

Synchronous Vibration Suppression for AMB Systems based on MCCF-PLL and Positive-negative-sequence SRF-PI

Tianci Sun, *Student Member, IEEE*, Dong Jiang, *Senior Member, IEEE*

Abstract—Active magnetic bearings (AMBs) support rotors through controllable electromagnetic forces and are widely used in high-performance rotating machinery. In practical AMB systems, synchronous vibration is a major vibration source, and its suppression requires accurate rotational speed and phase information without an additional speed sensor. However, conventional synchronous-reference-frame (SRF)-based observation and compensation methods are sensitive to amplitude mismatch, phase nonorthogonality, direct current (DC) offsets, and harmonic contamination, while insufficient sequence separation further degrades speed observation and vibration suppression. To address these issues, this paper proposes a sensorless synchronous vibration suppression method based on a multiple-complex coefficient-filter-based phase-locked loop (MCCF-PLL) and positive-negative-sequence SRF-PI control. The MCCF-PLL extracts the fundamental positive-sequence current component for robust speed and phase observation, while the displacement error is decomposed into positive- and negative-sequence components and regulated by corresponding SRF-PI compensators. The compensation angles are designed using sequence-domain dynamic compliance analysis and updated online during variable-speed operation. Simulation results verify the effectiveness of the proposed method under non-ideal signal conditions. Experimental results demonstrate improved speed-observation performance at 3000 r/min, effective vibration suppression at 1800 and 3600 r/min, and the ability to maintain effective suppression during the run-up from 900 to 3900 r/min and the run-down from 3900 to 900 r/min. These results confirm its feasibility for practical AMB systems.

Index Terms—Active magnetic bearing (AMB), Frequency estimation, Multiple-complex coefficient-filter-based phase-locked loop (MCCF-PLL), Synchronous-reference-frame PI (SRF-PI), Vibration suppression.

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Tianci Sun and Dong Jiang are with the School of Electrical and Electronic Engineering, Huazhong University of Science and Technology, Wuhan 430074, China (e-mail: suntc@hust.edu.cn; jiangd@hust.edu.cn).

(Corresponding author: Dong Jiang)

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I. INTRODUCTION

ACTIVE magnetic bearings (AMBs) realize contactless rotor support by means of controllable electromagnetic forces. Owing to their advantages of zero mechanical friction, no lubrication requirement, and actively adjustable supporting characteristics [1], AMBs have shown broad application prospects in high-speed motors, turbomachinery, flywheel energy storage systems, and vacuum rotating equipment [2]. However, synchronous vibration is a common problem in AMB-supported rotating machinery during practical operation. The synchronous unbalance force caused by rotor mass unbalance is one of the major vibration sources [3]. In addition, unbalanced magnetic pull, coupling misalignment, and sensor run-out may also become important sources of periodic disturbances [4].

Synchronous vibration not only deteriorates the rotor orbit and reduces rotational accuracy, but may also increase the control current [5], and in severe cases even lead to rotor-stator rub. On the other hand, AMB-suspended rotating machinery is often not equipped with an additional speed sensor [6]. Therefore, how to achieve accurate observation of synchronous rotational speed and phase under sensorless conditions, and on this basis design effective vibration suppression algorithms, has become an important issue in the control of AMB systems.

With regard to synchronous observation under sensorless conditions, existing methods can be broadly classified into the following categories.

Methods based on phase-locked loop (PLL). PLL-based frequency estimation is widely used in practical applications, particularly the synchronous reference frame PLL (SRF-PLL). Yang et al. proposed a quadrature PLL method implemented on a field-programmable gate array (FPGA) [7]. Xu et al. applied PLL to rotational speed and phase observation and derived tuning formulas for the PLL bandwidth and damping ratio [6]. To improve the robustness of PLL under nonorthogonality of x/y signals, Peng et al. proposed an improved dual second-order generalized integrator PLL (DSOGI-PLL) [8]. In addition, Xu et al. proposed a multiresonance controller-quadrature PLL (MRC-QPLL) to mitigate the influence of harmonics on speed estimation [9].

Methods based on frequency-locked loop (FLL). FLL methods usually require the construction of an error signal for

frequency adaptation [10]. Sun et al. applied generalized integrator-FLL (GI-FLL) to rotational speed observation [11]. Wu et al. proposed a phase-shifted second-order generalized integrator frequency-locked loop (SOGI-FLL) to improve low-frequency stability [12]. Cui et al. proposed a modified double SOGI-FLL method for rotational speed observation [13]. Moreover, Zeng et al. employed a complex bandpass filter-FLL (CBF-FLL) to extract rotor frequency information from displacement signals [14].

Another class of methods in the AMB field is frequency estimation based on adaptive notch filters (ANFs) [5], [15]. ANF-based methods have relatively simple structures, but Xu et al. indicate that their dynamic performance and estimation accuracy still need further improvement [9].

In summary, PLL-based methods feature clear structures, provide both frequency and phase information, and can be integrated with compensation based on synchronous coordinate transformation (CT). However, when input signals suffer from nonorthogonality, their estimation performance usually degrades, which is common in practical AMB displacement or current signals. FLL-based methods exhibit good frequency-tracking capability, but they usually lack a phase-locking mechanism and suffer from limitations in low-frequency stability and dynamic performance.

For synchronous disturbance suppression, existing methods can be classified into current vibration suppression, force vibration suppression, and displacement vibration suppression according to the controlled variable [16]-[17]. This study focuses on synchronous displacement vibration suppression, which is of more direct engineering significance in high-precision rotating equipment such as electric spindles and magnetically suspended flywheels [18].

Sun et al. applied a PSGI to displacement vibration suppression [11], while Han et al. adopted repetitive control [19]. Subsequently, some studies introduced modern feedback control and state observation concepts into synchronous displacement suppression [20]. Bi et al. improved iterative learning control (ILC) into automatic learning control (ALC) for displacement and current vibration suppression through switching strategies [21]. Gao et al. further improved ALC to accelerate convergence and enhance suppression performance [22]. In recent years, increasing attention has been paid to SRF-based synchronous vibration suppression methods. Zheng et al. proposed an SRF-based synchronous vibration suppression method, which represents a typical SRF-based synchronous compensation method in the AMB field [23]. On this basis, Wu et al. proposed a SOGI-SRF-PI unbalance compensation method based on single-phase CT [24]. Xu et al. proposed a sensorless unbalance control strategy based on CT compensation, which is essentially a synchronous-reference-frame PI (SRF-PI) type compensation structure [6].

Overall, SRF-based methods have attracted increasing attention because of their clear control objectives, systematic tuning, and integration with sensorless frequency observation. However, insufficient sequence-component separation makes conventional SRF processing susceptible to negative-sequence components and harmonic contamination.

Therefore, it is necessary to establish a control framework capable of stably providing $\omega(t)$ and $\theta(t)$ under complex signal conditions, and further achieving selective suppression of synchronous components on this basis.

A promising direction is to employ sequence-component extraction to provide cleaner sequence-domain information for synchronous observation and compensation. In power electronics, the multiple-complex coefficient-filter-based phase-locked loop (MCCF-PLL) employs multiple complex-coefficient filters (CCFs) to selectively extract positive-sequence and negative-sequence fundamental components as well as harmonic components [25]-[26]. Motivated by this idea, this paper proposes a synchronous displacement vibration suppression method for AMB systems based on MCCF-PLL and positive-negative-sequence SRF-PI control. MCCF-PLL is introduced on the current side to obtain more stable synchronous information of the fundamental sequence component, and the rotational speed and phase are observed based on the fundamental positive-sequence current. In the displacement error channel, MCCF is introduced to extract the fundamental positive-sequence and negative-sequence components, which are then combined with the SRF-PI synchronous compensation structure to achieve sequence-selective suppression of synchronous displacement components. Finally, both simulations and experiments are conducted to verify the effectiveness of the proposed method.

The remainder of this paper is organized as follows. Section II introduces the AMB model and analyzes the limitations of the conventional SRF-PLL under non-ideal conditions. Section III presents the MCCF-PLL-based phase and speed estimation method. Section IV develops the positive-negative-sequence SRF-PI synchronous vibration suppression method and its stability analysis. Section V and Section VI provide the simulation and experimental results, respectively. Finally, Section VII concludes the paper.

II. AMB MODELING AND ANALYSIS OF SRF-PLL

A. Structure of the Radial AMB System

In general, a typical AMB-supported rotor system is actively controlled in five degrees of freedom. A single radial plane of an AMB, which contains two translational degrees of freedom in the x - and y -directions, is illustrated in Fig. 1.

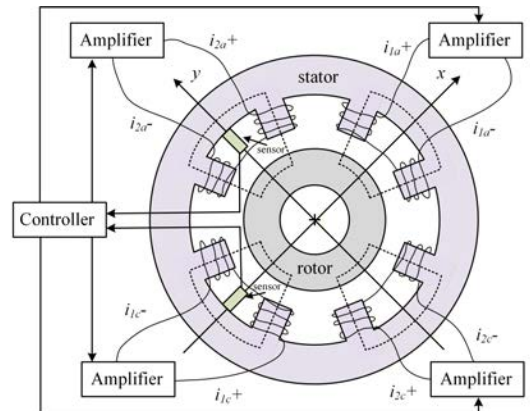


Fig. 1. Structure of the radial AMB.

During each control period, the displacement sensors measure rotor position, while the current sensors acquire winding current signals. The outer position controller generates the control command according to the displacement error. Subsequently, the inner current controller produces the duty-cycle command based on this reference and drives the power amplifier to regulate the winding currents. The electromagnets then generate the required electromagnetic forces. When the rotor deviates from its equilibrium position, the control system adjusts the currents of relevant poles to produce a net restoring force toward the equilibrium position, thereby dynamically correcting the rotor position and maintaining stable levitation within the air gap.

The control block diagram of a single-degree-of-freedom AMB is shown in Fig. 2. The AMB system can be regarded as a dual closed-loop control system, which mainly includes the position controller (typically implemented as a proportional-integral-derivative (PID) controller), the current loop (including the current controller and the power amplifier), the AMB electromagnetic actuator, and the rotor dynamic plant. Here, Δi_x denotes the compensating current injected into the current-reference channel, and i_d denotes the equivalent disturbance current converted from the external disturbance force.

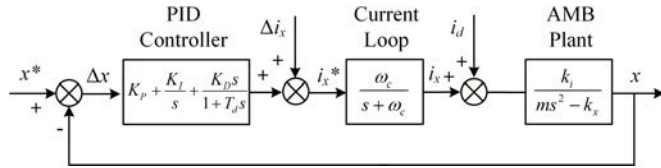


Fig. 2. Control block diagram of the single-degree-of-freedom AMB system.

The PID controller transfer function can be expressed as:

$$C_{PID}(s) = k_p + \frac{k_i}{s} + \frac{k_d s}{1 + T_d s} \quad (1)$$

where k_p , k_i , and k_d are the proportional, integral, and derivative gains, respectively, and T_d is the time constant of the first-order low-pass derivative filter.

The closed-loop current loop can be modeled as a low-pass element, whose transfer function is given by (2):

$$G_i(s) = \frac{\omega_{cl}}{s + \omega_{cl}} \quad (2)$$

where ω_{cl} denotes the closed-loop bandwidth of the current loop, which reflects the dynamic response of the current controller and power amplifier.

The actuator electromagnetic force is a nonlinear function of rotor displacement and coil current. When the displacement is much smaller than the air gap, the resultant force produced by a pair of opposing actuators along the same degree of freedom can be linearized as (3):

$$k_i i + k_x x = m \ddot{x} \quad (3)$$

where x is the displacement, $\ddot{x}$ is the acceleration, m is the rotor mass, i is the control current, k_i is the force-current stiffness, and k_x is the force-displacement stiffness. Applying the Laplace transform gives the AMB plant transfer function:

$$G_p(s) = \frac{X(s)}{I(s)} = \frac{k_i}{ms^2 - k_x} \quad (4)$$

It should be noted that (3) and (4) are local simplified models for a single radial degree of freedom. Gyroscopic effects are not explicitly included; within the present experimental range of 900–3900 r/min, they are not dominant, while higher-speed cases require a more complete multi-degree of freedom (DOF) model.

B. Sequence-domain Analysis of Non-ideal Signals

In this study, the rotational speed is estimated from the x/y control current signals in the same radial plane. Let the two signals be denoted by $i_x(t)$ and $i_y(t)$. Under ideal conditions, these signals can be regarded as sinusoidal control current components superimposed on direct current (DC) components. The DC components represent the steady control currents required to balance the projection of gravity along the corresponding degrees of freedom, while the ac components are caused by synchronous vibration induced by rotor unbalance. For the ac components, the ideal case corresponds to equal amplitudes, no harmonic contamination, and a phase difference of 90° . The following expression is adopted:

$$\begin{aligned} i_x(t) &= c_x + A_x \cos(\theta(t) + \varphi_x) + \sum_{h=2}^H A_{xh} \cos(h\theta(t) + \varphi_{xh}) \\ i_y(t) &= c_y + A_y \cos(\theta(t) + \varphi_x \pm 90^\circ + \Delta\varphi) + \sum_{h=2}^H A_{yh} \cos(h\theta(t) + \varphi_{xh} \pm 90^\circ + \Delta\varphi) \end{aligned} \quad (5)$$

Here, $\theta(t)$ is the rotor phase, h and H denote the harmonic order and maximum harmonic order, respectively, and φ_x and φ_{xh} are the phase offsets of the fundamental and h -th harmonic components. Amplitude mismatch is characterized by $A_x \neq A_y$ (and $A_{xh} \neq A_{yh}$); phase nonorthogonality is described by $\Delta\varphi$. Moreover, c_x and c_y denote DC offsets or low-frequency drifts. Therefore, the subsequent synchronous observation and sequence-component analysis focus on the AC components of $i_x(t)$ and $i_y(t)$, since the DC components do not contain rotational speed or phase information.

Defining the x - and y -axis signals in complex-vector form, the standard positive- and negative-sequence decomposition can be obtained through analytical derivation as (6):

$$\mathbf{i}(t) = i_x(t) + j i_y(t) = I_+ e^{j\theta(t)} + I_- e^{-j\theta(t)} \quad (6)$$

where the complex amplitudes of the positive- and negative-sequence components I_+ and I_- are respectively given by:

$$\begin{aligned} I_+ &= \frac{1}{2} (A_x + A_y \cos \Delta\varphi) - \frac{j}{2} A_y \sin \Delta\varphi \\ I_- &= \frac{1}{2} (A_x - A_y \cos \Delta\varphi) - \frac{j}{2} A_y \sin \Delta\varphi \end{aligned} \quad (7)$$

It can be seen from (7) that amplitude mismatch and nonorthogonality decompose the originally ideal single positive-sequence fundamental component into the superposition of a positive-sequence term $e^{j\theta}$ and a negative-sequence term $e^{-j\theta}$. Moreover, the magnitude of the negative-sequence component is directly related to the degree of asymmetry between the x - and y -axis signals.

C. SRF-PLL Principle and Limitations

The structure of the SRF-PLL is shown in Fig. 3. It operates by applying a Park transformation to the input vector with respect to the estimated angle $\theta(t)$ and using the q-axis component as the phase error. In Fig. 3, HPF denotes a high-pass filter introduced to remove DC offsets and low-frequency drift components from the current signals before phase detection. The PI regulator updates the estimated angular frequency $\omega(t)$, while the phase integrator updates $\theta(t)$, such that the q-axis component converges to zero in steady state.

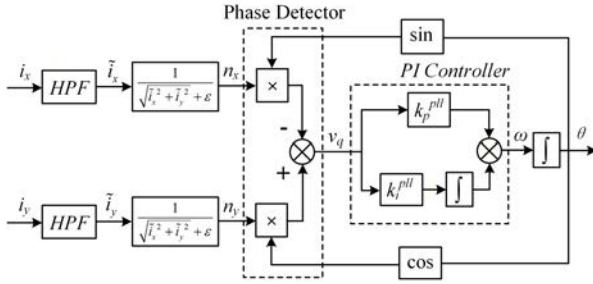


Fig. 3. Structure of the SRF-PLL.

When the observed signals contain amplitude mismatch or phase nonorthogonality, the fundamental component becomes a superposition of positive- and negative-sequence components, as shown in (6). In the vicinity of the locked state, the corresponding error signal can be written as (8):

$$e_{pll}(t) = v_q(t) = \text{Im}\{I_+ e^{j(\theta_a - \theta)} + I_- e^{-j(\theta_a + \theta)}\} \approx I_+ \tilde{\theta}(t) + \text{Im}\{I_- e^{-j2\theta_a(t)}\} \quad (8)$$

where $v_q(t)$ is the q-axis phase-detection error in the synchronous reference frame; θ_a and ω_a denote the actual phase angle and actual angular frequency, respectively. In this expression, the first term is the desired phase-locking error signal, whereas the second term is a periodic component at $2\omega_a$. Thus, the negative-sequence component does not become a DC term in the SRF, but appears as an oscillatory term whose amplitude is proportional to I_- .

In addition to the negative-sequence component, the actual current signals may also contain higher-order harmonics. To analyze the influence of these harmonics on the PLL, the input complex vector is uniformly expressed as (9):

$$\mathbf{i}(t) = I_1^+ e^{j\theta_a(t)} + \sum_{h=2}^H (I_h^+ e^{jh\theta_a(t)} + I_h^- e^{-jh\theta_a(t)}) \quad (9)$$

The fundamental negative-sequence component is temporarily neglected. After CT and taking the imaginary part, the phase-detection error is obtained as (10):

$$e_{pll}(t) = v_q(t) \approx I_1^+ \tilde{\theta}(t) + \sum_{h=2}^H (\text{Im}\{I_h^+ e^{j(h-1)\theta_a(t)}\} + \text{Im}\{I_h^- e^{-j(h+1)\theta_a(t)}\}) \quad (10)$$

It can be seen from this expression that the synchronous rotating transformation does not eliminate the harmonic components; it shifts their frequencies to $(h-1)\omega_a$ and $(h+1)\omega_a$, which correspond to the positive- and negative-sequence components of the h -th harmonic, respectively.

In summary, both negative-sequence components and higher-order harmonics disturb the SRF-based synchronization process. In the SRF-PLL, they introduce

ripple into the phase-detection error, which leads to speed ripple and phase error. Because the SRF-PI compensator has the same synchronous rotating-frame structure, it suffers from similar interference. Without sequence separation before SRF processing, these components generate oscillatory terms in the synchronous frame, distort the control error, and weaken the vibration-suppression performance. Although low-pass filtering or bandwidth reduction can partially mitigate these effects, they usually degrade dynamic performance.

III. MCCF-PLL-BASED SYNCHRONOUS ESTIMATION

To suppress the interference of negative-sequence components and harmonics on synchronous observation at the source, a CCF is introduced at the front end of the PLL in this section, thereby constructing the MCCF-PLL. The idea is to combine the two-channel signals in the same radial plane into a complex vector, and then exploit the rotational selectivity of the CCF in the complex plane to perform narrowband extraction and conditioning of the target synchronous component, such that the signal fed into the PLL is made as close as possible to a pure positive-sequence fundamental component. Fig. 4 illustrates the positive- and negative-sequence CCF structures for the n -th synchronous component.

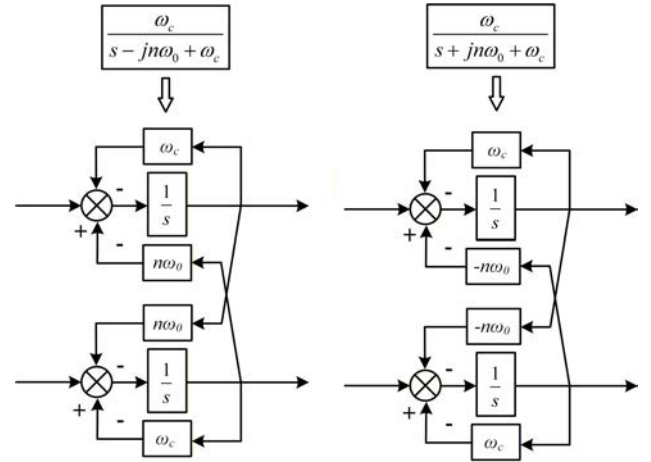


Fig. 4. Structure of the CCF.

By expressing the sequence output in complex form as $\mathbf{i}_{xyn\pm}(t) = i_{xn\pm}(t) + j i_{yn\pm}(t)$, the corresponding first-order transfer function in the complex domain can be derived as:

$$H_{n+}(s) = \frac{i_{n+}(s)}{i(s)} = \frac{\omega_c}{s - jn\omega_0 + \omega_c}, H_{n-}(s) = \frac{i_{n-}(s)}{i(s)} = \frac{\omega_c}{s + jn\omega_0 + \omega_c} \quad (11)$$

where ω_0 denotes the fundamental angular frequency, and ω_c denotes the CCF bandwidth. Its frequency response is shown in Fig. 5, where $n = 1$, $\omega_0 = 2 \times \pi \times 100$ rad/s, and $\omega_c = 0.2\omega_0$.

As can be seen from Fig. 5, H_{n+} attains its maximum response (0 dB) at $\Omega = n\omega_0$. In contrast, H_{n-} exhibits a passband around $-n\omega_0$. Therefore, the positive-sequence and negative-sequence CCFs constitute two band-pass extractors with the same center frequency but opposite rotational selectivity. In the vicinity of the locked frequency, they are able to direct the positive-sequence and negative-sequence components of the complex vector formed from the x - and y -axis signals into their respective channels.

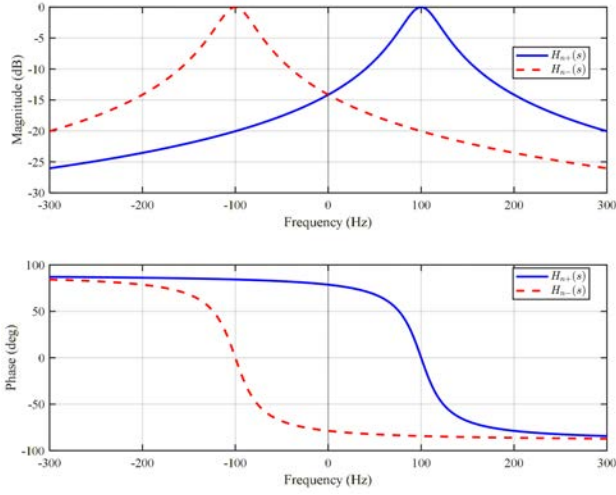


Fig. 5. Bode diagram of CCFs.

The frequency observation structure based on the MCCF-PLL is shown in Fig. 6. The x - and y -axis currents in the radial plane are selected as input signals, and the cases $n = 1$ and $n = 2$ are considered. If additional harmonic components are present, the corresponding CCF branches can be incorporated.

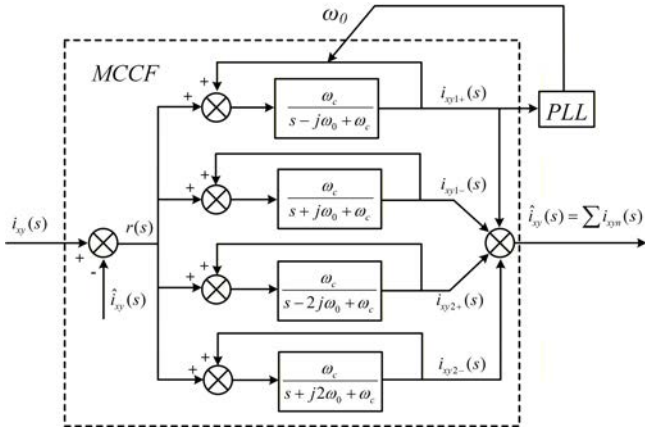


Fig. 6. Structure of the MCCF-PLL.

The reconstructed sum of all extracted synchronous components can be written as (12):

$$\hat{\mathbf{i}}_{xy}(t) = \sum_{n=1}^N (\hat{\mathbf{i}}_{xym+}(t) + \hat{\mathbf{i}}_{xym-}(t)) \quad (12)$$

where N denotes the number of extracted harmonic orders. The outer loop constructs the residual signal $r(t)$ through the summing node and feeds it into each extraction branch. Moreover, each CCF branch contains an internal summing node that feeds the output back to the input:

$$\mathbf{i}_{xym+}(s) = \frac{H_{n+}(s)}{1 - H_{n+}(s)} r(s) = \frac{\omega_c}{s - jn\omega_0} r(s) \quad (13)$$

This structure transforms the leaky first-order filter into a synchronous internal-model integrator with high gain, thereby providing stronger capability for capturing synchronous components and reducing the steady-state error.

Fig. 7 shows the overall frequency responses of the MCCF network, including the Bode diagrams from the input complex

vector $\mathbf{i}_{xy}$ to the reconstructed output and to the fundamental positive-sequence output. It can be seen that the overall network selectively preserves and reconstructs the target synchronous components while significantly attenuating undesired sequence components and harmonics.

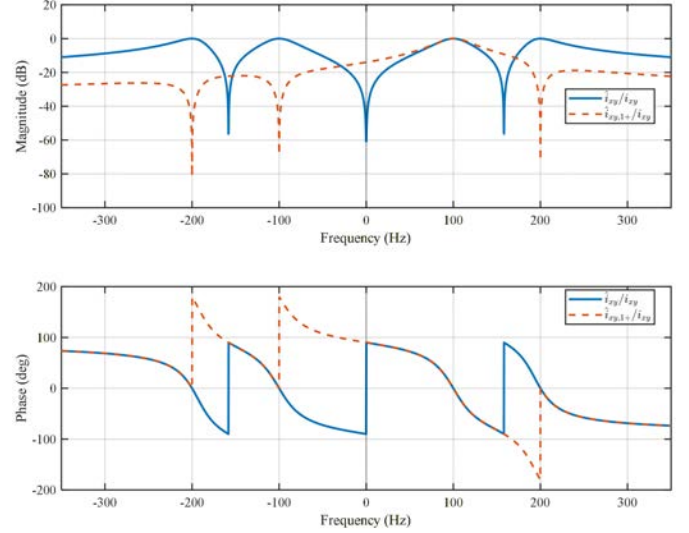


Fig. 7. Bode diagram of the MCCF.

Moreover, the MCCF effectively extracts the fundamental positive-sequence component and provides a cleaner synchronized input for the PLL. The PLL output $\omega(t)$ is fed back to the MCCF to update the center frequency and bandwidth ω_c of each branch, forming an adaptive closed-loop synchronous observation framework. In the original reference, the bandwidth is typically chosen as $\omega_c(t) = 0.707\omega_0(t)$. However, simulation and experimental results in this study indicate that, for AMB applications, $\omega_c(t) = 0.1\omega_0(t)$ or $\omega_c(t) = 0.2\omega_0(t)$ is more appropriate. Compared with grid-synchronization scenarios, AMB current signals generally contain stronger nonidealities, heavier harmonic and noise contamination, and more pronounced speed fluctuations. Hence, a smaller bandwidth is preferred to enhance robustness, at the expense of some dynamic response speed.

IV. POSITIVE-NEGATIVE-SEQUENCE SRF-PI-BASED SYNCHRONOUS VIBRATION SUPPRESSION

A. The Principle of Synchronous Vibration Suppression

To suppress synchronous displacement vibration, the structure shown in Fig. 8 combines MCCF-based sequence component extraction with positive-negative-sequence SRF-PI compensation. The outer position loop generates the radial displacement errors Δx and Δy , from which the MCCF extracts the positive- and negative-sequence synchronous components. With the synchronous angle $\theta(t)$ provided by the MCCF-PLL as the rotating reference, SRF-PI controllers are constructed in the corresponding SRFs. The controller outputs are then transformed back to the stationary reference frame using the compensation angles θ_c^+ , θ_c^- , and injected into the current-reference channel to suppress synchronous displacement vibration.

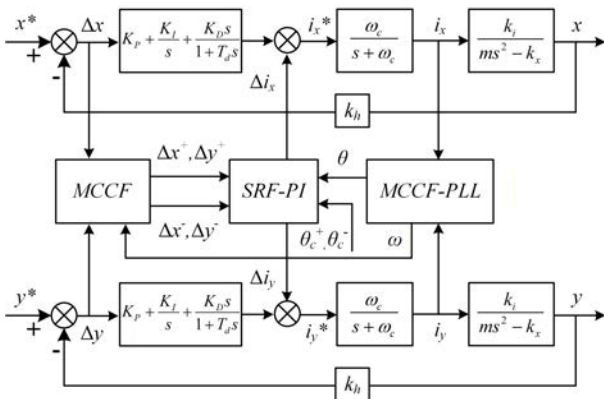


Fig. 8. Structure of the synchronous vibration suppression system.

The principle of vibration suppression is illustrated below by taking the fundamental positive-sequence component as an example. Assume that the displacement error is dominated by the synchronous component. After processing by the MCCF, the positive-sequence and negative-sequence components can be obtained and expressed in complex-vector form as (14):

$$\Delta \mathbf{z}^+(t) = \Delta x^+(t) + j\Delta y^+(t) \quad (14)$$

The separated Δx^+ and Δy^+ components are then fed into the positive-sequence SRF-PI structure shown in Fig. 9.

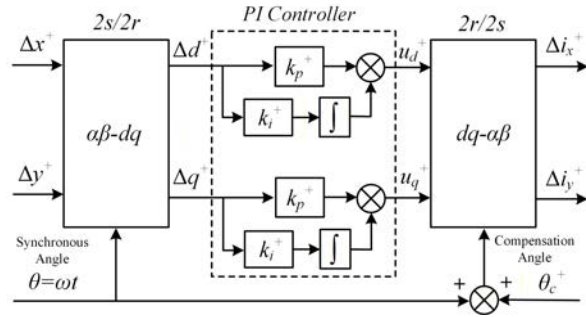


Fig. 9. Structure of the positive-sequence SRF-PI.

In the positive-sequence SRF-PI controller, the extracted positive-sequence error components are transformed into the positive-sequence synchronous rotating reference frame with respect to the angle $\theta(t)$:

$$\begin{bmatrix} \Delta d^+ \\ \Delta q^+ \end{bmatrix} = \begin{bmatrix} \cos\theta & \sin\theta \\ -\sin\theta & \cos\theta \end{bmatrix} \begin{bmatrix} \Delta x^+ \\ \Delta y^+ \end{bmatrix} \quad (15)$$

In the vicinity of the locked operating point, if the input is dominated by the target positive-sequence synchronous component, Δd^+ and Δq^+ become approximately constant in steady state. Therefore, a PI regulator can be directly employed to achieve zero steady-state error:

$$\begin{aligned} u_d^+(t) &= k_p^+ \Delta d^+(t) + k_i^+ \int \Delta d^+(t) dt \\ u_q^+(t) &= k_p^+ \Delta q^+(t) + k_i^+ \int \Delta q^+(t) dt \end{aligned} \quad (16)$$

Afterward, the controller output is mapped back to the stationary reference frame through the inverse transformation with a compensation angle θ_c^+ , thereby yielding the positive-sequence compensating current:

$$\begin{bmatrix} \Delta i_x^+ \\ \Delta i_y^+ \end{bmatrix} = \begin{bmatrix} \cos(\theta + \theta_c^+) & -\sin(\theta + \theta_c^+) \\ \sin(\theta + \theta_c^+) & \cos(\theta + \theta_c^+) \end{bmatrix} \begin{bmatrix} u_d^+ \\ u_q^+ \end{bmatrix} \quad (17)$$

Here, θ_c^+ is introduced to compensate for the phase lag caused by the controlled plant and the current loop at the synchronous frequency. By rotating the controller output current vector, the compensation current can be aligned with a more favorable phase direction for vibration suppression, thereby improving the compensation effectiveness and closed-loop stability margin. The compensation angle is placed in the inverse transformation rather than in the forward transformation because it acts on the controller output current vector instead of redefining the synchronous error. In this way, synchronous error regulation in (15)-(16) and output phase compensation in (17) remain clearly separated, facilitating subsequent stability analysis and parameter tuning.

The negative-sequence branch is constructed similarly, except that the SRF is based on $-\theta(t)$, because the negative-sequence component rotates in the opposite direction. Its compensation angle θ_c^- is tuned independently. For the n -th harmonic component, the structure can be extended in parallel using $n\theta(t)$ and $-n\theta(t)$, with θ_{cn}^+ and θ_{cn}^- tuned according to the frequency-response results.

Finally, the compensating currents generated by the positive-negative-sequence SRF-PI controllers are summed to obtain the final compensating current injected into the system:

$$\begin{aligned} \Delta i_x &= \Delta i_x^+ + \Delta i_x^- \\ \Delta i_y &= \Delta i_y^+ + \Delta i_y^- \end{aligned} \quad (18)$$

B. Stability Analysis

To analyze the local closed-loop stability of the SRF-PI synchronous compensation, the positive-sequence channel is considered. In the stability analysis, the MCCF is approximated as a unity-gain element, and the equivalent closed-loop plant from the positive-sequence compensating current Δi^+ to the positive-sequence displacement error $\Delta \mathbf{z}^+$ is defined as $R_+(s)$:

$$R_+(s) = \frac{G_1(s)G_p(s)}{1 + C_{PID}(s)G_1(s)G_p(s)} \quad (19)$$

The transfer function of the positive-sequence SRF-PI branch in the stationary frame can be written as (20):

$$H_+(s) = e^{j\theta_c^+} \left(k_p^+ + \frac{k_i^+}{s - j\omega} \right) \quad (20)$$

where k_p^+ and k_i^+ are the proportional and integral gains of the positive-sequence compensation, respectively. Accordingly, the closed-loop characteristic equation of the positive-sequence channel is given by (21):

$$1 + R_+(s)H_+(s) = 0 \quad (21)$$

Substituting (19) and (20) into the above equation yields:

$$(s - j\omega) + k_i^+ H_+(s) = 0 \quad (22)$$

where $H_+(s)$ is given by:

$$H_+(s) = \frac{k_i^+ e^{j\theta_c^+} R_+(s)}{1 + k_p^+ e^{j\theta_c^+} R_+(s)} \quad (23)$$

$H_+(s)$ denotes the complex-domain transfer function obtained by absorbing the proportional compensation action into the equivalent open-loop system. It can thus be seen that

θ_c^+ , k_p^+ , and k_i^+ all affect the distribution of the closed-loop poles in the complex plane. To investigate the local stability after the introduction of the synchronous compensation branch, let $k_i^+ = 0$ first. In this case, (22) reduces to $s - j\omega = 0$.

That is, when $k_i^+ = 0$, the closed-loop characteristic equation has a pole at $s = j\omega$. As k_i^+ increases gradually from zero, this pole starts from the vicinity of $j\omega$ and moves along the root locus. For the system to become stable in this local neighborhood, the initial direction of motion of this pole should point toward the left half of the complex plane. By treating the left-hand side of (22) as an implicit function and differentiating it with respect to k_i^+ , one obtains:

$$\left. \frac{ds}{dk_i^+} \right|_{s=j\omega, k_i^+=0} = -H_+(j\omega) \quad (24)$$

This indicates that the initial trajectory of the pole departing from $j\omega$ is determined by the complex quantity $-H_+(j\omega)$. For this pole to move toward the left half-plane, the real part of its initial velocity must be negative. Therefore, the following local stability condition must be satisfied:

$$\text{Re}[-H_+(j\omega)] < 0 \quad (25)$$

This condition can also be equivalently expressed in terms of the departure-angle condition:

$$-90^\circ < \arg \left(\frac{e^{j\theta_c^+} R_+(j\omega)}{1 + k_p^+ e^{j\theta_c^+} R_+(j\omega)} \right) < 90^\circ \quad (26)$$

This is the local stability criterion for the positive-sequence synchronous compensation branch. It can be seen that k_p^+ affects both the magnitude and phase of the equivalent plant through the denominator term of $H_+(j\omega)$. If k_p^+ is chosen improperly, the denominator may approach zero near the target synchronous frequency, making the above local stability condition more difficult to satisfy. Therefore, in engineering practice, $k_p^+ = 0$ is adopted in this study from the viewpoint of engineering implementation and stability. This is also because, in the synchronous rotating reference frame, the target synchronous error becomes approximately a DC quantity near the locked condition, and the integral term alone is sufficient to achieve zero steady-state error compensation.

Letting $R_+(j\omega) = |R_+(j\omega)|e^{j\phi_R}$, where ϕ_R denotes the phase of $R_+(j\omega)$, the stability criterion in (26) can then be rewritten as:

$$-90^\circ < \theta_c^+ + \phi_R < 90^\circ \quad (27)$$

To make the pole move as far as possible toward the left from the vicinity of $j\omega$, and obtain the maximum local damping, one may set $\theta_c^+ + \phi_R = 0$. Therefore, the theoretically optimal value of the positive-sequence compensation angle is given by:

$$\theta_c^+ = \theta_{c,th}^+ = -\phi_R \quad (28)$$

where $\theta_{c,th}^+$ denotes the theoretical optimal positive-sequence compensation angle. This result indicates that the positive-sequence compensation angle θ_c^+ should be chosen as the negative of the phase of the positive-sequence dynamic compliance function $R_+(s)$ at the target synchronous frequency.

For the AMB model considered in this study, the equivalent positive-sequence $R_+(s)$ can be expressed as:

$$R_+(s) = \frac{1.019 \times 10^5 s}{s^4 + 7854s^3 + 4.651 \times 10^6 s^2 + 8.072 \times 10^8 s + 8.15 \times 10^{10}} \quad (29)$$

The Bode diagram of $R_+(s)$ is shown in Fig. 10. At 100 Hz, ϕ_R is -132.03° , and thus the theoretical optimal compensation angle is obtained as $\theta_{c,th}^+ = 132.03^\circ$.

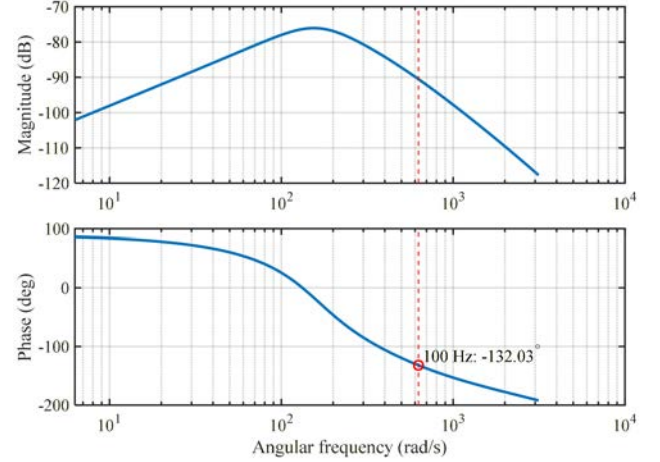


Fig. 10. Bode diagram of $R_+(s)$.

On this basis, by fixing $k_i^+ = 1 \times 10^6$ and sweeping the compensation angle θ_c^+ , the effect on the real part of the dominant closed-loop pole is obtained, as shown in Fig. 11.

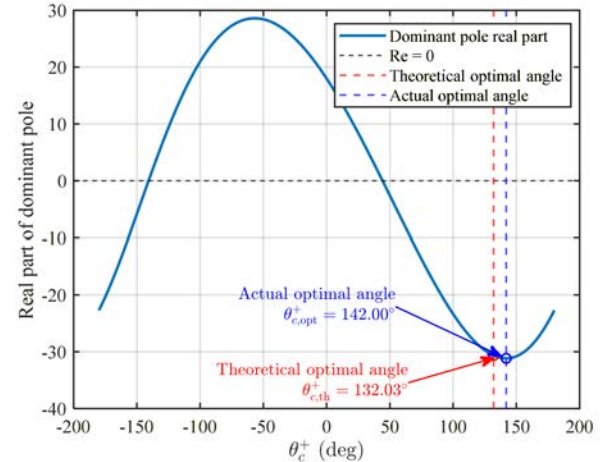


Fig. 11. Real part of the dominant pole versus θ_c^+ ($k_i^+ = 1 \times 10^6$).

It can be seen that the real part of the dominant pole reaches its minimum near $\theta_{c,opt}^+ = 142^\circ$, which exhibits a slight deviation from the theoretically optimal angle. This is because, when k_i^+ takes a relatively large value within the practical operating range, the influence of the synchronous compensation branch on the closed-loop pole distribution becomes more pronounced. Considering that, in experiments, the phase characteristic of $R_+(s)$ can be readily obtained through frequency-sweep testing, and that the deviation between $\theta_{c,opt}^+$ and $\theta_{c,th}^+$ is small, the system still maintains a satisfactory stability margin when the theoretically optimal angle is adopted. Therefore, for the analytical model considered in this study, $\theta_{c,th}^+$ is adopted as the reference compensation-angle θ_c^+ setting.

After the compensation angle $\theta_c^+ = 132.03^\circ$ is fixed, the influence of the integral gain k_i^+ on the closed-loop pole distribution is investigated. Fig. 12 shows the trajectories of all closed-loop poles as k_i^+ varies.

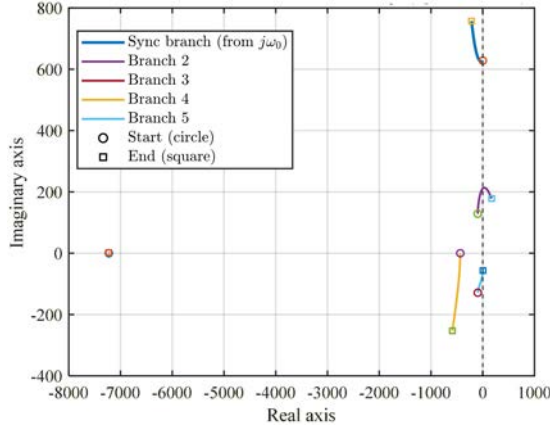


Fig. 12. Closed-loop pole trajectories with respect to k_i^+ .

It can be seen that, as k_i^+ increases, both the pole branches introduced by the synchronous compensation and the original closed-loop poles shift in the complex plane. When k_i^+ is too small, the compensation effect is limited and cannot provide sufficient vibration suppression. As k_i^+ continues to increase, the poles on Branches 3 and 5 gradually move toward the right side of the imaginary axis, indicating a reduction in stability margin. Therefore, an excessively large integral gain may lead to instability.

Fig. 13 further shows the variation of the maximum real part of the closed-loop poles with k_i^+ . When k_i^+ is small, the maximum real part remains close to zero but still lies in the left half-plane, indicating that the system is stable but the compensation capability is limited. As k_i^+ increases, the closed-loop poles first move further to the left. After k_i^+ exceeds a certain range, the maximum-real-part curve turns upward, indicating that the dominant pole shifts from the synchronous compensation pole to other original closed-loop poles of the system. Near $k_{i,crit}^+ = 3.0 \times 10^6$, the maximum real part becomes positive, indicating that a closed-loop pole crosses the imaginary axis into the right half-plane and the system tends to become unstable.

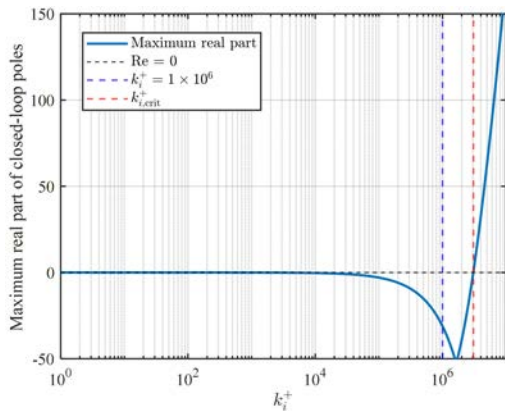


Fig. 13. Maximum real part of the closed-loop poles vs. k_i^+ .

Considering both synchronous vibration suppression capability and stability margin, the analytical study suggests

$\theta_c^+ = \theta_{c,th}^+ = 132.03^\circ$ and $k_i^+ = 1 \times 10^6$ as a feasible parameter pair for the simplified transfer-function model. In engineering tuning, the compensation angle θ_c^+ can first be determined from frequency-sweep identification, and then k_i^+ can be gradually increased from a small value to enhance the suppression capability. Since the influence of the synchronous compensation branch on the closed-loop pole distribution becomes more pronounced as k_i^+ increases, θ_c^+ may also be fine-tuned around the theoretical value $\theta_{c,th}^+$.

For the negative-sequence channel, the dynamic compliance function $R_-(s)$ can be defined analogously, and the compensation angle can be chosen as $\theta_c^- = \theta_{c,th}^- = -\arg[R_-(j\omega)]$. The harmonic channels can be extended in the same manner.

V. SIMULATION RESULTS

A simulation model is established in MATLAB/Simulink, and the parameters are listed in Table I. It should be noted that these parameters are not required to be identical to those in the preceding stability analysis. The stability analysis uses a simplified continuous-time transfer-function model to derive the compensation-angle selection rule and the local influence of the integral gain on pole migration, whereas the simulation is performed on a more detailed system model. Therefore, the controller parameters in Table I are re-tuned for the simulation model following the same design principle in Section IV.

TABLE I
SIMULATION PARAMETERS

Parameters	Magnitude
Mass of rotor, m/kg	10
Stiffness, $k_s/(\text{N} \cdot \text{m}^{-1})$	2.40×10^6
Stiffness, $k_f/(\text{N} \cdot \text{A}^{-1})$	129.7
Eccentricity, $e/(\mu\text{m})$	50
Bandwidth of current loop, $\omega_{cl}/(\text{rad} \cdot \text{s}^{-1})$	4000π
Bandwidth of MCCF, ω_c	$0.1 \times \omega_0$
Gain of SOGI, k_0	0.6
Gain of SOGI-FLL, τ	-60
Gain of PLL, k_p^{pll}	1329.7
Gain of PLL, k_i^{pll}	12279
Gain of SRF-PI, k_p^{pi}	0
Gain of SRF-PI, k_i^{pi}	2
Compensation angles, θ_c^+, θ_c^-	36°
Compensation angles, $\theta_{c2}^+, \theta_{c2}^-$	77°

The first part evaluates the frequency estimation performance of the MCCF-PLL in comparison with the PLL and SOGI-FLL methods. The second part presents the synchronous vibration suppression results based on positive-negative-sequence SRF-PI control.

A. Frequency Estimation Results based on MCCF-PLL

First, comparisons are conducted under standard orthogonality, where the rotational speed is set to 100 Hz, the injected signals in the x - and y -directions are mutually orthogonal, and $A_x = A_y = 2.0 \times 10^{-4}$. At $t = 0.02$ s, the three frequency estimation methods are activated. The initial value

of the SOGI-FLL method is set to 60 Hz, while those of PLL and MCCF-PLL are set to 0. As shown in Fig. 14, after a transient process, all three methods achieve convergence. In terms of convergence speed, the PLL is the fastest, followed by the MCCF-PLL, and the SOGI-FLL is the slowest.

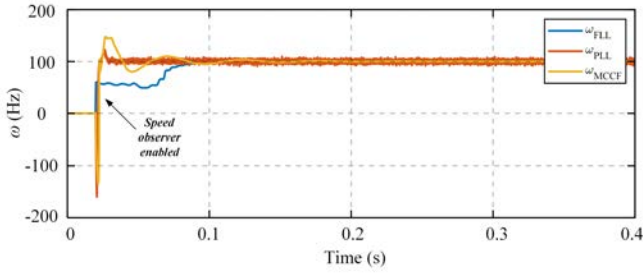


Fig. 14. Frequency estimation under ideal orthogonal conditions.

Verification is conducted under non-ideal operating conditions. Disturbances are injected with amplitudes $A_x = 2.0 \times 10^{-4}$, $A_y = 1.0 \times 10^{-4}$, $A_{x2} = 4 \times 10^{-5}$, and $A_{y2} = 2.0 \times 10^{-5}$, with the phase difference deviating from the ideal orthogonal condition by $\pi/8$, where A_{x2} and A_{y2} represent the amplitudes of the second harmonic components. Under such conditions, the x - and y -direction control currents exhibit severe distortion and contain harmonic components, corresponding to the non-ideal operating conditions of AMB systems, as shown in Fig. 15.

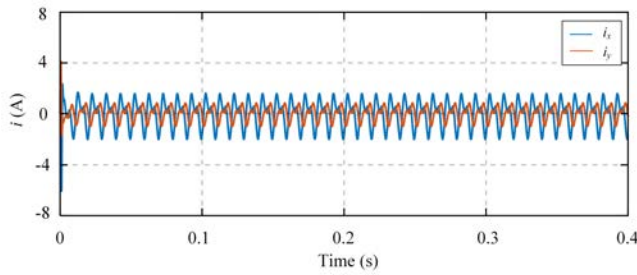


Fig. 15. x - and y -axis control currents under nonideal conditions.

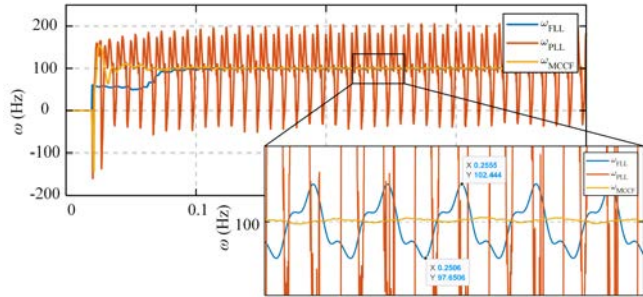


Fig. 16. Non-ideal conditions frequency estimation comparison.

In Fig. 16, the first subfigure presents the overall frequency estimation results of the different methods, while the second subfigure shows a local enlargement for a clearer comparison of the steady-state fluctuations. It can be seen that the traditional PLL method exhibits evident harmonic-related fluctuations, whereas the SOGI-FLL method still shows a fluctuation range of about ± 2.5 Hz. By contrast, the MCCF-PLL method maintains the estimation fluctuation within approximately ± 0.1 Hz, demonstrating better steady-state stability and higher frequency estimation accuracy under non-ideal operating conditions.

B. Synchronous Vibration Suppression Results based on Positive-negative-sequence SRF-PI

Following the compensation-angle selection principle in Section IV, the transfer functions $R_+(s)$ and $R_-(s)$ of the positive- and negative-sequence channels were identified for the simulation model. Since a second-harmonic suppression loop is included, identification was performed at 100 and 200 Hz, respectively:

$$R_+(j\omega)|_{\omega=2\pi \times 100} = R_-(j\omega)|_{\omega=2\pi \times 100} = 10.3 \angle -36^\circ \quad (30)$$

$$R_+(j\omega)|_{\omega=2\pi \times 200} = R_-(j\omega)|_{\omega=2\pi \times 200} = 6.5 \angle -77^\circ$$

The compensation angles were then selected accordingly:

$$\begin{aligned} \theta_c^+ &= \theta_c^- = 36^\circ \\ \theta_{c2}^+ &= \theta_{c2}^- = 77^\circ \end{aligned} \quad (31)$$

Simulations of displacement vibration were conducted under the same non-ideal operating conditions as in Part A. In the simulation, the positive-negative-sequence SRF-PI suppression was activated at $t = 0.2$ s, and the second-harmonic positive- and negative-sequence suppression channels were activated at $t = 0.5$ s. The x - and y -direction displacement waveforms are shown in Fig. 17, and the sequence components of displacement are shown in Fig. 18.

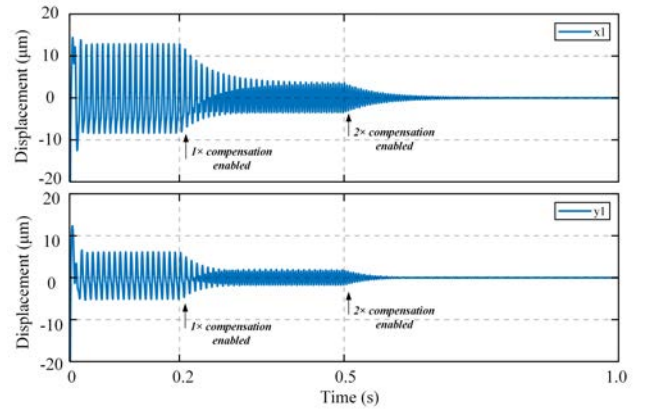


Fig. 17. x - and y -axis displacement responses.

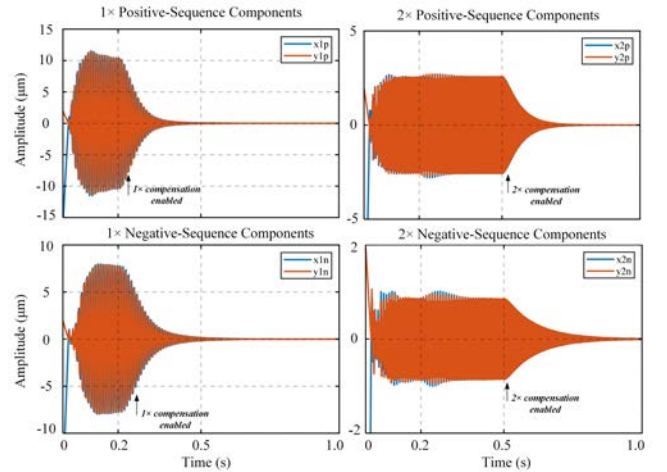


Fig. 18. Positive- and negative-sequence displacement components under fundamental and second-harmonic suppression.

After the fundamental positive- and negative-sequence suppression channels were activated at $t = 0.2$ s, the

corresponding fundamental components were significantly reduced. After the second-harmonic suppression channels were further activated at $t = 0.5$ s, the corresponding harmonic components were suppressed to near zero. These results confirm the effectiveness of the proposed method for selective suppression of synchronous vibration components under non-ideal operating conditions.

VI. EXPERIMENTAL RESULTS

To verify the effectiveness of the proposed synchronous estimation and vibration suppression strategy, experiments were carried out on an AMB test rig as shown in Fig. 19. The shaft is driven by a high-speed motor, and two touchdown bearings are employed to protect the AMBs. The control board acquires the displacement and current signals, executes the control algorithm, and generates pulse width modulation (PWM) commands for the power amplifier. The DC bus voltage is 70 V, and the AMB bias current is 5 A. The switching device is an insulated gate bipolar transistor (IGBT) module. The sampling and control frequencies are both 20 kHz, and the proposed synchronous estimation and vibration suppression algorithms can be executed within one control period.

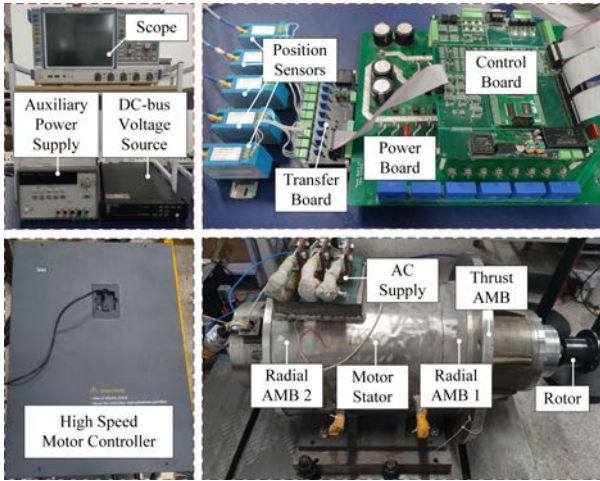


Fig. 19. Structure of AMB motor test rig.

The experimental parameters are listed in Table II. In practical experiments, the controller parameters were further tuned on the test rig to account for sampling and actuation delays, sensor noise, amplifier nonlinearity, and unmodeled dynamics, while preserving the same control structure and parameter-selection principle as those used in the simulation.

A. Constant-speed Experiments

Under the 3000 r/min operating condition, the dynamic synchronous-estimation results of the PLL, SOGI-FLL, MCCF-PLL, and MCCF-PLL-LPF methods are shown in Fig. 20. Fig. 20(a) presents the frequency-estimation responses, while Fig. 20(b) shows the corresponding synchronous phase outputs. Here, MCCF-PLL-LPF denotes the MCCF-PLL frequency estimate further processed by a 10 Hz low-pass filter. The enabling instant of each method is marked in the figure. It can be seen that the PLL exhibits the most severe

TABLE II
EXPERIMENTAL PARAMETERS

Parameter	Magnitude
Mass of rotor, m/kg	46
Stiffness, $k_s/(\text{N}\cdot\text{m}^{-1})$	2.60×10^6
Stiffness, $k_f/(\text{N}\cdot\text{A}^{-1})$	260
Rated motor power/kW	200
Rated motor speed/($\text{r}\cdot\text{min}^{-1}$)	18,000
Touchdown bearing air gap/ μm	250
AMB air gap/ μm	500
Bandwidth of MCCF, ω_c	$0.1 \times \omega_0$
Gain of SOGI, k_0	2
Gain of SOGI-FLL, τ	2000
Gain of PLL, k_p^{pll}	200.0
Gain of PLL, k_i^{pll}	4000.0
Gain of SRF-PI, k_p^{pi}	0
Gain of SRF-PI, k_i^{pi}	0.004

oscillation and fails to settle within the ± 2 Hz band, with a steady-state fluctuation standard deviation of 13.103 Hz. The SOGI-FLL achieves a relatively smooth steady-state estimate, but its settling time reaches 0.447 s. In contrast, the MCCF-PLL shows a significantly faster dynamic response, with a settling time of 0.272 s, although some steady-state fluctuation remains (0.327 Hz). After low-pass filtering, the MCCF-PLL-LPF further reduces the steady-state fluctuation to 0.071 Hz while preserving a fast settling time of 0.244 s. The phase results in Fig. 20(b) further indicate that the MCCF-PLL can provide continuous synchronous phase information for the subsequent SRF-PI compensation. These results show that the proposed MCCF-based speed observation scheme achieves a better trade-off between dynamic response and steady-state accuracy under non-ideal operating conditions.

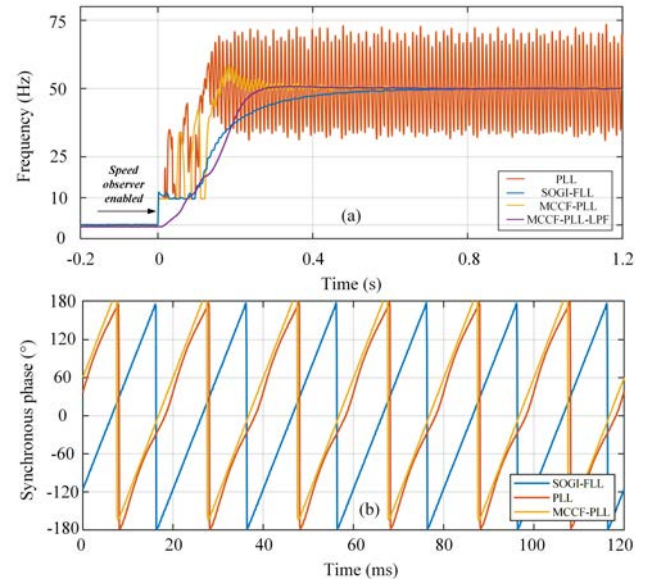


Fig. 20. Synchronous-estimation results of different speed-observation methods at 3000 r/min. (a) Frequency estimation. (b) Synchronous phase.

Based on the above speed-observation results, the sequence-domain dynamic characteristics are further identified to design the compensation angles for vibration

suppression. First, sequence-domain frequency-sweep identification is performed by injecting forward- and backward-rotating sequence excitations, through which the dynamic compliance functions of the positive-sequence and negative-sequence channels, $R_+(s)$ and $R_-(s)$, are obtained. The resulting phase-frequency responses are shown in Fig. 21.

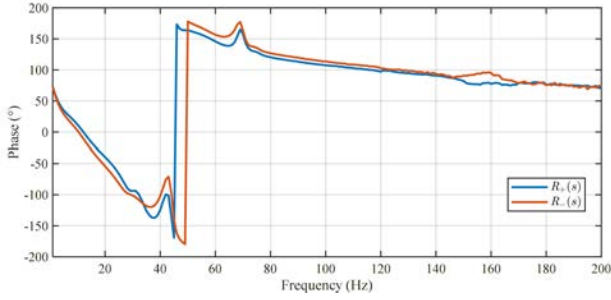


Fig. 21. Phase-frequency responses of the sequence-domain dynamic compliance functions.

The compensation angles are selected following $\theta_c^+ = -\arg[R_+(j\omega)]$, $\theta_c^- = -\arg[R_-(j\omega)]$. In the experiments, the dynamic compliance functions are obtained from sequence-domain frequency-sweep identification on the test rig. The corresponding values for 1800 r/min (30 Hz) and 3600 r/min (60 Hz) are listed in Table III.

TABLE III
COMPENSATION ANGLES USED IN THE EXPERIMENTS

Motor speed/(r·min ⁻¹)	θ_c^+	θ_c^-
1800	93.7°	101.2°
3600	-145.9°	-156.6°

Figs. 22 and 23 show the x - and y -direction displacement waveforms and the corresponding current waveforms under the 1800 and 3600 r/min operating conditions, respectively. In each figure, the left panel presents the x - and y -direction displacement responses, while the right panel shows the corresponding x - and y -direction current responses. The instant at which the synchronous vibration suppression algorithm is activated is also marked in the figures. After the suppression algorithm is enabled, the displacement amplitudes in both directions decrease significantly and converge to much lower levels. Meanwhile, the current oscillation amplitudes are also reduced. Under the 1800 r/min condition, the residual vibration in the suppressed displacement waveform is mainly attributed to the second-harmonic component rather than the target fundamental synchronous component.

To further quantify the suppression performance, Fig. 24 presents the spectral comparison of the x - and y -direction displacement signals within the initial 1 s and the final 1 s under the 1800 and 3600 r/min operating conditions, where the dashed, dash-dotted, and dotted lines indicate the $1\times$, $2\times$, and $3\times$ frequency locations, respectively. At 1800 r/min, the $1\times$ synchronous displacement amplitude is reduced from 29.206 to 0.015 μm in the x direction and from 74.327 to 0.151 μm in the y direction, corresponding to reductions of 99.95% and 99.80%, respectively. At 3600 r/min, the $1\times$ synchronous displacement amplitude is reduced from 67.419 to 3.748 μm in the x direction and from 22.494 to 0.694 μm in

the y direction, corresponding to reductions of 94.44% and 96.92%, respectively. These results demonstrate that the proposed method can effectively suppress the dominant synchronous displacement components under different rotational-speed conditions, thereby verifying its feasibility in practical AMB systems.

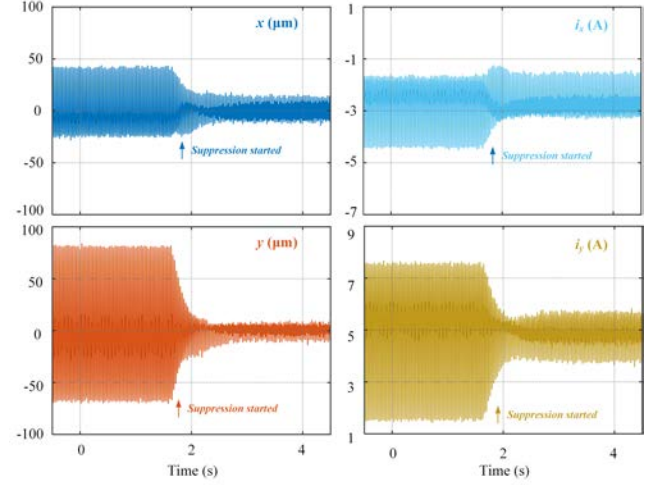


Fig. 22. x - and y -direction displacement and current waveforms at 1800 r/min.

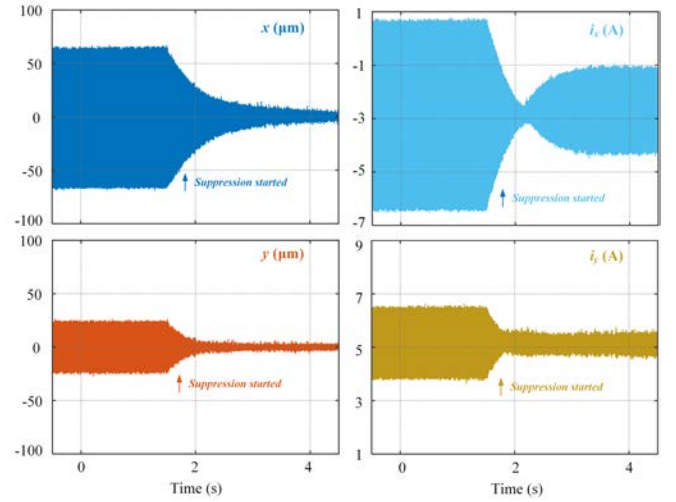


Fig. 23. x - and y -direction displacement and current waveforms at 3600 r/min.

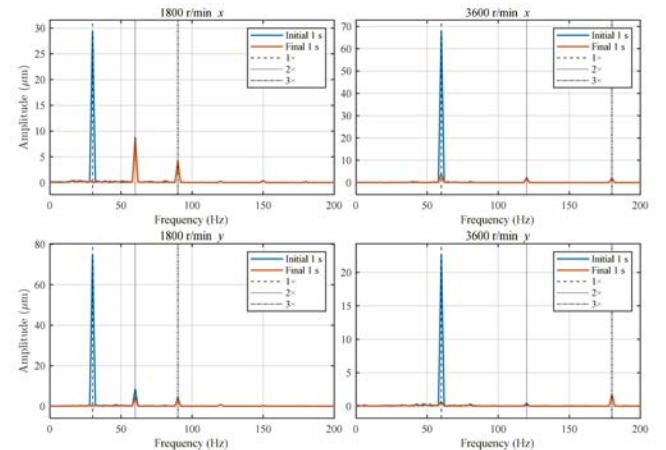


Fig. 24. Spectral comparison of x - and y -direction displacement signals in the initial 1 s and final 1 s at 1800 and 3600 r/min.

B. Variable-speed Experiments

To further verify the effectiveness of the proposed method under continuously varying rotational-speed conditions, variable-speed experiments were conducted during the run-up

from 900 to 3900 r/min (15–65 Hz) and the run-down from 3900 to 900 r/min (65–15 Hz), as shown in Figs. 25 and 26, respectively. In these experiments, the suppression branch was enabled before the run-up or run-down interval.

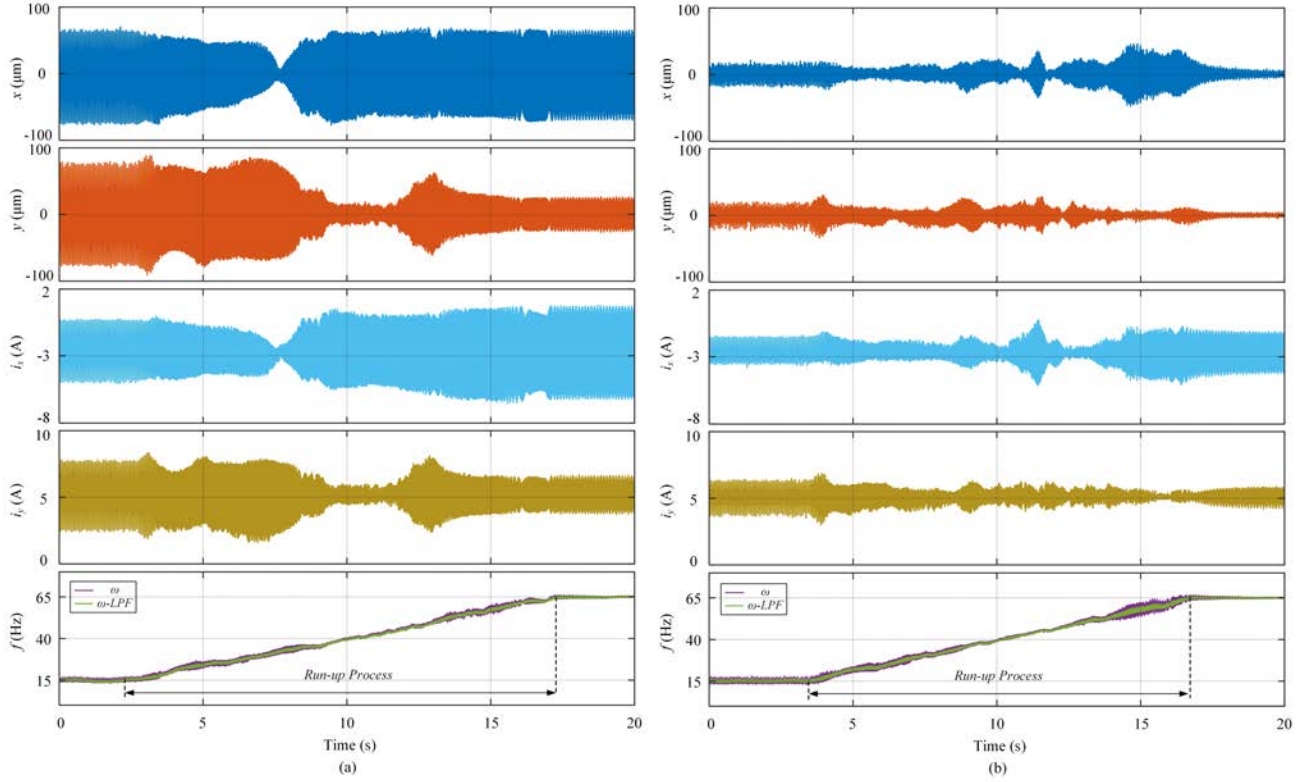


Fig. 25. Variable-speed vibration suppression results during the run-up from 900 to 3900 r/min (15–65 Hz). (a) Without suppression. (b) With suppression.

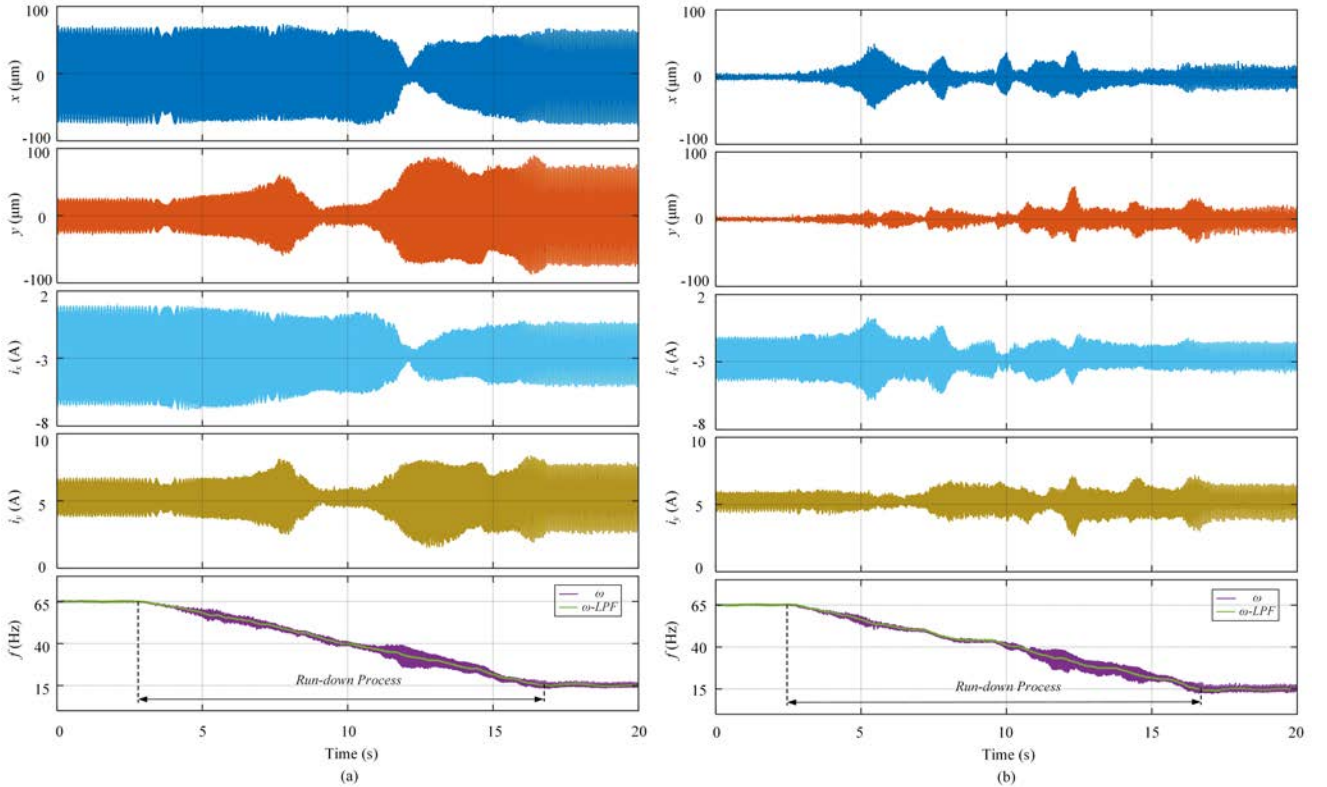


Fig. 26. Variable-speed vibration suppression results during the run-down from 3900 to 900 r/min (65–15 Hz). (a) Without suppression. (b) With suppression.

During variable-speed operation, the synchronous rotational speed was first estimated online by the proposed MCCF-PLL. Considering that the raw observed speed may contain instantaneous fluctuations, the MCCF-PLL-LPF defined above was used to obtain a smoother rotational-speed signal. Based on the compensation-angle design results shown in Fig. 21, the compensation angle was then updated online through table lookup and interpolation according to the low-pass-filtered observed speed. In this way, the compensation angle could vary continuously with the operating frequency, thereby maintaining proper phase compensation during the variable-speed process.

The experimental results in Figs. 25 and 26 present the x - and y -direction displacement waveforms, the corresponding current waveforms, and the observed rotational speed under the run-up and run-down conditions. Compared with the cases without suppression, the displacement amplitudes in both directions are effectively reduced when the suppression branch is enabled throughout the speed-varying process. Meanwhile, the observed speed and its low-pass-filtered result vary smoothly, indicating that the online compensation-angle updating strategy based on lookup-table interpolation remains effective during both acceleration and deceleration processes.

These results demonstrate that the proposed method is applicable not only under constant-speed conditions but also under continuous variable-speed conditions within the tested range. By combining low-pass-filtered speed observation with online compensation-angle interpolation, the proposed method maintains effective synchronous vibration suppression during both the run-up from 900 to 3900 r/min and the run-down from 3900 to 900 r/min.

In addition to the vibration-suppression performance, the real-time computational feasibility of the proposed method was also evaluated. The execution time was measured on the TMS320F28377D controller by using the internal CPU timer as a free-running counter. During the test, the synchronous estimation and vibration suppression branches were enabled, and 100 consecutive ADC interrupt periods were recorded. The measured execution time covers the main computational procedures in one control period, including current/displacement sampling, synchronous estimation, synchronous vibration suppression, PID/PI calculation, and PWM duty-cycle update. The average execution time of each module is summarized in Table IV.

TABLE IV
EXECUTION TIME OF THE MAIN PROGRAM MODULES

Program module	Average execution time/ μ s	Percentage of 50 μ s control period
Current/displacement sampling	0.96	1.92%
Synchronous estimation	6.02	12.04%
Synchronous vibration suppression	3.55	7.10%
Control loop calculation and PWM duty cycle update	6.56	13.12%
Other auxiliary procedures	0.79	1.58%
Total execution time	17.88	35.76%

As shown in Table IV, the total average execution time of the ADC interrupt program is 17.88 μ s, which accounts for

only 35.76% of the 50 μ s control period. Therefore, the proposed MCCF-PLL-based synchronous estimation and positive-negative-sequence SRF-PI vibration suppression algorithm can be completed within one control period with sufficient real-time margin. This result further verifies the feasibility of implementing the proposed method on the adopted DSP-based AMB control platform.

Both the constant-speed and variable-speed experiments show that the proposed method can achieve effective synchronous vibration suppression without necessarily increasing the total control current. This is because the total control current is the vector sum of the PID feedback current and the injected compensating current. Once the vibration is suppressed, the required feedback current is reduced. In addition, since the two current components are not always in phase, their superposition does not necessarily increase the total current amplitude. Therefore, with a properly selected compensation phase, better vibration suppression can be achieved with a reasonable current response.

VII. CONCLUSIONS

This paper proposes a sensorless synchronous vibration suppression method for AMB systems based on MCCF-PLL and positive-negative-sequence SRF-PI control. The main conclusions are as follows.

1) The proposed MCCF-PLL can extract the fundamental positive-sequence current component under non-ideal signal conditions, thereby improving synchronous speed and phase observation. Compared with conventional PLL and SOGI-FLL methods, it achieves a better balance between dynamic response and steady-state accuracy.

2) By combining MCCF-based sequence decomposition with positive-negative-sequence SRF-PI compensation, the proposed method realizes sequence-selective suppression of synchronous displacement vibration. The compensation angles are designed according to sequence-domain dynamic compliance analysis and can be updated online during variable-speed operation.

3) Simulation results verify the effectiveness of the proposed method under amplitude mismatch, phase nonorthogonality, and harmonic contamination. Experimental results further show that effective synchronous vibration suppression is achieved at 1800 and 3600 r/min, and the method remains effective during the run-up from 900 to 3900 r/min and the run-down from 3900 to 900 r/min.

Overall, these results verify the feasibility and engineering effectiveness of the proposed method for practical AMB systems. Future work will focus on experimentally extending the method to higher-order harmonic vibration suppression and validating its performance under higher-speed operating conditions.

REFERENCES

- [1] G. Schweitzer, and E. H. Maslen, *Magnetic Bearings: Theory, Design, and Application to Rotating Machinery*. Berlin, Germany: Springer, 2009.

- [2] X. J. Li, A. Palazzolo, and Z. Y. Wang, "A Combination 5-DOF Active Magnetic Bearing for Energy Storage Flywheels," *IEEE Transactions on Transportation Electrification*, vol. 7, no. 4, pp. 2344–2355, Dec. 2021.
- [3] R. Herzog, P. Buhler, and C. Gahler *et al.*, "Unbalance Compensation Using Generalized Notch Filters in the Multivariable Feedback of Magnetic Bearings," *IEEE Transactions on Control Systems Technology*, vol. 4, no. 5, pp. 580–586, Sept. 1996.
- [4] Y. B. Zheng, N. Mo, and Y. Zhou *et al.*, "Unbalance Compensation and Automatic Balance of Active Magnetic Bearing Rotor System by Using Iterative Learning Control," *IEEE Access*, vol. 7, pp. 122613–122625, Aug. 2019.
- [5] G. Liu, J. L. Li, and S. Q. Zheng *et al.*, "Suppression of Synchronous Current Using Double Input Improved Adaptive Notch Filter Algorithm," *IEEE Transactions on Industrial Electronics*, vol. 67, no. 10, pp. 8599–8607, Oct. 2020.
- [6] H. W. Xu, J. Li, and Y. Lu *et al.*, "Unbalance Control for High-speed Active Magnetic Bearing Systems without Speed Sensors," *IEEE Transactions on Transportation Electrification*, vol. 10, no. 4, pp. 10147–10157, Dec. 2024.
- [7] R. D. Yang, W. Z. Pei, and Y. Q. Yang *et al.*, "A Novel Method for Rotational Speed Observation of Magnetically Suspended High Speed Motor Using Quadrature Phase-locked-loop," *IEEE Sensors Journal*, vol. 22, no. 9, pp. 8604–8613, May 2022.
- [8] L. Peng, Z. C. Fu, and T. Xiao *et al.*, "An Improved Dual Second-order Generalized Integrator Phased-locked Loop Strategy for an Inverter of Flexible High-voltage Direct Current Transmission Systems Under Nonideal Grid Conditions," *Processes*, vol. 11, no. 9, pp. 2634, Sept. 2023.
- [9] C. Xu, X. S. Tang, and G. J. Liu, "A Novel Method for Frequency Estimation of AMB Rotor based on Multiresonance Controller-quadrature Phase-locked Loop," *IEEE Transactions on Industrial Electronics*, vol. 72, no. 2, pp. 1849–1859, Feb. 2025.
- [10] P. Rodríguez, A. Luna, and I. Candela *et al.*, "Multiresonant Frequency-locked Loop for Grid Synchronization of Power Converters Under Distorted Grid Conditions," *IEEE Transactions on Industrial Electronics*, vol. 58, no. 1, pp. 127–138, Jan. 2011.
- [11] H. B. Sun, D. Jiang, and J. C. Yang, "Synchronous Vibration Suppression of Magnetic Bearing Systems without Angular Sensors," *CES Transactions on Electrical Machines and Systems*, vol. 5, no. 1, pp. 70–77, Mar. 2021.
- [12] H. T. Wu, J. Zhou, and Y. Zhang *et al.*, "Adaptive Auto Balancing of Magnetically Suspended Rotor based on Second Order Generalized Integrator-frequency Locked Loop," *Proceedings of the CSEE*, vol. 41, no. 4, pp. 1505–1514, Feb. 2021.
- [13] P. L. Cui, Y. B. Li, and J. L. Li *et al.*, "Synchronous Vibration Force Suppression of Magnetically Suspended CMG based on Modified Double SOGI-FLL," *IEEE Transactions on Industrial Electronics*, vol. 70, no. 11, pp. 11566–11575, Nov. 2023.
- [14] H. Zeng, X. B. Xu, and J. F. Li *et al.*, "A Modified Feedforward Repetitive Controller for AMB Rotor System Using Adaptive Frequency Estimator," *IEEE Transactions on Power Electronics*, vol. 40, no. 10, pp. 14722–14732, Oct. 2025.
- [15] J. L. Li, G. Liu, and P. L. Cui *et al.*, "Suppression of Harmonic Vibration in AMB-rotor Systems Using Double-input Adaptive Frequency Estimator," *IEEE Transactions on Industrial Electronics*, vol. 69, no. 3, pp. 2986–2995, Mar. 2022.
- [16] S. Q. Zheng, Q. Chen, and H. L. Ren, "Active Balancing Control of AMB-rotor Systems Using a Phase-shift Notch Filter Connected in Parallel Mode," *IEEE Transactions on Industrial Electronics*, vol. 63, no. 6, pp. 3777–3785, Jun. 2016.
- [17] B. Liu, J. C. Fang, and G. Liu *et al.*, "Unbalance Vibration Control and Experiment Research of Magnetically Suspended Flywheels," *Journal of Mechanical Engineering*, vol. 46, no. 12, pp. 188–194, Jun. 2010.
- [18] C. Mao, and C. S. Zhu, "Unbalance Compensation for Active Magnetic Bearing Rotor System Using a Variable Step Size Real-time Iterative Seeking Algorithm," *IEEE Transactions on Industrial Electronics*, vol. 65, no. 5, pp. 4177–4186, May 2018.
- [19] B. C. Han, Y. Liu, and S. Q. Zheng, "Research on Vibration Suppression for Magnetic Suspension Motor based on Repetitive Control Method," *Journal of Vibration, Measurement & Diagnosis*, vol. 35, no. 3, pp. 486–492, Jun. 2015.
- [20] T. Schuhmann, W. Hofmann, and R. Werner, "Improving Operational Performance of Active Magnetic Bearings Using Kalman Filter and State Feedback Control," *IEEE Transactions on Industrial Electronics*, vol. 59, no. 2, pp. 821–829, Feb. 2012.
- [21] C. Bi, D. Z. Wu, and Q. Jiang *et al.*, "Automatic Learning Control for Unbalance Compensation in Active Magnetic Bearings," *IEEE Transactions on Magnetics*, vol. 41, no. 7, pp. 2270–2280, Jul. 2005.
- [22] H. Gao, L. X. Xu, and Y. L. Zhu, "Unbalance Vibratory Displacement Compensation for Active Magnetic Bearings," *Chinese Journal of Mechanical Engineering*, vol. 26, no. 1, pp. 95–103, Feb. 2013.
- [23] S. Q. Zheng, B. C. Han, and R. Feng *et al.*, "Vibration Suppression Control for AMB-supported Motor Driveline System Using Synchronous Rotating Frame Transformation," *IEEE Transactions on Industrial Electronics*, vol. 62, no. 9, pp. 5700–5708, Sept. 2015.
- [24] H. T. Wu, J. Zhou, and L. Ji, "Unbalance Compensation of Magnetically Suspended Rotor based on Single Phase Coordinate Transformation," *Journal of Zhejiang University (Engineering Science)*, vol. 54, no. 5, pp. 963–971, May 2020.
- [25] X. Q. Guo, W. Y. Wu, and Z. Chen, "Multiple-complex Coefficient-filter-based Phase-locked Loop and Synchronization Technique for Three-phase Grid-interfaced Converters in Distributed Utility Networks," *IEEE Transactions on Industrial Electronics*, vol. 58, no. 4, pp. 1194–1204, Apr. 2011.
- [26] Z. Y. Yan, Y. D. Feng, and B. D. Wang *et al.*, "A Double Fundamental Frequency PLL without Cascaded SOGI Only based on SAI Under Nonideal Grid Conditions," *IEEE Transactions on Power Electronics*, vol. 41, no. 5, pp. 7684–7693, May 2026.



Tianci Sun (S'23) was born in Shandong, China, in 2001. He received the B.S. degree in electrical engineering and automation from Chongqing University, Chongqing, China, in 2023, and the M.S. degree in electrical engineering from Huazhong University of Science and Technology, Wuhan, China, in 2026. His

current research interests include synchronous-vibration suppression and surge disturbance suppression in active magnetic bearing systems.



Dong Jiang (S'05-M'12-SM'16) received B.S. and M.S. degrees in electrical engineering from Tsinghua University, Beijing, China, in 2005 and 2007, respectively. He began his Ph.D. study at the Center for Power Electronics Systems (CPES) at Virginia Tech, USA, in 2007 and was transferred to the University of

Tennessee, USA, with his advisor in 2010. He received his Ph.D. degree in power electronics and motor drives from the University of Tennessee in Dec. 2011. He was with United Technologies Research Center (UTRC) in Connecticut, USA, as a Senior Research Scientist/Engineer from Jan 2012 to July 2015. He has been with Huazhong University of Science & Technology (HUST) in China as a professor since July 2015. Dong Jiang's major research area is power electronics and motor drives, with more than 100 published IEEE journal and conference papers in this area. He has four best paper awards in IEEE conferences. He is an associate editor of IEEE Transactions on Industry Applications.