \\
\int_0^t (b - \hat{b}) (u_d e_d + u_q e_q) dt \geq - \gamma_2^2 \\
\int_0^t (c - \hat{c}) e_v dt \geq - \gamma_3^2
\end{cases}
\]

(25)

Through the above analysis, if you want to maintain the stability of the system, you need to meet the conditions of equation (25). Taking \( \gamma_1 \) as an example, the analysis is as follows. From the first inequality, it can be written as:

\[
\int_0^t (a - \hat{a})(e_d i_d + e_q i_q) dt \geq - \gamma_1^2
\]

(26)

Assume \( a \) as \( a - \hat{a} \) proportional integral structure:

\[
a - \hat{a} = \int_0^t p_1(\tau) d\tau + p_2(\tau)
\]

(27)

Incorporate equation (27) into equation (26):

\[
a - \hat{a} = \int_0^t \int_0^t p_1(\tau) d\tau (e_d \dot{i}_d + e_q \dot{i}_q) dt
\]

\[
+ \int_0^t p_2(\tau) (e_d \dot{i}_d + e_q \dot{i}_q) dt \geq - \gamma_1^2
\]

(28)

Decompose formula (28) into two inequalities and discuss separately:

\[
\begin{cases}
\int_0^t \int_0^t p_1(\tau) d\tau (e_d \dot{i}_d + e_q \dot{i}_q) dt \geq - \gamma_{11}^2 \\
\int_0^t p_2(\tau) (e_d \dot{i}_d + e_q \dot{i}_q) dt \geq - \gamma_{12}^2
\end{cases}
\]

(29)

For the first inequality in inequality (29), use the following formula:

\[
\int_0^t \frac{df(t)}{dt} kf(t) dt = k \left( f^2(t) - f^2(0) \right) \geq k \frac{1}{2} f^2(0)
\]

(30)

Take \( \frac{df(t)}{dt} = (e_d \dot{i}_d + e_q \dot{i}_q) \), \( kf(t) = \int_0^t p_1(\tau) d\tau \) and derivate both sides of \( kf(t) = \int_0^t p_1(\tau) d\tau \) to get:

\[
p_1(\tau) = k_i (e_d \dot{i}_d + e_q \dot{i}_q), k_i > 0
\]

(31)

Incorporating equation (31) into the first inequality in (29) can make this inequality true. For the second inequality, just take \( p_2(\tau) = k_p (e_d \dot{i}_d + e_q \dot{i}_q), k_p > 0 \) to make it true. So we can get:

\[
a - \hat{a} = k_i \int_0^t (e_d \dot{i}_d + e_q \dot{i}_q) d\tau + k_p (e_d \dot{i}_d + e_q \dot{i}_q)
\]

(32)

The derivation process of the adaptive laws for other motor parameters is similar. Through the above description, the adaptive parameter system conforms to the Popov superstability theory and can achieve stability.

Among them, \( \gamma_1, \gamma_2, \gamma_3 \) are bounded positive real numbers. The adaptation law of the design identification parameter is PI form to ensure that equation (24) holds. By introducing the basic form of continuous derivable function and proportional integral and the above derivation process, the parameter adaptation law can be obtained:

\[
\begin{cases}
\dot{a} = -(k_p + \frac{k_i}{\gamma_1^2}) (i_d e_d + i_q e_q) + \hat{a}(0) \\
\dot{b} = (k_p + \frac{k_i}{\gamma_2^2}) (u_d e_d + u_q e_q) + \hat{b}(0) \\
\dot{c} = -(k_p + \frac{k_i}{\gamma_3^2}) e_v + \hat{c}(0)
\end{cases}
\]

(33)

In the equation, \( k_p \) and \( k_i \) are the proportional integral parameters to be adjusted, and \( \hat{a}(0), \hat{b}(0), \) and \( \hat{c}(0) \) are the initial values of the identification parameters. \( e_d \) and \( e_q \) are the current component errors of the Cross axis respectively. Based on the mathematical and physical relationship between the identification value and \( R_s, L_r, \) and \( \psi_p \), the resistance, inductance and flux linkage values can be calculated. Figure 3 is a schematic diagram of the structure of the adaptive parameter recognition algorithm structure.
The reduction process is assumed to be a transient process. In the simulation, the dynamic motor is mostly a long dynamic process, but special cases are not ruled out. In the simulation, the dynamic motor flux linkage coefficient of viscous friction mover mass and the response of the regulator.

Table 1. Simulation experiment motor parameters.

| Parameter meaning                     | Value          |
|---------------------------------------|----------------|
| Stator winding $R_s$                  | 2.6 Ω          |
| d-axis inductance $L_d$               | 0.062 mH        |
| q-axis inductance $L_q$               | 0.0627 mH       |
| Mover mass $m$                        | 5.6 kg          |
| Coefficient of viscous friction $B$   | 0.2 ms          |
| Polar distance $V$                    | 0.018 m         |
| Flux linkage $w_b$                    | 0.24 WB         |

Although the flux linkage value reduction time becomes shorter, it does not affect the test result. Set the reference value of the flux linkage to 0.20 $w_b$, at 0.4 s. Figure 6 is the result of simulation identification when the flux linkage is set to decrease.

Figure 7 is a comparison waveform of the output current control amount of the speed regulator with and without parameter identification. Reduce the value of the motor flux to 0.20 $w_b$ at 0.4 s, and the control output of the speed controller combined with parameter identification can quickly reach the desired current. And after the control quantity output is stable, compared with the output fluctuation range, the output of the speed controller with parameter identification is more stable.

Figure 8 shows the comparison result of the speed tracking response curve of the speed regulator with and without parameter identification. At 0.4 s, the flux linkage of the motor decreases from 0.24 $w_b$ to 0.20 $w_b$, and the change of the flux linkage as the interference of uncertain factors causes a certain degree of decrease in the steady state speed. The controller with parameter identification obtains the information of the flux linkage change. When the controller flux linkage parameter is 0.4 s, the change is synchronized with the motor parameter change. The control law output of equation (12) is adjusted in time, so that the output control amount is adjusted, and the influence of the interference of uncertain factors on the speed is reduced.

Figure 9 shows the comparison results of the speed tracking response curve of the speed regulator when the without parameter identification of Figure 9 is obtained is a partial enlarged view. At 0.2 s, the load is 100 N, and the motor flux linkage value is changed to 0.20 $w_b$ at 0.4 s. But the controller flux linkage value remains unchanged. No parameter identification makes the controller’s flux linkage parameters inconsistent with the actual flux linkage value of the linear motor. The conclusion is that when the flux linkage changes, it is equivalent to adding a very small load to the system, which interferes with the speed.

For the convenience of observation, the speed error waveform get by the terminal sliding mode controller when the without parameter identification of Figure 9 is obtained is a partial enlarged view. At 0.2 s, the load is 100 N, and the motor flux linkage value is changed to 0.20 $w_b$ at 0.4 s. But the controller flux linkage value remains unchanged. No parameter identification makes the controller’s flux linkage parameters inconsistent with the actual flux linkage value of the linear motor. The conclusion is that when the flux linkage changes, it is equivalent to adding a very small load to the system, which interferes with the speed.

Figure 10 shows the comparison results of the three speed control methods of PI control, traditional SMC, and TSMC based on parameter identification proposed in this paper. During the motor starting phase, the time for the linear motor to stabilize at a given speed has been significantly improved. In order to facilitate observation, 200 N load is added at 0.2 s. Through comparison, it can be obtained that the speed response curves of traditional sliding mode control and proportional integral control are basically the same, but the anti-load ability of traditional sliding mode is stronger than that of proportional integral method. The terminal sliding mode speed regulator based on parameter identification has relatively good performance in anti-load interference and recovery stability time.

Figure 4. Block diagram of the overall system design.

Figure 5(a)–(c) are the identification waveform results of the magnetic flux linkage, inductance and resistance of the linear motor. When the time is about 0.1 s, the output value of the parameter observer converges to the true value, and the flux, inductance, and resistance. The real values are shown in Table 1. This paper mainly discusses the mutation of the flux linkage and the response of the regulator.

Figure 6 is the tracking of the flux linkage identification result when the motor flux linkage parameter is set to 0.20$w_b$. In practice, the flux linkage change of the motor is mostly a long dynamic process, but special cases are not ruled out. In the simulation, the dynamic reduction process is assumed to be a transient process.

System simulation results

The simulation system adopts $i_d = 0$ control strategy, and the current loop controller adopts PI control mode. The results obtained using simulation software to illustrate the method proposed in the article. The effectiveness of the method is demonstrated through analysis and comparison. Table 1 shows the specific parameters of the simulation.

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| Polar distance $V$ | 0.018 m     |
| Flux linkage $w_b$ | 0.24 WB     |
Figure 11 shows the comparison results of the three speed controllers’ steady-state speed errors. Because the controller proposed in this paper contains an integral term in the output nonlinear control quantity, it is equivalent to adding a low-pass filter to the output control quantity, so the overall output chattering is suppressed, so it achieves a better control effect, thus get the speed waveform with less chattering, make the speed control more stable. However, it can be concluded that the other two control methods do not

Figure 5. Parameter identification results: (a) flux identification result, (b) inductance identification result, and (c) resistance identification result.

Figure 6. Variable flux identification results.

Figure 7. Controlled q-axis current comparison chart.

Figure 8. Parameter identification terminal sliding mode speed comparison.
significantly suppress chattering when the speed reaches a steady state.

Figure 12 shows the comparison waveforms of the speed control performance of the three speed controllers. The initial speed is set to 0.8 m/s. When the speed is stable, the rising speed is 1 m/s at 0.1 s, and when the speed reaches a steady state, it decelerates to 0.8 m/s at 0.2 s. During the startup process, the rise time of the terminal sliding mode has a small increase, but compared to the other two controllers, the adjustment time is very small. During the speed-up process, the terminal sliding mode controller has a significant improvement in the convergence speed and overshoot. During the deceleration process, the terminal speed controller has a small overshoot, but the overall tracking effect is better. Converge to the expected value in a short time.

**System experiment**

The experimental verification of this article is based on the STM32 development board and Simulation software’s automatic code generation tool. The use of STM32CubeMX software to generate the underlying configuration code allows R&D personnel to focus on the algorithm implementation part and improve work efficiency. The algorithm experiment platform is shown in Figure 13. The permanent magnet synchronous linear motor was used as the controlled object, and the control algorithm c code was generated using a code generation tool, and the algorithm was verified on the STM32 development board.

For motor parameter identification, due to the fixed structure of the motor, the experiment in this article only gives the parameter identification value when the motor parameters remain unchanged to verify the correctness of the identified parameters. Figure 14(a)–(c) are the identification values of resistance, inductance and flux linkage respectively. Comparing the values in the above linear motor parameter table, it can be seen that the identification results of each parameter...
converge to the corresponding parameter value within a certain period of time. Its error fluctuates in the range of 4%–6%, which can illustrate the effectiveness of the identification algorithm.

Figure 15 shows the speed response waveform when the desired speed of the system is set to 1 m/s. In the start-up phase, the rise time is about 0.01 s, the adjustment time is about 0.02 s, there is a small overshoot, and the overall controller dynamic response index has a good performance. Figure 16(a) shows the comparison waveform of the speed output between the TSMC and the traditional SMC. Add a load of 200 N at 0.1 s to get a load response. As shown in Figure 16, the terminal sliding mode’s anti-load capability and the time to recover stability are better than traditional sliding mode controllers.

In terms of speed adjustment, as shown in Figure 16(a), in the process of acceleration and deceleration, the overshoot of the traditional sliding mode is larger and the adjustment time is longer. Figure 16(b) shows the control output signals of the two controllers. It can be concluded from the figure that in the no-load phase, the SMC reaches a stable output at about 0.04 s, while the TSMC-MRAS reaches a stable output at about 0.02 s. In addition, when speed regulation occurs, the control output of this control method is more stable. Through comparison, we can see the effectiveness and superiority of the improved terminal sliding mode speed regulator in linear motors.

**Conclusion**

With the goal of enhancing the control performance of linear motors, this paper proposes a terminal sliding mode speed regulator based on parameter
identification. Combine parameter identification with terminal sliding mode speed regulator. The identification resistance and inductance are used for parameter observation, and the flux linkage value is used to improve the speed controller, so that the terminal sliding mode speed controller can adjust the flux linkage value online, thereby adjusting the control amount and improving the speed output. The parameter identification realizes that the motor parameters are consistent with the controller parameters, enhances the speed regulation and anti-interference ability of the entire system, and improves the system robustness.

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