Research Article

A New Linear Motor Force Ripple Compensation Method Based on Inverse Model Iterative Learning and Robust Disturbance Observer

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Permanent magnet linear motors (PMLMs) are gaining increasing interest in ultra-precision and long stroke motion stage, such as reticle and wafer stage of scanner for semiconductor lithography. However, the performances of PMLM are greatly affected by inherent force ripple. A number of compensation methods have been studied to solve its influence to the system precision. However, aiming at some application, the system characteristics limit the design of controller. In this paper, a new compensation strategy based on the inverse model iterative learning control and robust disturbance observer is proposed to suppress the influence of force ripple. The proposed compensation method makes fully use of not only achievable high tracking accuracy of the inverse model iterative learning control but also the higher robustness and better iterative learning speed by using robust disturbance observer. Simulation and experiments verify effectiveness and superiority of the proposed method.

1. Introduction

The linear motor originated in 1840, when Charles Wheatstone invented the first linear motor [1]. Today, with the rapid development of technology such as material processing and electronic technology, linear motor has entered a new stage of vigorous development. Nowadays, it has become an indispensable electric actuator in the industry field. The PMLM has various advantages, such as high force density, low heat loss, simple mechanical structure, long stroke, and high accuracy. Therefore, the linear motor is widely used in ultra-precision machining equipment. Among them, the demand of ultra-precision motion platforms in IC manufacturing industry is gradually increased. Obviously, the PMLM has become the most widely used actuator in semiconductor processes.

Different from rotary motors, linear motors do not require coupling mechanisms such as gearboxes, chains, or screw couplings. This feature greatly reduces the nonlinearity, flank

clearance, and frictional disturbances, especially under the support of aerostatic bearing or magnetic bearing [2]. However, the advantages of using mechanical transmission have disappeared either, such as its inherent ability to reduce model uncertainty and external disturbance. The most important and well-known nonlinear character of PMLM is the force ripple [3], which is caused by the magnetic structure and depends on the position and velocity. From the perspective of control, the force ripple is the key problem affecting system performance. Hence, it is imminent to solve this problem.

Nowadays, the force ripple suppression problem has become the focus. The first one is to reduce the force ripple from the design view by optimizing the motor structure. Bianchi et al. compensated the force ripple by method of sliding pole and verified the effectiveness of the method by means of finite element analysis and experiments [4]. Jung et al. suppressed the cogging force by optimizing the arrangement of the permanent magnets and obtained the optimal

length of the segmented permanent magnet model by three-dimensional equivalent magnetic circuit network method, which verified its applicability in the PMLM [5]. Chao proposed a double-layer Halbach secondary structure for coreless linear motor and verified the rationality of this motor design by means of finite element simulation to optimize the size parameters of the double-layer Halbach permanent magnet array [6].

Although these methods are proved to be very effective, the improvement accuracy is limited. Moreover, in some high-precision and high-speed applications, the method of replacing the moving icon core with epoxy resin material becomes worthless due to its lack of great thrust. Therefore, it is extremely important to seek appropriate control strategy to reduce the influence of the force ripple on system performance.

There are a lot of studies compensating the force ripple by control strategy. Kim et al. designed a feedforward controller to compensate force ripple in a coreless linear motor using a T-type magnet array, and the experiment verified its effectiveness [7]. Bascetta et al. identified the force ripple model of brushless AC motors by frequency domain method and performed direct compensation [8]. Zhu et al. extracted and compensated the force ripple of the PMLM using the vector control method [9]. Otten et al. combined feedback control and neural network feedforward control based on the simplified second-order model of linear motor, and the results show that the online learning controller has great application in tracking control tasks [10]. Tan et al. designed a compound control structure in linear motor system, including a simple feedforward part, a PID feedback part, and an adaptive feedforward compensator. The computer simulation and real-time experiments verified the effectiveness of the proposed method in the application of high-precision trajectory tracking [11]. Xu and Yao designed two adaptive robust controllers and carried out experiments on the linear motor with icon core [12]. Chen et al. proposed a new method based on relay identification method to identify parameters of force ripple and friction model in the linear motor [13].

In addition, there are also a number of useful methods for improving the performance of tracking control system, which can also be used for suppressing the impact of force ripple. Tseng et al. proposed a fuzzy control design method for nonlinear systems with a guaranteed H_∞ model reference tracking performance, and simulation example illustrated the design procedures and tracking performance of the proposed method [14]. Tao et al. designed adaptive laws for updating the controller parameters when both the plant parameters and actuator-failure parameters are unknown, and the simulation results showed that desired system performance is achieved with the developed adaptive actuator failure compensation control designs [15]. Yin et al. developed a fractional-order sliding mode-based extremum seeking controller for the optimization of nonlinear systems, and simulation and experimental results show that the method can have a faster convergence speed and a smaller neighborhood around the optimal operational point [16, 17]. Although the above methods can also effectively

improve tracking performance, they are too complex to apply in practice. And the problem in linear motors is certain, that is, the force ripple. So the most conventional means is to reduce the effect of the force ripple via analysis of force ripple's characteristics and compensation strategy design. Hence, the compensator design is considered in this paper.

Based on the aforementioned references, it was found that the disturbance observer is a very common method to compensate force ripple of linear motor. The basic idea of disturbance observer is to estimate the real-time system disturbance and output equivalent disturbance then conduct compensation to the control output [18]. Hwang and Seok designed Jacobian linearized observer for the hybrid stepper linear motor to ensure that the state estimation converged to the true value and designed an input-output feedback linearization controller for the speed loop and position loop [19]. Cho et al. proposed a periodic adaptive disturbance observer to weaken the periodic disturbance in repeated motions [20]. However, the traditional observer does not use the previous system motion information, so that it could not fully compensate for the spatial periodic force ripple, which is common in many applications, such as lithography machine, numerical control machine tool, and other high-precision manufacturing equipment.

In recent years, iterative learning control (ILC) methods have been widely used in practical engineering, such as robotic arms, intelligent transportation systems, and biomedicine [21–23]. The iterative learning theory is applicable to the controlled system with repetitive motion characteristics. It memorizes and learns the factors that cause errors, such as the uncertainties contained in the system. It uses the deviation of the output trajectory and the input trajectory to continuously correct the unsatisfactory control instruction and improve the tracking performance [24]. A large part of the PMLM force ripple is related to the position, which is spatially periodic. Therefore, ILC is widely used in the compensation of linear motor force ripple. Zhang combined PID feedback control with iterative learning intelligent feedforward algorithm, which is applied to the PMLM to verify the effectiveness of the method [24]. Mishra and Tomizuka designed and analyzed the iterative learning controller based on partial information for previous cycles to solve the repetitive error caused by phase-mismatch and force ripple [25].

Aiming at the practical problems of linear motor systems, the paper presents a compensation structure combining disturbance observer and ILC. It is expected that we can use the previous motion information to improve the system accuracy; at the same time, the disturbance observer is adopted to speed up the iterative speed and deal with the nonperiodic disturbances. However, it is difficult to achieve high compensation accuracy for the design of conventional disturbance observer is easy to be constrained by the controlled object. Therefore, in this paper, a novel observer with notch filter is used to suppress the resonant peak of disturbance output gain caused by conventional disturbance observer. Consequently, the control performance is improved comprehensively.

This paper is organized as follows. In Section 2, the problem description is presented. The method of IMILC-Robust

DOB is illustrated in Section 3. And the effectiveness of the proposed method is verified by experiments in Section 4. Finally, Section 5 provides concluding remark of the paper.

2. Problem Description

The development of linear motor can be traced back to 1840, and it has become an indispensable electric drive device in today's industrial field. The linear motor has the advantages of high force density, low heat loss, simple mechanical structure, large stroke, high precision, and so on. Different from rotary motors, linear motors do not require indirect coupling mechanisms such as gearboxes, chains, or spiral couplings. This feature greatly reduces the influence of nonlinearity, side clearance, and friction disturbance in linear motor system, especially after the air flotation support scheme is adopted. However, the advantages of using mechanical actuators also disappear, for example, their inherent ability to reduce model uncertainty and to resist disturbance. Among disturbances, the force ripple is one of the most adverse factors affecting the accuracy of the servo system, which is composed by ripple disturbance, cogging effect, and end effect [26, 27]. It would facilitate the compensator design by finding out the mechanism of this disturbance.

Firstly, the generation principle of ripple disturbance is studied. In the working process of linear motor, if the three-phase currents passed into the primary winding are exactly the same and symmetrical sinusoidal, namely three-phase current amplitudes are equal and the phase difference is 120° , the motor can generate an ideal thrust. In fact, the output current obtained through the inverter contains some high harmonics. Besides, due to the errors in machining and assembly, there is a deviation between actual primary winding spatial distribution and the designed values, which will lead to asymmetrical three-phase currents. Therefore, there is always a certain current harmonics during the operation of the motor, which is the root cause of the force ripple, and this kind of disturbance is called ripple disturbance [28].

It is assumed that the linear motor is three sets, two poles and the whole distance, regardless of the cogging effect or end effect, and the secondary is infinitely long. Refer to [29], the thrust of permanent magnet linear motor is

$$F = \frac{(e_{ma}i_a + e_{mb}i_b + e_{mc}i_c)}{V} = \frac{3\pi}{2pw} [F_0 + F_6 \cos(6w) + F_{12} \cos(12w) + F_{18} \cos(18w) + \dots], \quad (1)$$

where w is angular frequency of the current (rad/s), i_a , i_b , and i_c are the three-phase currents passed into the primary winding (A), p is the pole pitch of the motor (m), F_0 is the motor average thrust, and $F_{6n}(n \geq 1)$ is the ripple amplitude of motor thrust.

From the above expression, it can be seen that there is always a certain current harmonics during the operation of the motor, which can be also called ripple disturbance. It

FIGURE 1: A schematic of the cogging force generation in linear motor.

depends on both the input current and relative position between the stator and the motor.

Secondly, the cogging force is analyzed briefly. Figure 1 shows the generation mechanism of cogging force.

- (1) When the mover tooth is completely above the stator magnet, as Figure 1(a) shows, the magnetic flux entering into the mover teeth can neutralize each other in the motion direction, and the mover is not disturbed by other forces
- (2) When the mover tooth moves above the gap of the magnetic steel, as shown in Figure 1(b), the components of the magnetic flux entering into the mover teeth in the moving direction cannot offset each other. The disturbance force generated by the magnetic field is called the cogging force

Due to the existence of iron core, the linear motor cogging effect exists at any position at any time, and it depends on the position rather than current.

Refer to [30], the cogging torque can be expressed as

$$\begin{cases} T_e = \frac{p^2 L_k \tau_c B_r^2}{2\pi\mu_0 ((1/L_m) + (1/L_g))} \sum_{k=a,b,c} E_{sk}(x), \\ E_{sk}(x) = \frac{d \left[\int_{-(1-K_0)\tau_c}^{(1-K_0)\tau_c} |A_k(x)| dx \right]}{dx}, \end{cases} \quad (2)$$

where p is the motor pole pairs, L_k is the length of iron core (m), τ_c is the average width of air gap in magnetic steel area (m), B_r is magnetic induction in radial direction (T), μ_0 is the magnetic permeability, L_m is the thickness of the magnetic steel (m), L_g is the air gap length (m), and K_0 is the notch factor.

From the above formula, it can be seen that the cogging force is only position dependable, which means that it will still exist without current. In addition, the expression of cogging torque is too complicated to use for compensating force ripple.

Similar to the mechanism of cogging force, the end effect also resulted from magnetic force caused by uneven magnetic flux, which is shown in Figure 2.

The linear motor core cannot be continuous or infinitely long, and it must have two side ends. Similarly, the magnetic field that enters the edge of moving iron core is deformed and it will generate a magnetic field force, making the core edge unbalanced, which is called the end effect.

FIGURE 2: A schematic of the end effect generation in linear motor.

According to the above principle, it can be found that the end effect is position dependable. Generally, it is believed that the moving iron core is long enough so that the force on both ends of the core is not affected by each other, which means that the coupling forces at both ends are not considered [31]. Therefore, it can be considered that the end forces are the sum of F_+ and F_- . Figure 3 shows the one-sided end force.

The relationship between F_+ and F_- is

$$F_-|_{z=z'} = -F_+|_{z=-z'}, \quad (3)$$

where z' is any position of permanent magnet.

Because the length of motor iron core is an integral of the pole pitch, the relationship of F_+ and F_- is

$$\begin{cases} F_-|_{z=z'} = -F_+|_{z=-(z'+\Delta)}, \\ \Delta = L + i\tau, \end{cases} \quad (4)$$

where τ is the pole pitch (m) and L is the length of iron core (m).

Expanding by Fourier series, the end force F_+ and F_- can be expressed as

$$\begin{cases} F_+ = F_0 + \sum_{n=1}^{\infty} F_{sn} \sin \frac{2n\pi}{\tau} z + \sum_{n=1}^{\infty} F_{cn} \cos \frac{2n\pi}{\tau} z, \\ F_- = -F_0 + \sum_{n=1}^{\infty} F_{sn} \sin \frac{2n\pi}{\tau} (z + \Delta) + \sum_{n=1}^{\infty} F_{cn} \cos \frac{2n\pi}{\tau} (z + \Delta). \end{cases} \quad (5)$$

It can be inferred from the above equation that the theoretical expression of end force is a periodic function that is only related to the relative position between motor primary and secondary, and the end effect exists at any moment of the system.

However, since the expression of end force is too complex and it cannot be directly applied to the control system, it is necessary to design other schemes to improve the control accuracy for the linear motor end effect problem.

From the above analysis, it is not hard to find that both the motor input current and the relative position between the stator and the motor can result in force ripple in the

linear motor system, whereas these factors are inherent in the linear motor system when the motor structure cannot be changed. And the expressions of ripple disturbance, cogging force, and end force are too complicated and they cannot be directly applied to the real-time control system. It leads to that it is necessary to design other control strategies to eliminate its impact on the system performance. Even so, the above deriving results are meaningful because they provide reference value for designing compensation strategy hereinafter.

3. Method of IMILC-RDOB

Based on previous analysis, the paper proposes a control strategy combining the inverse model iterative learning control (IMILC) and robust disturbance observer (RDOB) to compensate for the influence of linear motor force ripple. The structure diagram is shown in Figure 4.

Where P is the transfer function of control object, C is the transfer function of feedback controller, C_L is the transfer function of iteration learning controller, $\hat{Q} \cdot Q_\lambda \cdot P_n^{-1}$ is the transfer function of robust disturbance observer, $\hat{Q}$ is the transfer function of robust disturbance observer filter, $r \cdot y$ are the system input and output, respectively, and d is the system disturbance.

The robust disturbance observer can effectively alleviate the contradiction between the disturbance suppression ability and antinoise ability that exists in conventional disturbance observer. Its input signal is the real-time actual position, and it can compensate for nonperiodic and random disturbances that appear in the system. But in the system with a certain time delay, the compensation accuracy for large and low frequency disturbances is limited because it does not use the previous motion information. According to the discussion in Section 1, the iterative learning compensation strategy has a strong ability to compensate for periodic force ripple. As a result, the compensation structure proposed in the paper can not only handle the aperiodic and random disturbances flexibly but also accurately compensate large and low frequency disturbances. In the following, each part of the compensation method is described in detail. And at the end of this section, the compensation method proposed in the paper is simulated and analyzed to verify the validity and superiority.

3.1. Design of Inverse Model Iterative Learning Controller.

In order to achieve the faster convergence speed with simpler control structure, the inverse model iterative learning controller (IMILC) is selected. Refer to patent [32], if the ILC law is the model inverse of feedback control closed-loop system, it can achieve one step convergence theoretically, namely that $C_L = L$ and L is

$$L = \left(\frac{CP}{1 + CP} \right)^{-1}. \quad (6)$$

In fact, it is not physically possible to invert the feedback control closed-loop system model. And the system noise may

FIGURE 3: A schematic of end force F_+ and F_- per edge.

FIGURE 4: The structure of IMILC-RDOB.

FIGURE 5: The structure of inverse model iterative learning law.

result in instability. Therefore, it is improved to make the iterative learning law be expressed as

$$C_L = Q_L \cdot \bar{Q}_L \cdot L. \quad (7)$$

The role of $\bar{Q}_L$ is make the inverse model of feedback control system achievable. According to the control object and feedback controller in the paper, a second-order filter is selected and its expression is

$$\bar{Q}_L(s) = \frac{1}{T_L^2 s^2 + 2T_L \xi_L s + 1}, \quad (8)$$

while Q_L is a low-pass filter and its role is mainly to weaken the unmodeled dynamics of the controlled object and maintain the stability in the whole frequency band.

As a result, a first-order low-pass filter is chosen and its transfer function is

$$Q_L(s) = \frac{1}{\lambda_L s + 1}. \quad (9)$$

Since the nonrepetitive interference and measurement noise, a smaller learning gain K_{ILC} is introduced to balance these adverse effects. Reducing the learning rate can also lead to decrease in learning efficiency. Therefore, it is necessary to increase iterative times to achieve ideal tracking precision. The final inverse model iterative learning law is shown in Figure 5. And its expression is

$$C_L = K_{ILC} Q_L \cdot \bar{Q}_L \cdot L. \quad (10)$$

3.2. Design of Robust Disturbance Observer. The conventional disturbance observer is modified in this section to

FIGURE 6: The structure of control system with conventional disturbance observer.

FIGURE 7: Bode diagram of $(1 - Q)$ when ξ_q varies.

improve system performance while avoiding stimulating system uncertainty. The structure of conventional disturbance observer (DOB) is simple, which only requires a low-pass filter and an inverse model of the nominal model [33]. Its structure is shown in Figure 6. The introduction of filter Q is to handle the problem that conventional disturbance observer need the model inverse; meanwhile, it can filter out some measurement noise.

P_n is the system nominal model and $P_n = K_m \cdot P$. So according to the system diagram, we have

$$u_v = d + u'' \cdot K_m = (1 - Q)d + K_m \cdot u'. \quad (11)$$

Considering that the filter must be matched with the order of system inverse model and the order should be as

low as possible, the second-order low-pass filter is selected. Its transfer function is

$$Q(s) = \frac{1}{T_q^2 s^2 + 2T_q \xi_q s + 1}. \quad (12)$$

The filter break frequency is set as 60 Hz. Analyzing the frequency characteristics of $(1 - Q)$, its transfer function is

$$1 - Q(s) = \frac{2T_q \xi_q s ((T_q/2\xi_q)s + 1)}{T_q^2 s^2 + 2T_q \xi_q s + 1}. \quad (13)$$

Figure 7 shows the Bode diagram of $(1 - Q)$ when ξ_q varies.

FIGURE 8: Bode diagram of $(1 - \hat{Q})$.

Theoretically, the smaller the value of ξ_q is, the stronger the suppression ability is and the higher the system tracking precision will be. Whereas, there is peak more than 0 dB in the frequency characteristics of $(1 - Q)$ when the value of ξ_q is too small, which will amplify the mechanical resonance or plant uncertainty at the peak corresponding frequency band. As a result, it is necessary to select appropriate filter parameters.

Considering the above requirements, a notch filter is introduced to improve the $(1 - Q)$ in conventional disturbance observer, which can not only retain its strong disturbance suppression ability when ξ_q is small but also enhance the system stability.

The transfer function of improved filter is $\hat{Q}$, and there is

$$(1 - \hat{Q}) = (1 - Q)G_{NF}. \quad (14)$$

The transfer function of notch filter G_{NF} is

$$G_{NF}(s) = \frac{T_{NF1}^2 s^2 + 2T_{NF1}\xi_{NF1}s + 1}{T_{NF2}^2 s^2 + 2T_{NF2}\xi_{NF2}s + 1}. \quad (15)$$

First, the parameters of notch filter are designed, so that it can make $(1 - \hat{Q})$ keep better disturbance attenuation performance when its frequency characteristics is under 0 dB. Meanwhile, make sure that there is no peak above 0 dB. Based on the above consideration, it is selected that $T_{NF1} = T_{NF2} = T_q$, $\xi_{NF1} = \xi_q$ and makes the value of ξ_{NF2} be much

larger than ξ_q , which can balance out the peak above 0 dB. So the parameters are set as $T_{NF1} = T_{NF2} = T_q = 1/2\pi \times 60$, $\xi_{NF1} = \xi_q$, and $\xi_{NF2} = 5$ briefly. The Bode diagram of $(1 - \hat{Q})$ is shown in Figure 8.

From Figure 8, it can be found that there is harmonic peak of 14 dB in the vicinity of 60 Hz. It will depress the peak in the mid-high frequency band and retains the ability of disturbance attenuation for the low frequency band, which can make sure not to motivate the plant uncertainty.

Therefore, the parameters are set as $T_{NF1} = T_{NF2} = T_q$ and $\xi_{NF1} = \xi_q$ in the following applications. Then, the modified filter's transfer function is

$$\hat{Q}(s) = 1 - (1 - Q(s)) \cdot G_{NF}(s) = \frac{2T_q(\xi_{NF2} - \xi_q)s + 1}{T_q^2 + 2T_q\xi_{NF2}s + 1}. \quad (16)$$

The structure diagram of improved disturbance observer, namely the robust disturbance observer, is shown in Figure 9.

Among them, the filter Q_λ is to make the inverse model of system nominal model realizable. Usually, the inverse model is $P_n^{-1} = ms^2$ and then the part of $\hat{Q} \cdot P_n^{-1}$ cannot be physically realized. So the filter Q_λ is introduced to solve this problem. About the problem of type selection for the filter Q_λ , there are three main considerations.

- (1) The main function of the filter Q_λ is to make up for the order of robust disturbance observer, which make sure the robust disturbance observer be physically

FIGURE 9: The structure of control system with robust disturbance observer.

realized. In consequence, the order of filter should be as lower as possible

- (2) Because the disturbance observer input is actual position, noise is bound to be introduced into the system. The filter Q_λ should have the ability to suppress high-frequency noise
- (3) Due to the large amount of uncertainty in the high frequency band, the filter Q_λ should avoid that the plant uncertainty is motivated in the observer loop

Based on the above considerations, a first-order low-pass filter is selected as the filter Q_λ and its expression is

$$Q_\lambda(s) = \frac{1}{\lambda s + 1}. \quad (17)$$

3.3. Convergence Analysis. Convergence analysis is the core problem of the iterative learning control, and this section will provide convergence analysis of the above control structure. According to Figure 4, it can be found that the system output can be expressed as

$$y = \frac{K_m CP}{\hat{Q}Q_\lambda - \hat{Q} + 1 + K_m CP} (r + \bar{e}) + \frac{P(\hat{Q} - 1)}{\hat{Q}Q_\lambda - \hat{Q} + 1 + K_m CP} d, \quad (18)$$

where $\bar{e}$ is the output of iterative learning controller.

Therefore, the system error after $(k + 1)$ th iterations is

$$\begin{aligned} e_{k+1} &= r - y_{k+1} = r - \left(\frac{K_m CP}{\hat{Q}Q_\lambda - \hat{Q} + 1 + K_m CP} (r + \bar{e}_{k+1}) \right. \\ &\quad \left. + \frac{P(\hat{Q} - 1)}{\hat{Q}Q_\lambda - \hat{Q} + 1 + K_m CP} d \right) \\ &= \left(1 - \frac{K_m CP}{A} \right) r - \frac{B}{A} d - \frac{K_m CP}{A} \bar{e}_{k+1}, \end{aligned} \quad (19)$$

$$A = \hat{Q}Q_\lambda - \hat{Q} + 1 + K_m CP,$$

$$B = P(\hat{Q} - 1).$$

According to Figure 5, we have

$$\begin{aligned} \bar{e}_{k+1} &= \bar{e}_k + e_k \cdot C_L, \\ \bar{e}_1 &= 0. \end{aligned} \quad (20)$$

Therefore, e_{k+1} can also be expressed as

$$\begin{aligned} e_{k+1} &= G \left(1 - \frac{K_m CP}{A} \right) r - G \frac{B}{A} d - G \frac{K_m CP}{A} \bar{e}_k, \\ G &= 1 - \frac{K_m CPC_L}{A}. \end{aligned} \quad (21)$$

Therefore, it can be speculated that e_{k+1} is

$$\begin{aligned} e_{k+1} &= G^k \left(1 - \frac{K_m CP}{A} \right) r - G^k \frac{B}{A} d - G^k \frac{K_m CP}{A} \bar{e}_1 \\ &= G^k \left(1 - \frac{K_m CP}{A} \right) r - G^k \frac{B}{A} d. \end{aligned} \quad (22)$$

The above result is only a corollary, so mathematical induction is used to prove that the expression (22) is right.

When $k = 0$, the system error is

$$\begin{aligned} e_1 &= r - y_1 = r - \left(\frac{K_m CP}{A} (r + \bar{e}_1) + \frac{B}{A} d \right) \\ &= \left(1 - \frac{K_m CP}{A} \right) r - \frac{B}{A} d. \end{aligned} \quad (23)$$

So when $k = 0$, the expression (22) is right.

Next, k is set as $(m - 1)$, and according to (22) e_m is

$$e_m = G^{m-1} \left(1 - \frac{K_m CP}{A} \right) r - G^{m-1} \frac{B}{A} d. \quad (24)$$

While according to the characteristics of system, e_{m+1} can be calculated.

$$\begin{aligned}
e_{m+1} - e_m &= (r - y_{m+1}) - (r - y_m) = y_m - y_{m+1} \\
&= \left(\frac{K_m \text{CP}}{A} (r + \bar{e}_m) + \frac{B}{A} d \right) \\
&\quad - \left(\frac{K_m \text{CP}}{A} (r + \bar{e}_{m+1}) + \frac{B}{A} d \right) \\
&= \frac{K_m \text{CP}}{A} (\bar{e}_m - \bar{e}_{m+1}) \\
&= \frac{K_m \text{CP}}{A} (\bar{e}_m - (\bar{e}_m + e_m \cdot C_L)) \\
&= \frac{K_m \text{CPC}_L}{A} e_m.
\end{aligned} \tag{25}$$

Then, e_{m+1} can be expressed as

$$e_{m+1} = \left(1 - \frac{K_m \text{CPC}_L}{A} \right) e_m = G \cdot e_m. \tag{26}$$

Substituting (24) into (26), it follows that

$$\begin{aligned}
e_{m+1} &= G \cdot \left(G^{m-1} \left(1 - \frac{K_m \text{CP}}{A} \right) r - G^{m-1} \frac{B}{A} d \right) \\
&= G^m \left(1 - \frac{K_m \text{CP}}{A} \right) r - G^m \frac{B}{A} d.
\end{aligned} \tag{27}$$

When $k = m$, (22) is equal to (27). By mathematical induction, the expression (22) is right. And the convergence condition is expressed as

$$\left| 1 - \frac{K_m \text{CPC}_L}{A} \right| < 1. \tag{28}$$

According to parameter setting in simulation and experiments, the Bode diagram of convergence condition can be get as shown in Figure 10.

From Figure 10, it can be seen that in the 0~160 Hz frequency band, the convergence of proposed method can be guaranteed. When in more than 160 Hz frequency band, the left side of (28) will be a little bigger than 1, which have little effect on system convergence. And the force ripple mainly exists in the 0~120 Hz frequency band in the actual system. Therefore, the convergence of proposed algorithm in the frequency band that we cared about can be guaranteed. In summary, the proposed method can satisfy the convergence condition.

3.4. Simulation of IMILC-RDOB. According to Figure 9, the simulation model is built up. The controlled object uses the linear motor nominal model, namely $P(s) = 1/ms^2$. So the position controller is designed as PI-lead correction. The parameters of position controller are set as frequency width $h = 100$, cutting frequency $\omega_c = 60\text{Hz}$, and PI-type controller's break frequency $\omega_l = 20\text{Hz}$. The parameters of inverse model iterative learning controller are $T_L = 1/2\pi \times 1000$, $\xi_L = 0.7$, $\lambda_L = 1/2\pi \times 60$, and $K_{\text{ILC}} = 0.7$. The parameters of robust disturbance observer are $T_q = T_{\text{NF1}} = T_{\text{NF2}} = 1/2\pi \times 60$, $\xi_q = \xi_{\text{NF1}} = 0.1$, $\xi_{\text{NF2}} = 5$, and $\lambda = 1/2\pi \times 200$.

The system input is a third-order S curve, and its parameters are set as follows: the jerk is 500 m/s^3 , the acceleration is 8 m/s^2 , the velocity is 0.3 m/s , and the displacement is 0.2 m . System disturbance is replaced by the sum of a set of sine functions at different frequency, which amplitude is 16 and frequency includes 40 Hz, 60 Hz, 80 Hz, and 164 Hz, respectively.

In addition, since there are lots of uncertainties beyond 100 Hz, the components that simulate the uncertainty must be added in the simulation in order to simulate the actual situation as much as possible. In consequence, a first-order resonance is introduced to simulate the real plant uncertainty and its transfer function is

$$G_{\text{res}}(s) = \frac{T_{\text{res1}}^2 s^2 + 2T_{\text{res1}} \xi_{\text{res1}} s + 1}{T_{\text{res2}}^2 s^2 + 2T_{\text{res2}} \xi_{\text{res2}} s + 1}, \tag{29}$$

where damping ratio is $\xi_{\text{res1}} = \xi_{\text{res2}} = 0.01$, and the break frequency are $T_{\text{res1}} = 1/2\pi \times 120$ and $T_{\text{res2}} = 1/2\pi \times 160$, respectively.

The simulation results of the proposed compensation method are shown in Figure 11. And the comparison results of the proposed compensation method with only ILC and only disturbance observer are shown in Figure 12. Table 1 shows the maximum tracking error in the uniform velocity section under three compensation methods.

Considering the situation that system tracks the reference signal in the uniform velocity section emphatically, the following conclusions can be drawn.

- (1) From Figure 11, it can be seen that the simulation results verify the effectiveness of the IMILC-RDOB compensation method
- (2) From Figure 12, the conclusion can be drawn that among the three compensation methods, the IMILC-RDOB has the strongest disturbance compensation capability and the simulation results validate its superiority
- (3) After the same 7 iterations, the tracking precision obtained by IMILC-RDOB is much higher than that of the single application of IMILC, which indirectly verified that the proposed method has faster iterative speed

The above conclusions all indicate that the proposed compensation method has better disturbance suppression ability and has stronger robustness. And specific experimental verification will be carried out in the actual platform.

4. Experiment

4.1. Introduction of Experimental Platform

4.1.1. Overview of Experimental Platform. In this paper, the long-stroke stage in the double wafer stage system of lithography machine is used as the platform to verify performance of the proposed compensation structure, which is shown in

FIGURE 10: The Bode diagram of convergence condition expression.

FIGURE 11: The simulation results of IMILC-RDOB.

Figure 13. The linear motor in the red dotted line box is the motor used in the experiment.

The positive direction of X and Y in space is marked, where the zero point of X-direction is at the center of the guide rail. The linear encoder LIF471, with the measurement resolution of 50 nm, is used as the position sensor to provide feedback signals to the closed-loop system. The control output is computed by a motion control card according to the

feedback from linear encoder, then it is transmitted to the motor driver through the optical fiber. In this experiment, the sampling period is $200 \mu\text{s}$. The control cycle flow chart is shown in Figure 14.

In the actual operating conditions, the motion of X-direction motor is to track the step trajectory formed by multiple S curves, whose parameters are set as same as the above simulation.

FIGURE 12: Comparison of the three compensation methods.

TABLE 1: Maximum tracking error in the uniform velocity section under three compensation methods.

Method	IMILC-RDOB	IMILC	DOB
Max error (μm)	0.9520	3.5866	115.0290

Besides, to analyze the tracking performance, two indexes commonly used in the lithography machine should be defined, which are MA and MSD, respectively [34]. Their expressions are

$$\text{MA}(t) = \frac{1}{T_e} \int_{t-T_e/2}^{t+T_e/2} e(\tau) d\tau, \quad (30)$$

$$\text{MSD}(t) = \sqrt{\frac{1}{T_e} \int_{t-T_e/2}^{t+T_e/2} (e(\tau) - \text{MA}(\tau))^2 d\tau}, \quad (31)$$

where T_e is the exposure time length (s) with expression $T_e = s/v$, s is the slit width (m), and v is the scanning speed (m/s). In this experiment, the slit width is $s = 10\text{mm}$ and the scanning speed is $v = 0.3\text{m/s}$. From (30) and (31), it can be seen that MA represents the uniformity of the motion error, and MSD characterizes the accuracy rating of the motion error.

The nominal values of X-direction linear motor parameters are shown in Table 2.

4.1.2. Experimental Parameter Setting. For position loop controller, the plant model contains not only the linear motor but also the driver. To obtain a model with enough accuracy, the frequency sweep experiment is designed and the result is shown in Figure 15.

It should be noted that the signal noise ratio of high frequency band is relatively low; the obtained actual plant frequency characteristics have been covered by noise. As a result, only the data in low and middle frequency band of the frequency sweep experiment are available, shown as second-order inertial component. This result could be analyzed as follows: First, the low signal noise ratio at low frequency band cannot fully reflect the characteristics of system. Next, because the control output in low frequency band is small, the motor control is affected greatly by the force ripple, as well as cable tension and air-film friction. Due to the greatly changed position, the system characteristics in low frequency band are mainly expressed by the force ripple.

Synthesizing the above analysis, the data of middle frequency band are selected to build up the plant model. Fitting with double integral link model, its transfer function is

$$G_{\text{rigid}}(s) = \frac{1}{ms^2}, \quad (32)$$

where m is the entire mass of motor mover. Since the driving magnification is also taken into account in the experiment, in fact, its value is $m = 20 \times (\text{mover mass})$, the more smooth and accurate frequency points are selected to calculate its value and we have $m = 529.5177$.

Using classical automatic control theory, a reasonable PI-lead correction controller is designed for the above motor model, and its transfer function is

$$C(s) = \frac{1.9962 \times 10^5 s^2 + 3.2611 \times 10^7 s + 9.4570 \times 10^9}{2.6526 \times 10^{-4} s^2 + s}. \quad (33)$$

FIGURE 13: The overview of the experimental platform.

FIGURE 14: The control cycle flow chart.

TABLE 2: The parameters of X-direction linear motor.

Parameters (25°C)	Symbols	Units	Values
Phase resistance	R_h	Ω	2.41
Phase inductance	L_h	mH	18
Thrust coefficient	K_{ht}	N/A	93
Phase back-EMF coefficient	K_{he}	V/m/s	76
Peak current	$I_{h \max}$	A	15
Peak thrust	$F_{h \max}$	N	1200
Pole pitch	τ	mm	24

Theoretically, the closed-loop bandwidth of control system can reach 91.3 Hz, and the actual closed-loop bandwidth obtained by frequency sweep experiment is approximately 81.37 Hz. The reason why the actual closed-loop bandwidth

is slightly lower than the theoretical value mainly resulted from the large number of mechanical resonance and uncertainties, which lie nearby 100 Hz. In consequence, the PI-lead correction controller with the above parameters will be adopted in the future work.

The theoretical formula of force ripple is usually too complicated to derive available model. As a result, in order to reasonably select the break frequency of robust disturbance observer's filter, the force ripple data must be obtained through experiments. And then, the frequency analysis should be performed. On the basis of the above requirements and the existing experimental facilities, the data acquisition experiment is designed.

Since the cogging force is only related to motor position, the uniform velocity method for force measurement is designed, shown in Figure 16. The principle of the method is to obtain the controller output at each sampling point under uniform velocity, which is the opposite number of the cogging force at the corresponding position. Because

FIGURE 15: The actual plant frequency characteristic and the fitting curve of double integral model.

FIGURE 16: A schematic diagram of the cogging force measurement method.

FIGURE 17: The cogging force varies with the position.

TABLE 3: The statistics of cogging force data.

Data	Minimum	Maximum	Average	Variance
Input (m)	-0.1000	0.1000	0	0.0033
Output (N)	-7.6685	16.1804	3.9452	27.8808

the motor is air-floating, there is no external disturbance theoretically. So the controller output under the uniform velocity is the opposite of the cogging force.

The cogging force with the position -0.1 m to $+0.1$ m is shown in Figure 17. The statistical values of the cogging force data are shown in Table 3. The velocity of motor is 0.025 m/s in the data acquisition experiment.

FFT analysis is performed on the cogging force data and the frequency characteristics diagram is shown in Figure 18.

The fundamental frequency of the force ripple is approximately $f = v/\tau = 0.025\text{m/s}/0.012\text{m} = 2.0833\text{Hz}$, where τ is the motor pole pitch. According to the cogging force frequency characteristics, it can be seen that it includes fundamental frequency and harmonics with order of 2, 3, and 6. Among them, the peak of fundamental frequency is larger, so the filter break frequency should be at least twice of this frequency. For the S curve which uniform velocity is 0.3 m/s, the fundamental frequency in the force ripple is about 25 Hz. Combining with the situation that the closed-loop bandwidth under the designed PI-lead correction controller is 81.37 Hz, the break frequency of the low-pass filter in the robust disturbance observer is selected as $T_q = 1/2\pi \times 60$. Thus, the frequency components of the force ripple within the bandwidth can be filtered out.

4.2. Experiments. In this section, the experiments of the above compensation method will be performed. The parameters of the robust disturbance observer are $T_q = T_{\text{NF1}} = T_{\text{NF2}} = 1/2\pi \times 60$, $\xi_q = \xi_{\text{NF1}} = 0.01$, $\xi_{\text{NF2}} = 5$, and $\lambda = 1/2\pi \times 200$, respectively. The parameters of the inverse model iterative learning feedforward controller are $T_L = 1/2\pi \times 1000$, $\xi_L = 0.7$, $\lambda_L = 1/2\pi \times 60$, and $K_{\text{ILC}} = 0.7$. And other conditions are the same as the simulation's conditions. The tracking errors are shown in Figure 19 and the k th maximum tracking errors in the uniform velocity section are shown in Table 4.

It can be seen from the diagram that the system tracking performance is improved greatly after the introduction of iterative learning feedforward, and the effectiveness of the method has been verified. In the same case, the comparison among IMILC-RDOB, IMILC, and DOB is conducted. Tracking error comparison diagram of the three compensation methods is shown in Figure 20. FFT analysis of tracking error of the three compensation methods is shown in Figure 21. MA and MSD comparison diagrams of the three compensation methods are shown in Figure 22. The iterative speed comparison diagram between IMILC-RDOB and IMILC is shown in Figure 23. Various indexes of uniform velocity section of the three compensation methods are shown in Table 5. It is worth noting that, according to the reason mentioned in the previous chapter, when the damping ratio of conventional disturbance observer is selected to be smaller, the characteristics will arouse the resonance

oscillation in the system. Therefore, in the experiments, the damping ratio of conventional disturbance observer is selected as $\xi_{q\text{DOB}} = 0.5$.

The above results are analyzed and the conclusions can be drawn.

- (1) Firstly, according to Figure 20 and experimental data from Table 5, the maximum tracking error of uniform velocity section is analyzed. The IMILC-RDOB compensation method can improve the system tracking precision and reduce the times of iterations. After 10 iterations, the tracking precision using the IMILC-RDOB method can reach $0.30 \mu\text{m}$, which is only 6 times of the sensor resolution. Compared with only using iterative learning compensation, it increases by 33.33%. And it increases by 95.56% compared with disturbance observer
- (2) Secondly, the maximum MA of uniform velocity section is analyzed. According to experimental data from Table 5, after 10 iterations, the index of MA using IMILC-RDOB can reach $0.0265 \mu\text{m}$, which increases by 89.32% compared with iterative learning compensation and by 99.19% compared with disturbance observer
- (3) Thirdly, the maximum MSD of uniform velocity section is analyzed. It can be seen from Figure 23 and Table 5 that, after 10 iterations, the index of MSD using IMILC-RDOB can reach $0.1497 \mu\text{m}$, which increases by 39.95% compared with iterative learning compensation and by 95.28% compared with disturbance observer
- (4) Next, the settling time is considered. After 10 iterations, the settling time using the IMILC-RDOB method is 10.4 ms, which increases 82.31% compared with iterative learning compensation. Although the conventional disturbance observer seems to have a shorter settling time, it has no real value because of its large tracking error
- (5) Finally, Figure 23 and experimental data show that the IMILC-RDOB compensation method has higher convergence speed and convergence accuracy than the inverse model iterative learning compensation method. The proposed method can effectively compensate the force ripple in a certain frequency band and ensure that the system resonance not be motivated

The experimental results can fully verify the effectiveness of the IMILC-RDOB compensation method. At the same time, the comparison with DOB and IMILC proved that the proposed method has extreme superiority in terms of convergence speed, tracking precision, settling time, MA, and MSD.

5. Conclusion

Since the force ripple of linear motor presents periodicity in space and there are also nonperiodic disturbance in time,

FIGURE 18: The cogging force frequency characteristic.

FIGURE 19: The tracking error curve using IMILC-RDOB.

TABLE 4: The maximum tracking error using IMILC-RDOB.

k	1	2	3	4	5
Error (μm)	1.40	1.10	0.75	0.65	0.50
k	6	7	8	9	10
Error (μm)	0.50	0.40	0.35	0.40	0.30

it is impossible to obtain high control precision and stability by ILC only. The conventional disturbance observer is limited by the system structure in practical application so that it cannot compensate the influence of force ripple either. In view of their characteristics, the conventional disturbance observer was modified and combined with the inverse model

FIGURE 20: Tracking error comparison of the three compensation methods.

FIGURE 21: FFT analysis of tracking error of the three compensation methods.

iterative learning feedforward method. Based on that, the paper proposed a novel control strategy for force ripple suppression. The effectiveness of the proposed method is validated by simulation and experiments, respectively. The results show that the proposed method can shorten the iteration period and settling time while improving the system tracking precision, reflecting the practical value of the proposed method. However, there are also some shortcomings in the paper. It is known from the experimental parameter

setting that the robust disturbance observer cannot handle third and above harmonic components of force ripple. Because of lack of inhibiting ability for middle- and high-frequency force ripple components, the proposed robust disturbance observer cannot compensate force ripple completely. Therefore, for solving the above problem, there are still lots of work to be done in the future. (The modeling data used to support the findings of this study are available from the corresponding author upon request.)

FIGURE 22: Comparison of MA and MSD by the three compensation methods.

FIGURE 23: Iterative speed comparison between IMILC-RDOB and IMILC.

TABLE 5: Performance in uniform velocity section of the three compensation methods.

Methods Indexes	IMILC-RDOB	IMILC	DOB
Maximum tracking error (μm)	0.30	0.45	6.75
MA (μm)	0.0265	0.2482	3.2891
MSD (μm)	0.1497	0.2493	3.1724
Settling time (ms)	10.4	58.8	0.2

Data Availability

The data used to support the findings of this study are available from the corresponding author upon request.

Conflicts of Interest

The authors declare that they have no conflicts of interest.

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