

Article

A Pole-Placement-Based Variable-Gain Observer for Precision Motion Stages: Addressing the Disturbance-Noise Trade-Off

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Abstract

Accurate pattern transfer in the lithography process demands extreme positioning accuracy. However, various external disturbances acting on the wafer stage can lead to positioning errors. To address this issue, this paper proposes a pole-placement-based Variable-Gain Extended State Observer (VGESO). First, the trade-off between disturbance rejection and noise attenuation faced by conventional Extended State Observers (ESOs) in precision motion systems is analyzed. Then, a modified ESO structure is introduced, in which two pole-related parameters are employed to adaptively adjust the observer gains. These parameters enable effective suppression of both low-frequency disturbances and high-frequency measurement noise within their designated ranges. Finally, simulation results verify the effectiveness and superior performance of the proposed method.

Keywords: wafer stage; variable-gain observer; motion control; precision positioning

1. Introduction

Integrated Circuits (ICs) are the foundation of modern information technology. To enhance processor performance, an increasing number of sophisticated transistors are being integrated onto a single chip. The number of transistors per unit area continues to double approximately every 18 to 24 months, according to Moore's law [1,2]. Over the past 70 years, the flourishing development of IC technology has been inseparable from continuous advancements in lithography processes. Lithography machines are the core equipment in IC manufacturing, responsible for transferring the patterns on the mask to a wafer coated with a photoresist through an optical system [3]. To achieve precise pattern transfer, the wafer stage, which carries the wafer, and the mask plate, which holds the pattern, must maintain nanometer-level motion accuracy during the scanning process [4,5]. Meanwhile, to enhance yield, the motion stage of a lithography machine must achieve nanometer-level tracking precision along high-speed, high-acceleration, and large-stroke exposure trajectories [6].

Addressing this problem mainly involves the following two challenges. Firstly, the motion stage of a lithography machine is a complex mechatronic system, for which it is difficult to establish accurate mathematical models for controller design. Both the wafer stage and the mask stage are six-degree-of-freedom (6-DOF) systems with strong coupling among different DOFs, and their dynamic characteristics vary with the motion of the stage [7,8]. Moreover, the lightweight design of the platform increases structural



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flexibility, which further exacerbates the mismatch between the nominal model and the actual system [9,10]. Secondly, the servo control process of a lithography machine is subject to various disturbances. The motion stage actuator is affected by position- and velocity-dependent magnetic disturbance forces [11]. In addition, the distribution of the coils introduces cable forces [12]. The system is further influenced by measurement noise, air-film-induced vibrations, and structural vibrations caused by reaction forces during stage motion [13]. Consequently, effectively mitigating the effects of disturbances and modeling uncertainties is crucial for improving the motion control performance of lithography machines.

Based on common motion control system architectures, control methods for disturbance rejection can generally be classified into feedforward and feedback schemes [14]. The feedforward approach constructs an inverse model of the platform based on a pre-identified system model and employs this inverse model as a feedforward controller to compensate for disturbances. However, as discussed previously, the platform model cannot be obtained with sufficient accuracy, and therefore feedforward control alone cannot achieve ideal disturbance rejection performance. Feedback control schemes can be further categorized into three classes according to the extent of available model information. The first class consists of model-free methods, such as proportional–integral–derivative (PID) control [15]. Although PID control is widely adopted due to its simplicity and ease of implementation, conventional PID controllers have difficulty meeting the stringent requirements of nanometer-level motion accuracy. The second class includes model-based methods that rely heavily on accurate system models, such as disturbance observers (DOBs) [16] and model-dependent adaptive control strategies [17]. The third class comprises methods that exploit partial model information, including Active Disturbance Rejection Control (ADRC) [18]. These approaches design controllers by leveraging the available system knowledge, and in general, greater model information leads to improved control performance. Compared with the first class, the third class makes more effective use of available model information. Moreover, it can achieve performance comparable to that of the second class while requiring less detailed model knowledge. Consequently, this category offers a favorable trade-off between control performance and practical applicability.

ADRC exploits available system information to construct an extended state observer (ESO), which estimates the total disturbance by lumping external disturbances and internal uncertainties into an augmented state [19]. This formulation enables simultaneous compensation of both disturbances and uncertainties. Consequently, the accuracy and timeliness of total disturbance estimation serve as key criteria for evaluating ESO performance. To achieve fast and accurate disturbance estimation, the ESO is typically designed with high observer gains. However, high-gain designs inevitably amplify measurement noise, which can significantly degrade the servo performance of the motion stage and may even lead to instability. As a result, the gain selection of conventional ESOs must carefully balance disturbance rejection capability against sensitivity to measurement noise [20].

In general, two main strategies have been developed to address this issue. One effective strategy is to employ cascaded ESO architectures, in which disturbances are sequentially estimated and compensated at multiple layers. Several cascaded ESO structures have been reported in [21,22], where multiple observers with different bandwidths are designed to improve overall estimation accuracy. Another variant adopts a dual-ESO configuration, in which the observation residual of the first ESO is further estimated and compensated by a second ESO [23]. In addition, a variable-structure ESO was proposed in [24], where a cascaded-filter-based linear ADRC(LADRC) scheme was developed to enhance disturbance rejection performance. Furthermore, an additional tuning parameter was introduced to optimize the ESO gains, enabling a dual trade-off between disturbance rejection

tion and measurement noise suppression [20]. The second strategy focuses on restructuring the conventional ESO framework to decouple control performance from strict observer bandwidth limitations. One representative approach introduces a phase-lead element to improve disturbance observation capability; however, the influence of measurement noise is not explicitly addressed [25]. An improved adaptive linear active disturbance rejection control method was proposed in [26], while an adaptive optimization algorithm was developed in [27]. Moreover, a fractional-order ESO combined with a backstepping-based ADRC scheme was presented in [28]. These methods increase the degrees of freedom in ESO design, thereby reshaping the inherent coupling between disturbance rejection and noise attenuation and enabling more flexible and tailored control performance.

However, cascaded ESO architectures introduce multiple observer layers, which significantly increase the system order and the complexity of parameter tuning. This not only makes the controller design more cumbersome but also complicates stability analysis. Moreover, properly coordinating observer bandwidths across different layers is nontrivial, and inappropriate tuning may lead to phase lag, oscillations, or degraded disturbance rejection performance. In addition, the underlying coordination mechanism between disturbance rejection and measurement noise suppression has not yet been fully clarified. To address these issues, this paper redesigns the ESO parameters from a pole-placement perspective. The main contributions of this paper are summarized as follows:

1. A detailed analysis is conducted to reveal the inherent trade-off between low-frequency disturbance rejection and high-frequency measurement noise amplification in conventional ESOs. Based on this analysis, a corresponding solution strategy is developed.
2. A pole-placement-based variable-gain extended state observer (VGESO) is proposed. By directly adjusting the observer pole locations to determine the ESO parameters, the proposed approach achieves improved disturbance rejection and noise attenuation compared with conventional ESO and existing VGESO schemes. Moreover, a systematic and intuitive tuning strategy is presented. Despite the introduction of additional design parameters, the proposed method does not increase tuning complexity, thereby maintaining practical applicability.
3. The effectiveness of the proposed approach is validated through simulation studies on a motion stage system, demonstrating its capability to simultaneously suppress low-frequency disturbances and high-frequency noise.

The remainder of this paper is organized as follows. Section 2 presents the single-degree-of-freedom modeling of the lithography motion stage and formulates the problem under investigation. Section 3 analyzes the limitations of the conventional ESO and introduces the proposed method. Section 4 provides the theoretical stability analysis of the proposed approach. Section 5 presents comparative simulation results to evaluate the performance of the proposed method against existing approaches. Finally, Section 6 concludes the paper.

2. Problem Formulation

2.1. Wafer Stage Dynamics

During the lithography process, the wafer stage is required to perform six-degree-of-freedom (DOF) motion, with coupling existing among different DOFs. To facilitate controller design, force decoupling and measurement decoupling are applied to transform the system into six independent single-degree-of-freedom (SDOF) subsystems. Each SDOF

subsystem consists of one rigid-body mode and several flexible resonant modes, which can be described as follows:

$$P(s) = \frac{1}{ms^2} + \frac{1}{m} \sum_{i=1}^n \frac{k_i}{s^2 + 2\zeta_i\omega_i s + \omega_i^2} \quad (1)$$

where m denotes the total mass of the platform, ζ_i and ω_i represent the damping ratio and resonant frequency of the i th resonant mode, respectively, and k_i is the modal parameter determined by the system eigenmodes and the locations of the input and output. s represents the Laplace operator.

In practice, the above parameters cannot be obtained with complete accuracy. The higher-order dynamics beyond the second-order term are treated as internal uncertainties or internal disturbances of the platform, which can be represented as follows:

$$M(s) = \frac{1}{ms^2} - \frac{1}{\hat{m}s^2} + \frac{1}{m} \sum_{i=1}^n \frac{k_i}{s^2 + 2\zeta_i\omega_i s + \omega_i^2} \quad (2)$$

where $\hat{m}$ denotes the nominal mass of the platform. It is evident that the internal uncertainties of the considered system include both mass variations and unmodeled higher-order dynamics. By incorporating these uncertainties into the internal disturbance term, the actual system model can be rewritten as

$$P(s) = \frac{1}{\hat{m}s^2} + M(s) \quad (3)$$

Considering that the platform is subject to external disturbances such as electromagnetic interference, cable forces, and friction, $d(t)$ denotes the lumped external disturbance acting on the system. Applying the inverse Laplace transform, the above system can be rewritten in the time domain as

$$\ddot{y}(t) = b_0 u(t) + f(t) \quad (4)$$

where $b_0 = \frac{1}{\hat{m}}$, and $y(t)$ and $u(t)$ denote the output and input of the platform, respectively. The term $f(t)$, referred to as the total disturbance in the ADRC framework, represents the combined effect of internal uncertainties and external disturbances. It can be expressed as $f(t) = \frac{1}{\hat{m}}d(t) + (M(t)u(t))''$, where $M(t)$ is the inverse Laplace transform of $M(s)$.

It should be noted that the disturbance term $d(t)$ represents a lumped equivalent disturbance acting on the system. Although electromagnetic interference, cable effects, friction, and modeling uncertainties originate from different physical sources and may have different physical units, their influence on the system dynamics can be equivalently mapped into the input channel through the plant model. Specifically, these effects manifest as additional forces or torques acting on the system, and can therefore be expressed in the same unit as the control input after being processed by the system dynamics. This lumped-disturbance formulation is widely adopted in ADRC- and ESO-based control frameworks, as it enables unified estimation and compensation of various uncertainties and external perturbations without requiring precise modeling of each individual disturbance source.

The schematic diagram of the single-degree-of-freedom motion control system is illustrated in Figure 1. The objective of the controlled stage is to accurately track the reference trajectory, which is designed according to the specifications of the lithography exposure process. The feedback controller is primarily responsible for ensuring system stability and tracking accuracy by compensating for model uncertainties and external disturbances through continuous correction based on the tracking error between the reference signal and the measured output. In contrast, the feedforward controller aims to improve tracking

performance and dynamic response by predicting the required control action from the reference signal or known system dynamics, thereby alleviating the burden on the feedback loop. Based on Figure 1, the following relationship can be derived:

$$\begin{cases} y(s) = G(s)[r(s) - \zeta(s)] + T(s)[d(s) + u_d(s)] \\ y_\zeta(s) = G(s)r(s) + T(s)[d(s) + u_d(s)] + S(s)\zeta(s) \\ e_\zeta(s) = S(s)r(s) - T(s)[d(s) + u_d(s)] - S(s)\zeta(s) \end{cases} \quad (5)$$

where $S(s) = [1 + C_{fb}(s)P(s)]^{-1}$ is the sensitivity function, $T(s) = P(s)S(s)$ is the process sensitivity function, and $G(s) = C_{fb}(s)P(s)S(s)$ is the complementary sensitivity function.

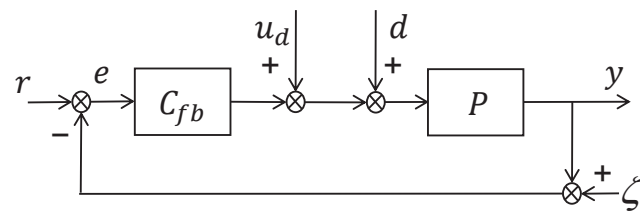


Figure 1. 1DOF control configuration.

2.2. Problem Description

During the operation of a lithography stage, extremely high positioning accuracy is required. As indicated by the above formulations, system errors mainly originate from the reference trajectory, external disturbances, and measurement noise. Errors induced by the reference trajectory can be effectively mitigated through trajectory planning and input shaping techniques, which are beyond the scope of this work. This study therefore focuses on suppressing the effects of disturbances. The effectiveness of the disturbance compensation input u_d directly determines the achievable positioning accuracy. If the condition $u_d = -d$ is satisfied, the disturbance can be completely eliminated. In practice, the compensation input u_d is generated by the ESO; however, it is inevitably affected by measurement noise. Accordingly, this work proceeds as follows: first, the influences of disturbances and measurement noise on the observation results and system output are analyzed; then, a pole-placement-based variable-gain extended state observer is proposed.

3. Control Design

ADRC is a widely used feedback control method that estimates and compensates for total disturbances by means of an ESO. Gao facilitated the practical implementation of ADRC by introducing the observer bandwidth concept, which enables a linearized and systematic tuning of ESO parameters [29]. In this section, ADRC is applied to the modeled motion stage to analyze the effects of total disturbances and measurement noise. Furthermore, primary and secondary observer poles are introduced to define two new design parameters, which enable simultaneous suppression of disturbances and measurement noise.

3.1. Design of ADRC

In Section 2, internal uncertainties and external disturbances are lumped into a total disturbance. The system block diagram of the ADRC framework is shown in Figure 2. This section focuses on the additional ADRC component for the integral-series standard form of the system. Let $y(t) = x(t) = [x_1(t) \ x_2(t) \ x_3(t)]^T$, $\dot{x}_3(t) = h(t)$. Accordingly, the system can be rewritten in the following state-space form:

$$\begin{cases} \dot{x}(t) = Ax(t) + Bu(t) + Eh(t) \\ y(t) = Cx(t) \\ y_{\zeta}(t) = Cx(t) + \zeta(t) \end{cases} \quad (6)$$

where $y_{\zeta}(t)$ is the output with measurement noise and $A = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ 0 & 0 & 0 \end{bmatrix}$, $B = \begin{bmatrix} 0 \\ b_0 \\ 0 \end{bmatrix}$,

$$E = \begin{bmatrix} 0 \\ 0 \\ 1 \end{bmatrix}, C = \begin{bmatrix} 1 \\ 0 \\ 0 \end{bmatrix}^T.$$

A third-order ESO is constructed in the following form:

$$\begin{cases} \dot{\hat{x}}(t) = A\hat{x}(t) + Bu(t) + L[y_{\zeta}(t) - \hat{y}(t)] \\ \quad = (A - LC)\hat{x}(t) + Bu(t) + Ly_{\zeta}(t) \\ \hat{y}(t) = C\hat{x}(t) \end{cases} \quad (7)$$

where $\hat{x}(t)$ is the state vector of the ESO. Specifically, $\hat{x}_1(t)$ represents the estimation of the system output, $\hat{x}_2(t)$ denotes the estimation of the output velocity, and $\hat{x}_3(t)$ corresponds to the estimation of the extended state of the system. The observer gain vector is defined as $L = [\beta_1 \ \beta_2 \ \beta_3]^T$. By applying the linearization method proposed in [29], the following relationships can be obtained

$$\beta_1 = 3\omega_0, \beta_2 = 3\omega_0^2, \beta_3 = \omega_0^3 \quad (8)$$

where ω_0 is the bandwidth of the ESO. $\hat{x}_3(t)$ is introduced into the feedback loop to compensate for the total disturbance, and proportional–derivative (PD) control is introduced to obtain the control law:

$$u(t) = \frac{1}{b_0}(u_0(t) - \hat{x}_3(t)) = \frac{1}{b_0}[k_1(r(t) - \hat{x}_1(t)) - k_2\hat{x}_2(t) - \hat{x}_3(t)]. \quad (9)$$

where $[k_1 \ k_2]^T$ is the gain of the PD controller.

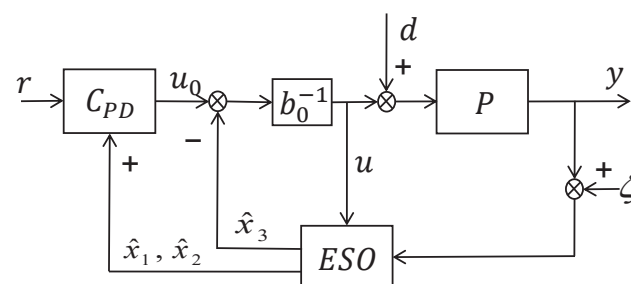


Figure 2. The system block diagram of the ADRC.

Remark 1. From the perspective of Equations (2)–(4), the motion platform is modeled as a second-order integrator chain subject to a lumped total disturbance. Under this formulation, the plant dynamics are intentionally simplified, and all unmodeled dynamics, parameter variations, and resonant effects are treated as part of the total disturbance acting on the system. This implies that even if partial information about the resonant modes is available, such information is not explicitly utilized in the model and is instead absorbed into the lumped disturbance term. Within the ESO framework, disturbances located within the observer bandwidth are estimated and compensated online. Consequently, the actual resonant frequencies play a more significant role in the observer

performance than the precise values or uncertainties of the resonant parameters themselves. As a result, uncertainties associated with the nominal parameters k_i and ω_i do not undermine the proposed methodology, since their effects are naturally handled through disturbance estimation rather than explicit model compensation.

3.2. Effects of the Disturbance and Measurement Noise

According to Formula (6), it is easy to obtain that

$$\begin{cases} \dot{\hat{x}}(t) = A_z \hat{x}(t) + B_z \begin{bmatrix} u(t) & y_\zeta(t) \end{bmatrix}^T \\ \hat{y}(t) = C_z \hat{x}(t) \end{cases} \quad (10)$$

where $A_z = \begin{bmatrix} -\beta_1 & 1 & 0 \\ -\beta_2 & 0 & 1 \\ -\beta_3 & 0 & 0 \end{bmatrix}$, $B_z = \begin{bmatrix} 0 & \beta_1 \\ b_0 & \beta_2 \\ 0 & \beta_3 \end{bmatrix}$, $C_z = C$. The transfer function matrix corresponding to the ESO is

$$C_z(sI - A_z)B_z = \frac{\begin{bmatrix} b_0s & \beta_1s^2 + \beta_2s + \beta_3 \\ b_0(s^2 + \beta_1s) & \beta_2s^2 + \beta_3s \\ -b_0\beta_3 & \beta_3s^2 \end{bmatrix}}{s^3 + \beta_1s^2 + \beta_2s + \beta_3} \quad (11)$$

The expression for $\hat{x}_3(s)$ is

$$\hat{x}_3(s) = \frac{\beta_3}{s^3 + \beta_1s^2 + \beta_2s + \beta_3} f(s) + \frac{\beta_3s^2}{s^3 + \beta_1s^2 + \beta_2s + \beta_3} \zeta(s) \quad (12)$$

Define the disturbance estimation error $e(t) = \hat{x}_3(t) - f(t)$ and substitute into Formula (11) to obtain the expression

$$e(s) = \frac{s^3 + \beta_1s^2 + \beta_2s}{s^3 + \beta_1s^2 + \beta_2s + \beta_3} f(s) + \frac{\beta_3s^2}{s^3 + \beta_1s^2 + \beta_2s + \beta_3} \zeta(s) \quad (13)$$

It can be seen that the estimation error is affected by both disturbance and measurement noise. The transfer function to evaluate the estimation performance is as follows

$$G_f(s) = \frac{e(s)}{f(s)} = \frac{s^3 + \beta_1s^2 + \beta_2s}{s^3 + \beta_1s^2 + \beta_2s + \beta_3} \quad (14)$$

$$G_\zeta(s) = \frac{e(s)}{\zeta(s)} = \frac{\beta_3s^2}{s^3 + \beta_1s^2 + \beta_2s + \beta_3}. \quad (15)$$

Combined with the tuning methodology of the linear extended state observer (LESO), the above expressions can be further reformulated as

$$G_f(s) = \frac{s^3 + 3\omega_0s^2 + 3\omega_0^2s}{s^3 + 3\omega_0s^2 + 3\omega_0^2s + \omega_0^3} \quad (16)$$

$$G_\zeta(s) = \frac{\omega_0^3s^2}{s^3 + 3\omega_0s^2 + 3\omega_0^2s + \omega_0^3}. \quad (17)$$

Combined with the practical characteristics of the system, it is assumed that the dominant frequency components of the total disturbance are much lower than the observer bandwidth. Under this assumption, the disturbance transfer function can be approximated as $G_f(s) = \frac{3s}{\omega_0}$, which indicates that a larger observer bandwidth leads to improved distur-

balance suppression performance. Moreover, the steady-state disturbance estimation error is bounded within $\frac{3}{\omega_0} |h(j\omega_0)|$. As illustrated in Figure 3, the disturbance rejection capability is significantly enhanced with an increase in the observer bandwidth ω_0 .

However, when the measurement noise predominantly lies in the high-frequency range relative to the observer bandwidth, the noise transfer function can be expressed as $G_\zeta(s) = \frac{\omega_0^3}{s}$. In this case, the squared term of β_3 increases rapidly, leading to a substantial amplification of high-frequency measurement noise as the observer bandwidth increases. This phenomenon may result in system instability or severe vibrations, which are unacceptable in high-precision servo control applications. Figure 4 depicts the frequency response of measurement noise in the LESO. Consequently, in the conventional LESO structure, improving disturbance rejection performance inevitably comes at the expense of reduced noise attenuation capability.

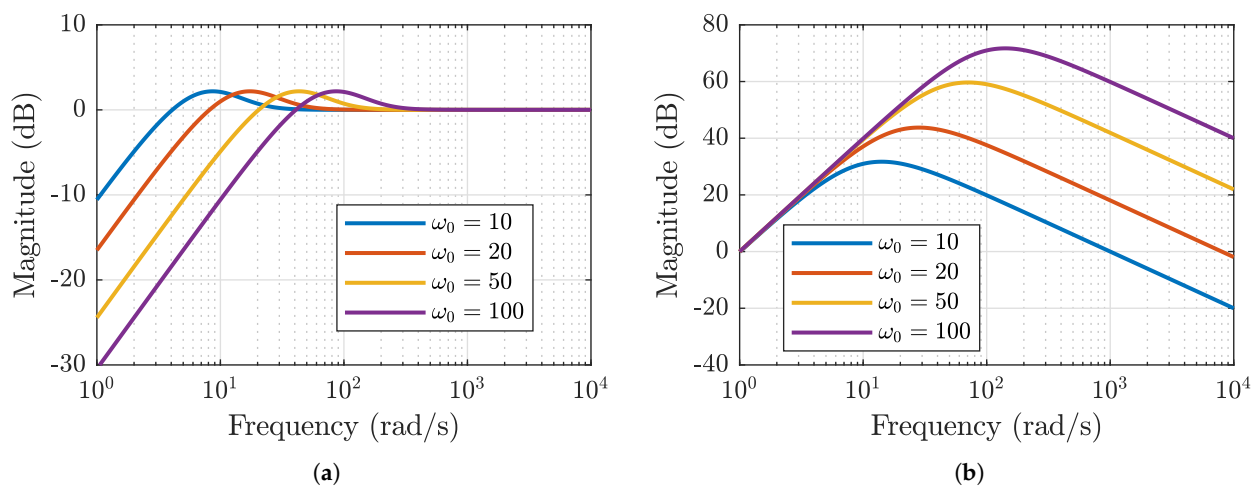


Figure 3. Amplitude–frequency responses of the disturbance estimation error under different observer bandwidths: (a) disturbance-to-estimation transfer characteristics; (b) noise-to-estimation transfer characteristics.

3.3. Design of the VGESO

The analysis in Section 3.2 reveals that the trade-off between disturbance rejection and measurement noise sensitivity in the conventional LESO is fundamentally rooted in its pole configuration. Specifically, increasing the observer bandwidth improves low-frequency disturbance estimation, but inevitably amplifies high-frequency measurement noise due to the fixed structure of the observer characteristic polynomial. This observation indicates that merely tuning the observer bandwidth is insufficient to simultaneously enhance disturbance rejection and suppress noise amplification. Instead, additional degrees of freedom are required at the structural level of the observer. Motivated by this insight, a variable-gain ESO based on pole placement is proposed in the following, where the observer gains are directly determined by the locations of the assigned poles.

Taking this second-order system as an example, a third-order ESO is designed, and the characteristic equation of the observer is as follows

$$|sI - (A - LC)| = s^3 + \beta_1 s^2 + \beta_2 s + \beta_3 \quad (18)$$

The poles of the designed observer satisfy the condition that the parameters of the above equation are real numbers. The pole $(-\omega_0, 0j)$ of the original LESO is set to a pole of the VGESO, which is called the main pole. l is the scaling factor of the other two secondary

poles, and θ is the angle between the secondary pole and the negative half-axis of the complex plane. According to Formula (7), the gain coefficient of VGESO can be solved as

$$\begin{cases} \beta_{1i} = (1 + 2 \cos \theta) \omega_0, \\ \beta_{2i} = (l^2 + 2l \cos \theta) \omega_0^2, \\ \beta_{3i} = l^2 \omega_0^3. \end{cases} \quad (19)$$

where $\begin{bmatrix} \beta_{1i} & \beta_{2i} & \beta_{3i} \end{bmatrix}$ is the gain of VGESO.

When $l = 1, \theta = 0$, the VGESO degenerates to the original LESO. In order to reduce the amplification effect of the observer on high-frequency noise, $0 < l < 1$ is set so that $\beta_{3i} < \beta_3$. At the same time, in order to enhance the disturbance rejection ability on low-frequency, $\frac{\beta_{2i}}{\beta_{3i}} < \frac{\beta_2}{\beta_3}$ should be satisfied. Solving the above inequality yields

$$\arccos l < \theta < \frac{1}{2}\pi. \quad (20)$$

Remark 2. Selecting three different poles on the real negative half-axis can also satisfy the characteristic equation of the observer. But doing so deteriorates the performance compared to LESO selecting three poles in the same position. This conclusion can be readily derived from the Cauchy inequality.

3.4. Effects of the VGESO Parameters

In the previous subsection, the design idea of VGESO is presented. In this subsection, the two introduced parameters will be analyzed by bode diagram to see how they affect the disturbance and noise.

Substituting (18) into (13) and (14) yields

$$G_{fi}(s) = \frac{s^3 + (1 + 2 \cos \theta) \omega_0 s^2 + (l^2 + 2l \cos \theta) \omega_0^2 s}{s^3 + (1 + 2 \cos \theta) \omega_0 s^2 + (l^2 + 2l \cos \theta) \omega_0^2 s + l^2 \omega_0^3} \quad (21)$$

$$G_{\zeta i}(s) = \frac{l^2 \omega_0^3 s^2}{s^3 + (1 + 2 \cos \theta) \omega_0 s^2 + (l^2 + 2l \cos \theta) \omega_0^2 s + l^2 \omega_0^3} \quad (22)$$

According to the admissible range of the parameter l obtained in the previous subsection, several representative values of different magnitudes are selected to investigate the influence of l on the performance of the VGESO. The Bode plots of the transfer functions from disturbance and measurement noise to the estimation error for $\omega_0 = 100$ rad/s and $\theta = 0$ are shown in Figure 4. When $l = 1$, the VGESO degenerates into a LESO. It can be observed that, as l decreases, the influence of measurement noise on the estimation result is gradually attenuated. However, this reduction in noise sensitivity is accompanied by a degradation in the disturbance estimation capability of the VGESO. Adjusting the parameter l has an effect similar to tuning the observer bandwidth. Therefore, when the observation bandwidth is constrained, enhanced noise immunity can be achieved by appropriately reducing the value of l .

Several representative values of θ are selected based on the admissible range obtained in the previous subsection to study its influence on VGESO performance. The Bode plots of the transfer functions from disturbance and noise to estimation for $\omega_0 = 100$ rad/s and $l = 1$ are shown in Figure 5. When $\theta = 0$, the VGESO degenerates into a LESO. Within the selected range, increasing θ enhances the disturbance observation capability in the low-frequency band without amplifying high-frequency noise. However, the estimation performance in the mid-frequency band, near the observation bandwidth, is slightly degraded. Consequently, when ω_0 is approximately fixed, the loss of disturbance estimation capability caused by reducing l can be compensated by adjusting θ . In practical engineering

applications, VGESO can achieve improved control performance when combined with frequency-domain analysis of system disturbances.

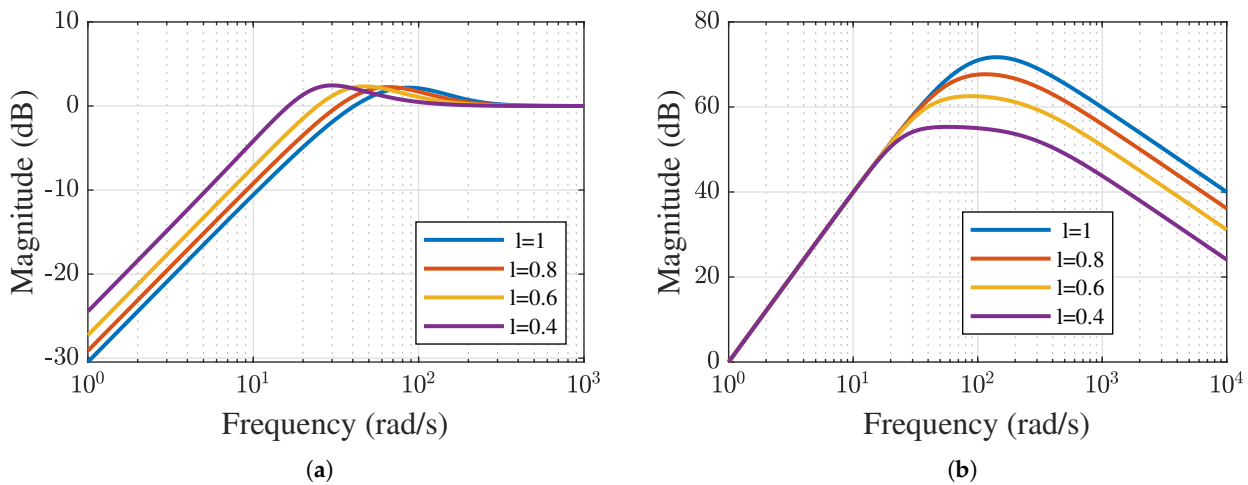


Figure 4. Amplitude–frequency responses of the disturbance estimation error under different l : (a) disturbance-to-estimation transfer characteristics; (b) noise-to-estimation transfer characteristics.

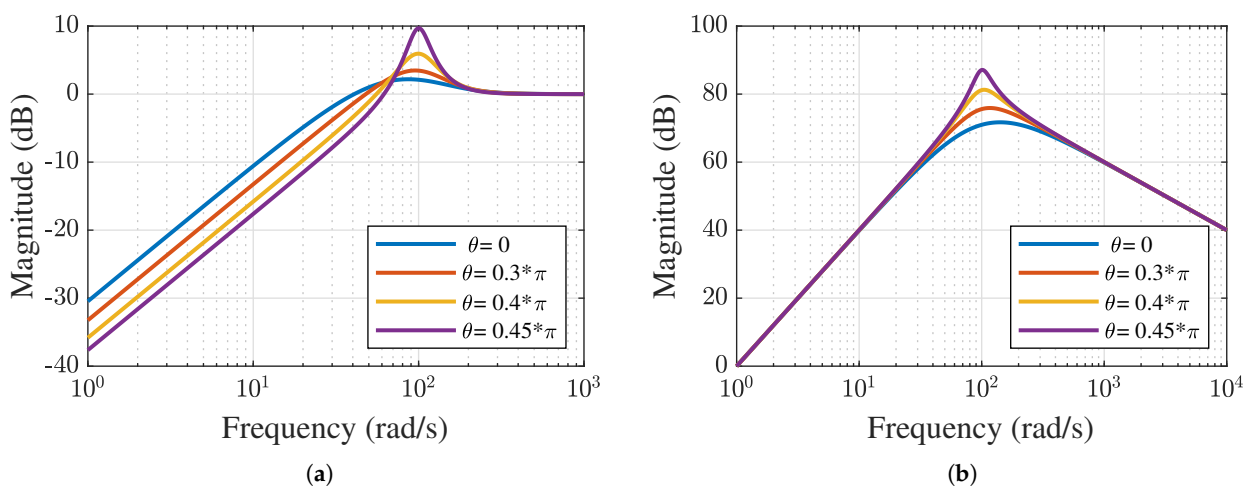


Figure 5. Amplitude–frequency responses of the disturbance estimation error under different θ : (a) disturbance-to-estimation transfer characteristics; (b) noise-to-estimation transfer characteristics.

Remark 3. Introducing two additional parameters from a pole-placement perspective to match the number of ESO parameters with the number of observer poles does not complicate the design process. On the contrary, it makes the tuning procedure more intuitive, since each parameter is associated with a specific effect. The improved extended state observer (IESO) proposed in [20] introduces an additional parameter to directly reduce β_3 , thereby alleviating noise amplification. This approach can also be interpreted as modifying the ESO pole locations. However, adjusting this parameter inevitably increases the dominant pole as well. For platforms with high-order flexible dynamics, such a change may lead to degraded stability margins or even system instability.

For practical implementation, the tuning of the proposed VGESO can be carried out in a sequential and intuitive manner. First, the observer bandwidth ω_0 is selected following the same principle as in a conventional ESO, such that sufficient disturbance estimation performance is achieved while maintaining closed-loop stability. Second, the parameter l is introduced to scale the secondary observer poles. By choosing $l < 1$, the high-frequency gain of the observer is effectively reduced, thereby suppressing the

amplification of measurement noise. Finally, the parameter θ is adjusted to compensate for the degradation in low-frequency disturbance rejection caused by the reduction of l . By appropriately selecting θ , the desired low-frequency performance can be recovered without sacrificing the noise attenuation benefit. This step-by-step tuning procedure provides a clear guideline for engineers and highlights the practical advantages of the proposed pole-placement-based VGESO.

To clearly distinguish the proposed VGESO from the conventional LESO, its main advantages are summarized as follows. First, the tuning parameters are assigned decoupled and physically interpretable roles. Specifically, the parameter l primarily governs the high-frequency gain of the observer, thereby enabling effective suppression of measurement noise, while the parameter θ mainly shapes the low-frequency disturbance estimation performance. Second, the proposed VGESO does not increase the observer order and retains the same third-order structure as the conventional LESO, which facilitates straightforward implementation and avoids additional computational burden. Third, by appropriately designing the observer poles, the gain β_3 can be significantly reduced compared with that of a conventional LESO with the same bandwidth. This directly alleviates noise amplification in the disturbance estimation channel, which is a critical issue in high-precision motion control systems. Finally, the pole-placement-based design provides greater flexibility than the bandwidth-only tuning strategy of the conventional LESO, allowing a more favorable trade-off between disturbance rejection and noise attenuation without sacrificing system stability.

4. Stability Analysis

This section analyzes the stability of the proposed VGESO. In general, the stability of an ADRC-based control system needs to be examined from both the ESO and the closed-loop control perspectives. The stability of the ESO requires that all roots of its characteristic equation lie in the left half-plane, which constitutes a fundamental prerequisite for the proposed design. Since the observer parameters are selected to satisfy the prescribed pole-placement criteria, the resulting ESO naturally guarantees internal stability.

It is pointed out in [30,31] that the internal stability of LESO- and LADRC-based systems composed of (6), (7), and (9) is independent of external disturbances. Moreover, since ADRC is an input–output-oriented control framework, the bounded-input bounded-output (BIBO) stability of the ADRC system is analyzed in the following.

$$u = \frac{1}{b_0} \left[k_1(r - z_1) + k_2(\dot{r} - z_2) + (\ddot{r} - z_3) \right] \quad (23)$$

Substituting (22) into (4) yields

$$\ddot{y} = k_1(r - \hat{x}_1) + k_2(\dot{r} - \hat{x}_2) + (\ddot{r} - \hat{x}_3) + f \quad (24)$$

Let $r_1 = r, r_2 = \dot{r}, r_3 = \ddot{r}$, The tracking error of the system is defined as $\varepsilon_i = r_i - x_i$, $i = 1, 2$. It yields

$$\begin{cases} \dot{\varepsilon}_1 = \dot{r}_1 - \dot{x}_1 = \varepsilon_2, \\ \dot{\varepsilon}_2 = \dot{r}_2 - \dot{x}_2 = -k_1\varepsilon_1 - k_2\varepsilon_2 - k_1e_1 - k_2e_2 - e_3. \end{cases} \quad (25)$$

Let $\varepsilon = \begin{bmatrix} \varepsilon_1 & \varepsilon_2 \end{bmatrix}^T, e = \begin{bmatrix} e_1 & e_2 & e_3 \end{bmatrix}^T$. Then, the closed-loop system can be rewritten as

$$\dot{\varepsilon} = A_1\varepsilon + A_2e \quad (26)$$

$$\text{where } A_1 = \begin{pmatrix} 0 & 1 \\ -k_1 & -k_2 \end{pmatrix}, A_2 = \begin{pmatrix} 0 & 0 & 0 \\ -k_1 & -k_2 & -k_3 \end{pmatrix}.$$

Theorem 1. *If the observation error of VGESO satisfies $\lim_{t \rightarrow +\infty} \|e\|_2 = 0$, there exists a control gain $\begin{bmatrix} k_1 & k_2 \end{bmatrix}^T$, such that the tracking error of the close-loop system (25) tends to zero.*

Proof. The solution of (23) is

$$\varepsilon(t) = \exp(A_1 t) \varepsilon(0) + \int_0^t \exp(A_1(t - \tau)) A_2 e(\tau) d\tau \quad (27)$$

□

The characteristic equation polynomial of A_1 is given by

$$|\lambda I - A_1| = \lambda^2 + k_2 \lambda + k_1 \quad (28)$$

By choosing $k_1, k_2 > 0$, the characteristic equation of A_1 possesses two negative real roots, i.e., $|\lambda I - A_1| = (\lambda + \lambda_1)(\lambda + \lambda_2)$, ensuring the stability of the system. So there exists a nonsingular matrix T such that

$$A_1 = T \operatorname{diag}\{-\lambda_1, -\lambda_2\} T^{-1} \quad (29)$$

There exist a $\beta > 0$ such that

$$\begin{aligned} \|\exp(A_1 t)\|_2 &= \|T\|_2 \|T^{-1}\|_2 \exp(-\lambda_1 t) \exp(-\lambda_2 t) \\ &\leq \beta \exp(-\lambda_1 t), \end{aligned} \quad (30)$$

Taking the limit yields

$$\lim_{t \rightarrow +\infty} \|\exp(A_1 t)\|_2 = 0 \quad (31)$$

Similarly, it can be derived that

$$\|\exp(A_1(t - \tau))\|_2 \leq \beta \exp(-\lambda_1(t - \tau)) \quad (32)$$

Since the observation error of the system tends to zero, for any $\eta > 0$, there exists t_0 such that e is bounded when $t > t_0$, i.e., $\|e\|_2 < M$.

The derivation of the second part of the right-hand side of Equation (26) can be obtained

$$\begin{aligned} &\int_0^t \exp(A_1(t - \tau)) A_2 e d\tau \\ &= \int_0^{t_0} \exp(A_1(t - \tau)) A_2 e d\tau + \int_{t_0}^t \exp(A_1(t - \tau)) A_2 e d\tau \\ &\leq \beta \|A_2\|_2 \alpha \exp(-\lambda_1 t) \int_0^{t_0} \exp(-\lambda_1 \tau) d\tau \\ &\quad + \beta \|A_2\|_2 \eta \exp(-\lambda_1 t) \int_0^{t_0} \exp(-\lambda_1 \tau) d\tau \\ &= \beta \|A_2\|_2 \alpha \int_0^{t_0} \exp(-\lambda_1 \tau) d\tau \exp(-\lambda_1 t) \\ &\quad + \beta \|A_2\|_2 \eta \frac{\exp(-\lambda_1 t) - \exp(-\lambda_1 t_0)}{\lambda_1} \\ &\leq M_1 \exp(-\lambda_1 t) + M_2 \eta. \end{aligned} \quad (33)$$

where $M_1 = \beta \|A_2\|_2 \alpha \int_0^{t_0} \exp(-\lambda_1 \tau) d\tau$, $M_2 = \frac{\beta \|A_2\|_2}{\lambda_1}$. Taking the limit yields

$$\lim_{t \rightarrow +\infty} \int_0^t \exp(A_1(t-\tau)) A_2 e d\tau = 0 \quad (34)$$

Combining equalities (30) and (33) yields

$$\lim_{t \rightarrow +\infty} \|\varepsilon\|_2 = 0 \quad (35)$$

The above theorem proves that the tracking error converges to zero if the estimation error converges to zero. More specifically, the output of the system will converge to the given trajectory. From a performance perspective, the disturbance and measurement noise enter the closed-loop system through the observation error vector e . As shown in Section 3, the proposed VGESO reduces the high-frequency amplification of measurement noise while preserving low-frequency disturbance estimation accuracy. Consequently, the magnitude of e is reduced over a wide frequency range compared with the conventional LESO. According to (27)–(35), the tracking error ε is directly driven by the observation error. Therefore, improving the disturbance estimation accuracy and suppressing noise-induced errors through VGESO leads to a tighter bound on the tracking error and enhanced closed-loop performance, even though the asymptotic stability condition remains unchanged.

5. Simulation Verification

To clearly illustrate the distinct roles of disturbance rejection and noise attenuation, two additional simulation cases are considered. In the first case, only external disturbances are applied while measurement noise is removed. In the second case, only measurement noise is injected without external disturbances. This separation allows the time-domain simulation results to directly reflect the frequency-domain analysis presented in Section 3, without altering the observer parameters.

The dynamic model of the system is given as follows

$$P(s) = \frac{1}{ms^2} + \frac{1}{m} \sum_{i=1}^3 \frac{k_i}{s^2 + 2\xi_i \omega_i s + \omega_i^2} \quad (36)$$

Assume that the actual mass of the system is $m = 24$. The nominal mass is set to $\hat{m} = 0.99 m$. The gain, the resonant frequency and the damping ratio of the three resonant points of the system are, respectively,

$$\begin{aligned} k_1 &= 0.013, \omega_1 = 1502\pi, \xi_1 = 0.04; \\ k_2 &= 0.0023, \omega_2 = 2922\pi, \xi_2 = 0.018; \\ k_3 &= 0.0038, \omega_3 = 5102\pi, \xi_3 = 0.0027. \end{aligned}$$

The system input is subjected to cogging force as

$$F_{\text{cogging}}(y) = \sum_{i=1}^7 \left(S_i \sin\left(\frac{2i\pi}{P} y\right) + C_i \cos\left(\frac{2i\pi}{P} y\right) \right) \quad (37)$$

where y is the system output. The magnetic dipole moment is set to $P = 0.0024$. The other constant coefficients are set to

$$\begin{aligned} S_1 &= 0.7, C_1 = -1.2, S_2 = 0.95, C_2 = 2.15, \\ S_3 &= 0.2, C_3 = -0.7, S_4 = -0.5, C_4 = -0.06, \\ S_5 &= 0.5, C_5 = -2.1, S_6 = -0.175, C_6 = -0.6. \end{aligned}$$

The total disturbance of the plant is the sum of high-order resonance, mass uncertainty and periodic cogging force. In the frequency domain, although the total disturbance in Equation (4) is distributed across nearly the full frequency range, the cogging component in the low-frequency band contributes predominantly, which is also the primary frequency range of focus in the simulation. According to [29], the feedback controller adopts a PD structure, as given in (9), with a control bandwidth of 60 Hz. The observer bandwidth of the ESO is set to approximately 160 Hz (equivalent to 1000 rad/s), and the parameters of the VGESO are configured as $l = 0.8$ and $\theta = 0.4\pi$.

The measurement noise is introduced to evaluate the noise sensitivity of the observer. The noise is modeled as band-limited white noise with dominant frequency components above the observer bandwidth. Specifically, the noise spectrum is concentrated in the range of 300–2000 Hz, which is significantly higher than both the control bandwidth (60 Hz) and the observer bandwidth (160 Hz). The RMS amplitude of the noise is set to the 1×10^{-7} m, consistent with practical high-precision position sensing systems. Figure 6 shows the third-order sigmoid reference trajectory adopted to emulate a typical scanning motion in lithography stages.

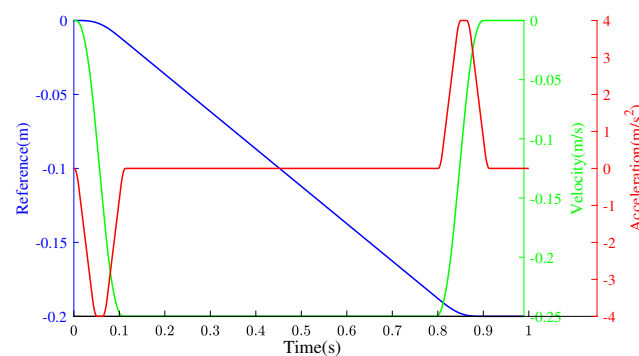


Figure 6. Reference trajectory.

Figure 7 illustrates the simulation results under the disturbance-dominated condition, where external disturbances are dominant and the influence of measurement noise is negligible. The analysis focuses on the constant-velocity segment of the motion profile. It can be seen that both the conventional ESO and the proposed VGESO are capable of accurately estimating the lumped disturbance, which significantly improves the positioning accuracy during steady motion. Under the same observer bandwidth, the VGESO provides enhanced low-frequency disturbance estimation, resulting in reduced residual tracking errors.

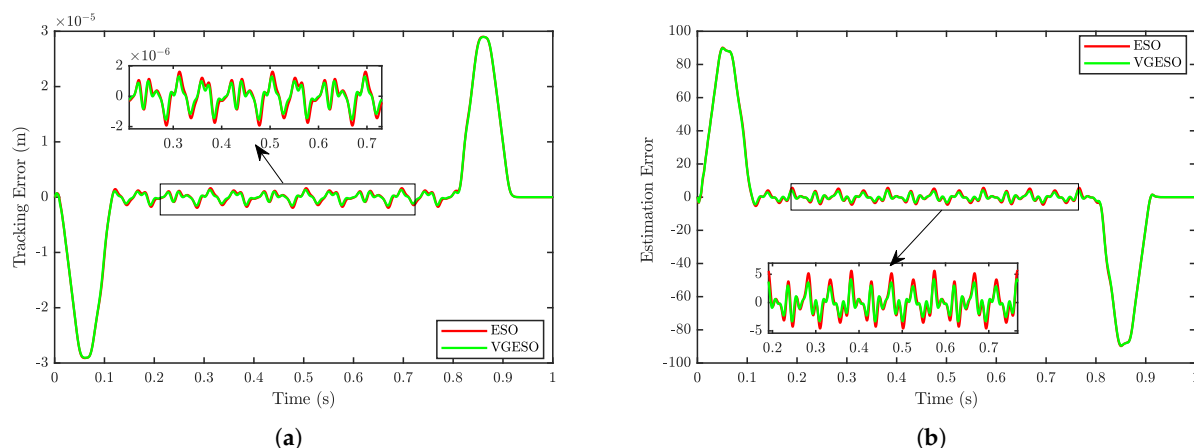


Figure 7. Performance comparison under disturbance-dominated condition: (a) tracking error; (b) estimation error.

Figure 8 presents the simulation results under the noise-dominated condition, in which the measurement noise becomes the dominant factor affecting estimation performance. Through pole reconfiguration, the proposed VGESO reshapes the noise propagation characteristics and suppresses high-frequency noise amplification. As a result, the VGESO achieves smoother tracking behavior and reduced high-frequency error components compared with the conventional ESO, without relying on bandwidth retuning.

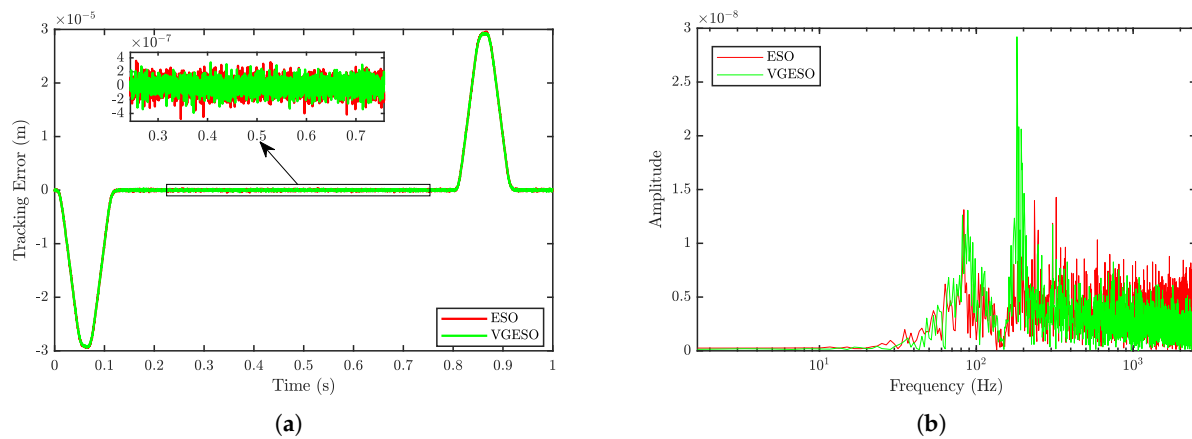


Figure 8. Performance comparison under the noise-dominated condition (window size = 50): (a) time-domain tracking error; (b) frequency-domain amplitude spectrum of the tracking error.

For clarity, Table 1 summarizes the key structural and tuning differences between the conventional ESO and the proposed VGESO.

Table 1. Parameter comparison between the conventional ESO and the proposed VGESO.

Item	Conventional ESO (LESO)	Proposed VGESO
Observer order	Third-order	Third-order
Gain parameterization	Single bandwidth parameter ω_0	Pole-placement-based parameters (ω_0, l, θ)
Observer poles	Three identical real poles at $-\omega_0$	One dominant real pole at $-\omega_0$ and a pair of complex-conjugate secondary poles
ESO gains	$\beta_1 = 3\omega_0$, $\beta_2 = 3\omega_0^2$, $\beta_3 = \omega_0^3$	$\beta_1 = (1 + 2 \cos \theta)\omega_0$, $\beta_2 = (l^2 + 2l \cos \theta)\omega_0^2$, $\beta_3 = l^2\omega_0^3$
Disturbance rejection	Improved by increasing ω_0	Enhanced via coordinated tuning of ω_0 and θ
Noise attenuation	Degraded at high ω_0 due to noise amplification	Improved by adjusting $l < 1$
Tuning flexibility	Limited (single degree of freedom)	Increased (independent regulation of low- and high-frequency behavior)

Although the proposed VGESO is developed and validated through simulations, several practical implementation aspects deserve further discussion. First, the VGESO preserves the same observer order as the conventional LESO, and only modifies the observer gain structure through pole placement. Therefore, it does not increase the computational burden compared with traditional ESO-based schemes. The observer can be implemented

using standard discrete-time realizations on commonly used real-time platforms, such as DSPs or FPGAs, without requiring additional memory or processing resources. Second, the introduction of the two tuning parameters, l and θ , does not complicate the tuning procedure. On the contrary, these parameters provide clear physical interpretations from a pole-placement perspective, allowing engineers to independently adjust noise attenuation and low-frequency disturbance rejection according to system requirements. This decoupled tuning characteristic improves robustness against modeling uncertainties and measurement noise in practical applications. Finally, practical challenges mainly arise from measurement noise characteristics and actuator bandwidth limitations. In high-precision motion systems, sensor noise spectra are usually known or can be identified experimentally. The proposed VGESO allows the observer poles to be flexibly adjusted to accommodate these noise characteristics without sacrificing low-frequency disturbance estimation performance, which enhances its engineering applicability.

6. Conclusions

This paper proposed a pole-placement-based VGESO for high-precision motion control of lithography stages. By introducing two pole-related parameters, the observer gains can be adjusted in a structured and intuitive manner, enabling a flexible trade-off between low-frequency disturbance rejection and high-frequency noise attenuation. Different from conventional high-gain ESO designs, the proposed VGESO reshapes the inherent coupling between disturbance suppression and noise amplification without increasing the observer order. Theoretical analysis and comparative simulations demonstrate that the proposed method achieves improved disturbance estimation accuracy and enhanced noise immunity over the conventional LESO, particularly under bandwidth-limited conditions. Future work will focus on experimental validation on a multi-degree-of-freedom motion stage and further investigation of the proposed observer in the presence of nonlinear effects and time-varying uncertainties.

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