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IEEE论文导读：如何改善低解析度霍尔元件的位置估测性能，应用于永磁同步电机矢量控制系统！

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An Improved Rotor Position Estimation With Vector-Tracking Observer in PMSM Drives With Low-Resolution Hall-Effect Sensors

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Abstract—This paper presents an improved approach for estimating high-resolution rotor position in permanent-magnet synchronous motor (PMSM) drives with low-resolution Hall-effect sensors. A vector-tracking position observer in conjunction with discrete Hall sensors' output signals has been proposed, which is similar to a phase-locked loop structure. It consists of a position error detector, based on the vector cross product of the unit back-electromotive-force vectors obtained from a stator electrical model, and a proportional–integral-typed controller to make the position error rapidly converge to zero. This structure does not only compensate the misalignment effect of Hall sensors but also enhance their transient operating capability. The effectiveness of the proposed approach has been verified from several experiments in a low-voltage PMSM drive for automotive applications.

Index Terms—Hall-effect sensors, high-resolution position estimation, permanent-magnet synchronous motor (PMSM), vector-tracking observer.

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Recently, some approaches to compensate the misalignment effect of Hall sensors have been introduced in [15]–[17]. The method in [15] has presented an automated process that can obtain the actual Hall state transition points in which an iterative routine determines the offset between ideal and actual points, and the resultant actual state transition values are stored in a lookup table during only initial commissioning. Since its main algorithm is based on the average speed, the performance may degrade at a variable-speed operation.

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In [16], a vector-tracking observer using the vector-cross-product phase-detection method has been demonstrated, which uses the quantized rotating position vector at 60° obtained from Hall sensors' signals. In [17], an improved method of [16] is presented to decouple the intrinsic harmonic term in the quantized rotating position vector. These methods have zero-lag tracking capability. However, since the observer is based on the mechanical model of a machine, their position estimation may be easily affected by the system inertia and load variations.

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As a practical compromise, Hall sensors are often employed since they require little cost and volume compared with shaft-mounted sensors and provide discrete absolute-position information with electrically $\pm 30^\circ$ resolution. From discrete sensors' information, the high-resolution position can be estimated through some signal processing or error correction techniques.

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A. Position Estimation With Average Rotor Speed

The high-resolution rotor position can be easily estimated through a method based on average rotor speed [13]–[16], where the six sectors are classified according to the states of the Hall sensor's signals. Assuming that the rotor speed within a sector is constant and the average speed in a current and previous sector is uniform, the rotor speed can be approximated as follows:

$$\omega_h = \frac{\frac{\pi}{3}}{\Delta t} \tag{1}$$

where Δt is the time interval of the previous sector. Then,

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where Δt is the time interval of the previous sector. Then, the rotor position is estimated by the following numerical integration:

$$\boxed{\widehat{\theta}_r = \theta_s + n \cdot \omega_h \cdot T_s \qquad (2)}$$

where θ_s is the measured absolute position in the current sector, n is the number of integration steps during the time interval of the current sector, and T_s is the sampling time. The rotor position $\widehat{\theta}_r$ satisfies $\theta_s \le \widehat{\theta}_r \le \theta_s + (\pi/3)$.

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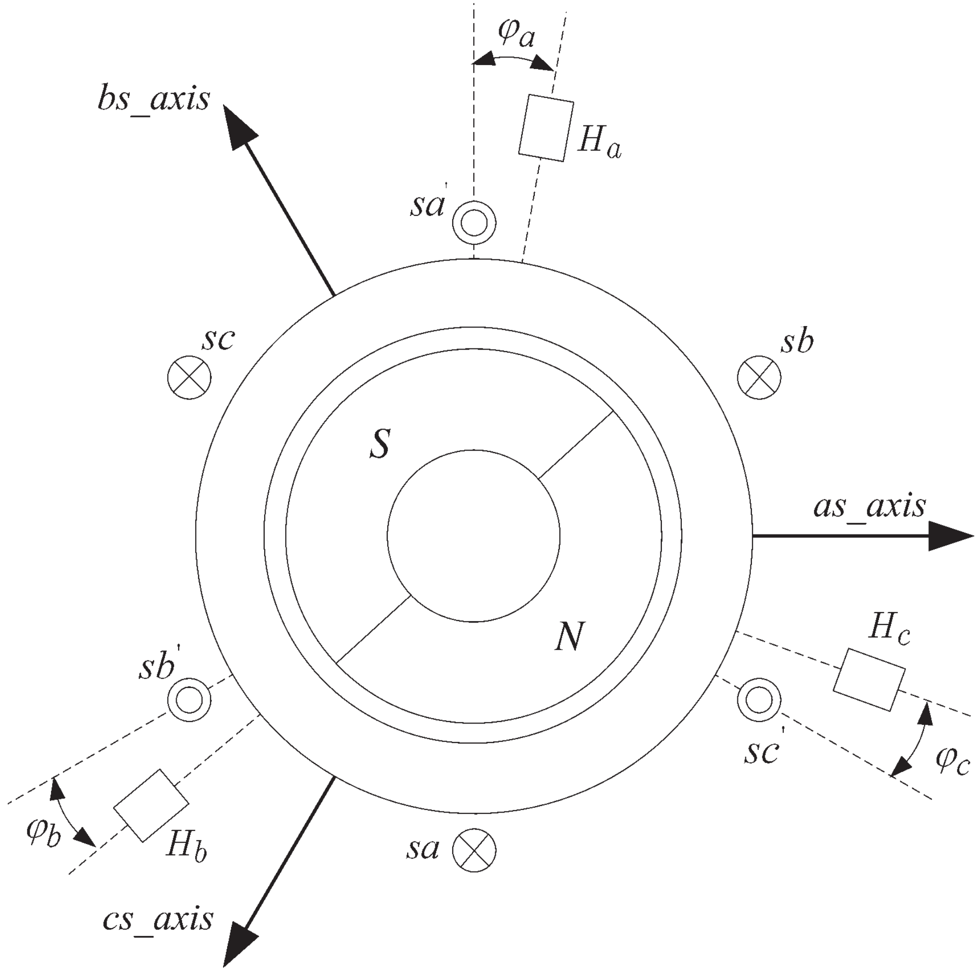

Fig. 1. Two-pole machine with misaligned Hall-effect sensors.

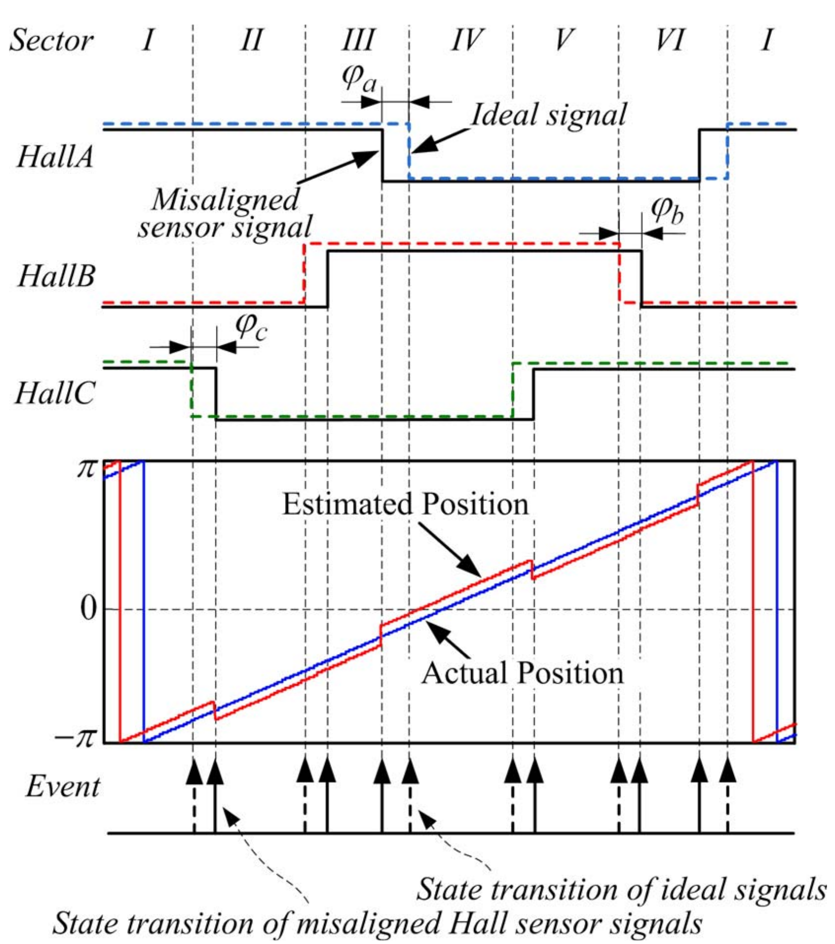

Fig. 2. Output signals of the misaligned Hall sensors and resultant position estimate.

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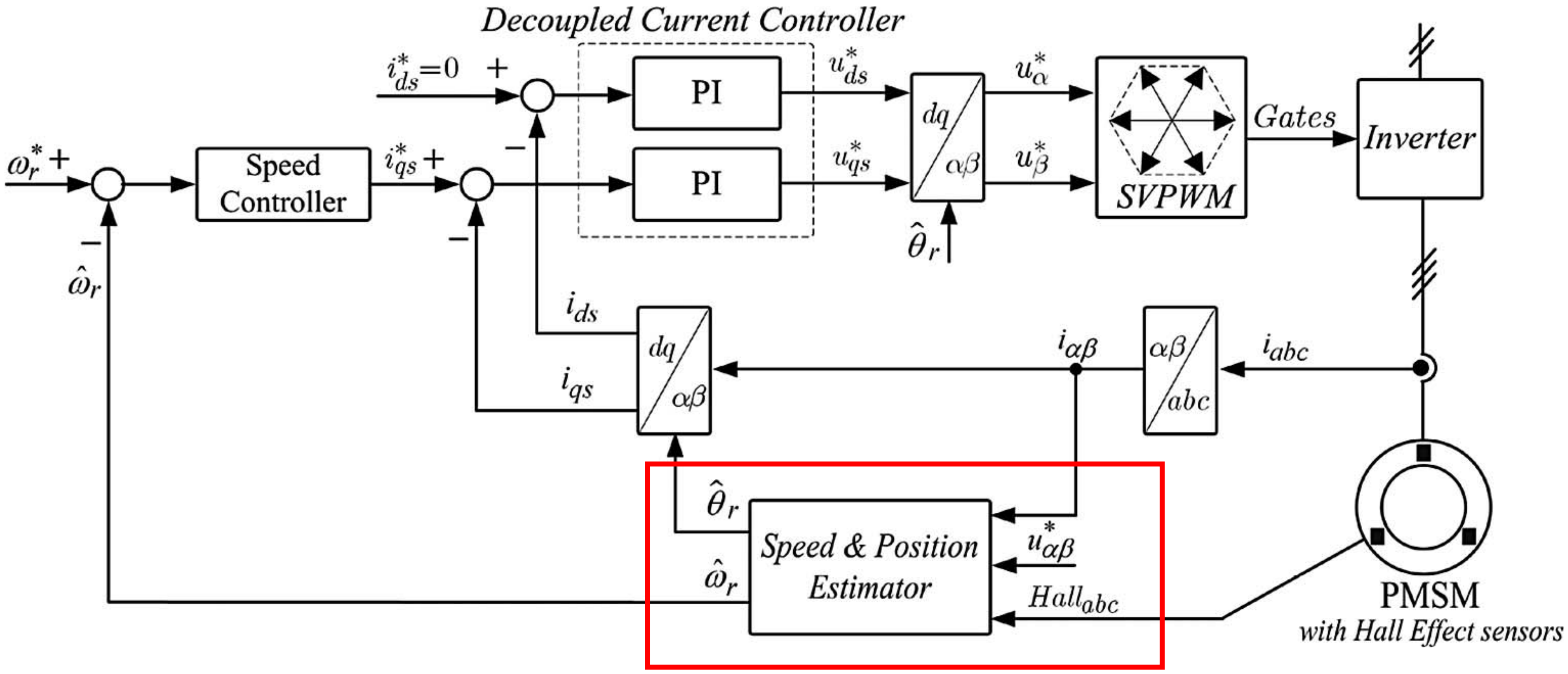

Fig. 3. Overall control scheme of the PMSM drive with Hall-effect sensors.

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In the case of the PMSM vector drives with Hall sensors, the general speed and decoupled current control loop in the synchronous reference frame with the position and speed estimator can be configured as shown in Fig. 3. The model equation in the stationary reference frame can be derived as follows:

$$\begin{bmatrix} u_\alpha^* \\ u_\beta^* \end{bmatrix} = \begin{bmatrix} R_s + pL_s & 0 \\ 0 & R_s + pL_s \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \hat{\omega}_r \lambda_m \begin{bmatrix} -\sin \widehat{\theta}_r \\ \cos \widehat{\theta}_r \end{bmatrix} \tag{3}$$

$$\hat{\omega}_r = \omega_h + \omega_{\mathrm{corr}} \tag{4}$$

where $u_{\alpha\beta}^*$ and $i_{\alpha\beta}$ represent the reference stator voltages and the measured stator currents, respectively. p is a derivative operator. R_s, L_s, and λ_m denote the stator resistance, the stator inductance, and the rotor flux linkage, respectively. $\hat{\omega}_r$ is the estimated rotor angular speed, including the average rotor speed ω_h, which is feedforwardly added to the proposed observer. ω_{corr} represents speed correction for the intrinsic error in ω_h.

Considering a digital control system with enough high sampling frequency, (3) can be expressed in discrete form. If the estimated rotor position, speed, and model parameters are the same as the actual values, the following equation can be obtained:

$$\boxed{\begin{bmatrix} (u_\alpha^* - R_s i_\alpha) - L_s \frac{i_\alpha(k) - i_\alpha(k-1)}{T_s} \\ \left(u_\beta^* - R_s i_\beta\right) - L_s \frac{i_\beta(k) - i_\beta(k-1)}{T_s} \end{bmatrix}} = \boxed{\hat{\omega}_r \lambda_m \begin{bmatrix} -\sin \widehat{\theta}_r \\ \cos \widehat{\theta}_r \end{bmatrix}} \tag{5}$$

where k is the sampling instant. The right side of (5) represents the estimate $\widehat{\overline{E}}$ of the back EMF established by the permanent magnet of the rotor. The left side represents the reference back EMF $\overline{E}^*$ calculated from the stator electrical circuits, where $\overline{E}^*$ can be filtered through a programmable low-pass filter (LPF), as represented in [18], to minimize the noise component without gain loss and phase delay. Then, the filtered reference back-

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gain loss and phase delay. Then, the filtered reference back-EMF vector is represented as

$$\overline{E}_f^* = \frac{\overline{E}^*}{\tau s + 1}(1 + j\tau\widehat{\omega}_r) \tag{6}$$

where τ is the time constant, which can be adjusted to the rotor speed $\hat{\omega}_r$.

The reference unit back EMF of $\overline{E}^*$ can be obtained as

$$\overline{e}^* = \frac{\overline{E}_f^*}{\left|\overline{E}_f^*\right|} = \begin{bmatrix} -\sin\theta_r^* \\ \cos\theta_r^* \end{bmatrix} \tag{7}$$

where θ_r^* is the phase angle of $\overline{e}^*$, which reflects the actual rotor position. Meanwhile, the estimated unit back EMF of $\widehat{\overline{E}}$ is equal to $\widehat{\overline{e}} = [-\sin\widehat{\theta}_r \ \cos\widehat{\theta}_r]^T$, which can be directly calculated from the estimated rotor position.

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If the estimated rotor position and speed are the same as the actual values, $\overline{e}^*$ and $\hat{\overline{e}}$ are in-phase with each other. However, these may be out of phase in practice due to the estimation errors of speed/position and the parameter mismatches. Assuming that the difference of θ_r^* and $\widehat{\theta}_r$ is reasonably small, the position estimation error can be approximately detected from the following cross product of the unit back-EMF vectors $\overline{e}^*$ and $\hat{\overline{e}}$:

$$\begin{aligned} \|\overline{e}^* \times \widehat{\overline{e}}\| &= -\sin\theta_r^* \cos\widehat{\theta}_r + \cos\theta_r^* \sin\widehat{\theta}_r \\ &= \sin\left(\widehat{\theta}_r - \theta_r^*\right) \simeq -\theta_{\mathrm{err}} \end{aligned} \tag{8}$$

where θ_{err} $(= \theta_r^* - \widehat{\theta}_r)$ is the position estimation error. Then,

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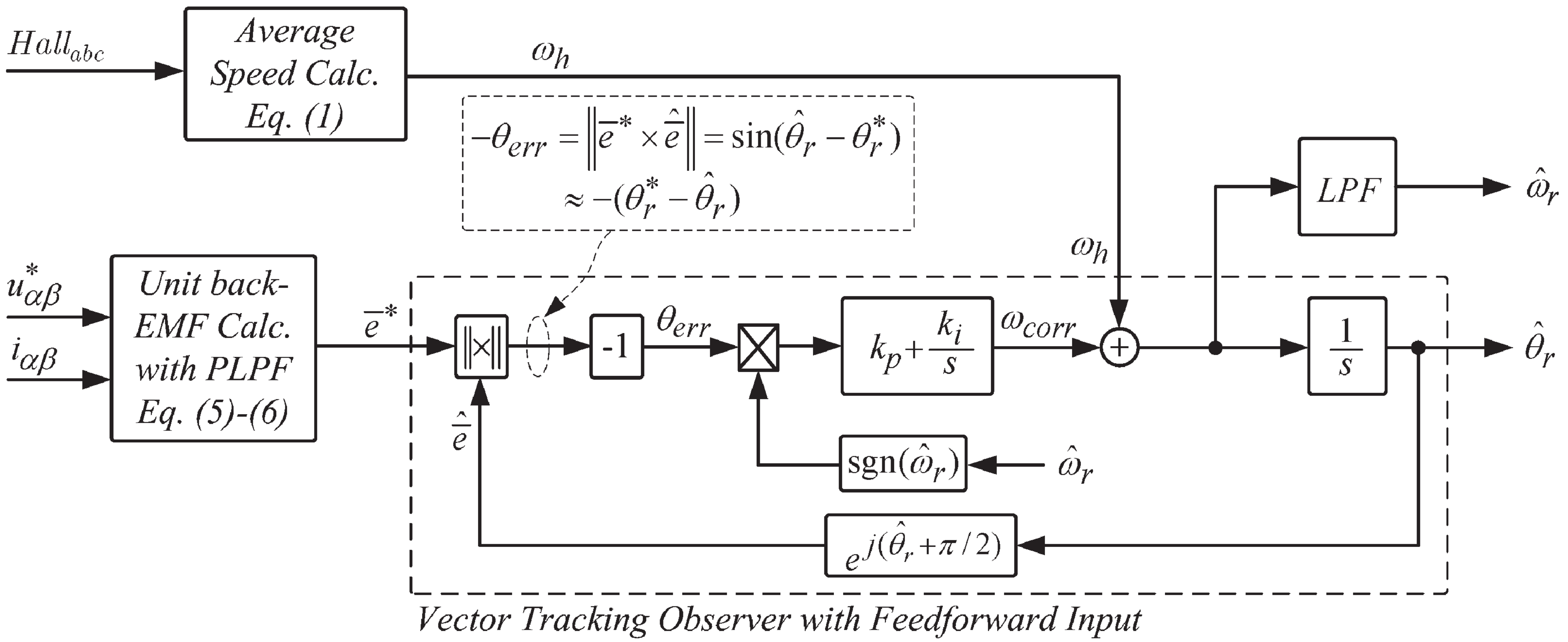

Fig. 4. Proposed vector-tracking position observer.

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where θ_{err} $(= \theta_r^* - \widehat{\theta}_r)$ is the position estimation error. Then, the speed correction ω_{corr} shown in (4) can be obtained by the following PI-typed controller as a loop filter of the proposed estimator so that the resultant value of (8) becomes zero:

$$\omega_{\text{corr}} = -\left(k_p + \frac{k_i \cdot T_s}{1 + z^{-1}}\right) \cdot \left[\|\overline{e}^* \times \widehat{\overline{e}}\| \cdot \text{sgn}(\widehat{\omega}_r)\right] \qquad (9)$$

where the sign function $\text{sgn}(\widehat{\omega}_r)$ is to reflect the rotating direction of the rotor. k_p and k_i represent the proportional and integral gains, respectively.

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In (9), ω_{corr} represents a speed correction corresponding to the position estimation error, i.e., the phase difference between the reference and estimated unit back-EMF vectors. From (4), the average rotor speed ω_h, including the intrinsic speed error, is appropriately corrected, and then, the estimated rotor speed $\hat{\omega}_r$ is converged to the actual rotor speed. Therefore, the high-resolution rotor position can be estimated by the following simple integration:

$$\widehat{\theta}_r(k) = \widehat{\theta}_r(k-1) + T_s\widehat{\omega}_r(k) \tag{10}$$

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C. Performance Evaluation of the Proposed Approach

The proposed vector-tracking position estimator described in (7)–(10) is shown in Fig. 4. In practical use, the estimated speed $\hat{\omega}_r$ has to be filtered through an LPF to enhance the noise immunity as shown in Fig. 4. The vector-tracking observer loop with feedforward input for the average speed ω_h can be equivalently transformed to a second-order linear model as shown in Fig. 5. From Fig. 5(b), the transfer function for the estimated position can be derived as

$$\widehat{\theta}_r = \frac{k_p s + k_i}{s^2 + k_p s + k_i}\theta_r^* + \frac{s^2}{s^2 + k_p s + k_i}\theta_h \qquad (11)$$

where θ_h represents the corresponding position to the average speed ω_h, which depicts the position feedforward input from the Hall sensors' signals.

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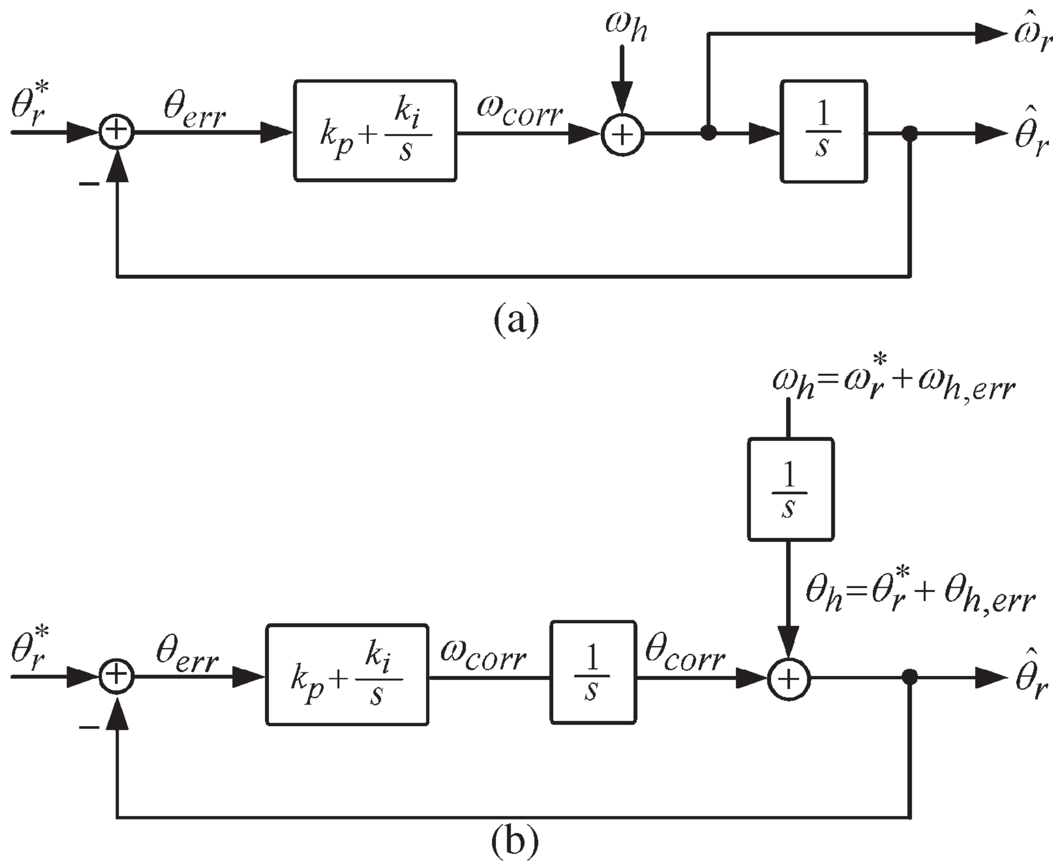

Fig. 5. Equivalent linear model of the vector-tracking observer with feedforward input. (a) Equivalent linear model. (b) Equivalent model of (a).

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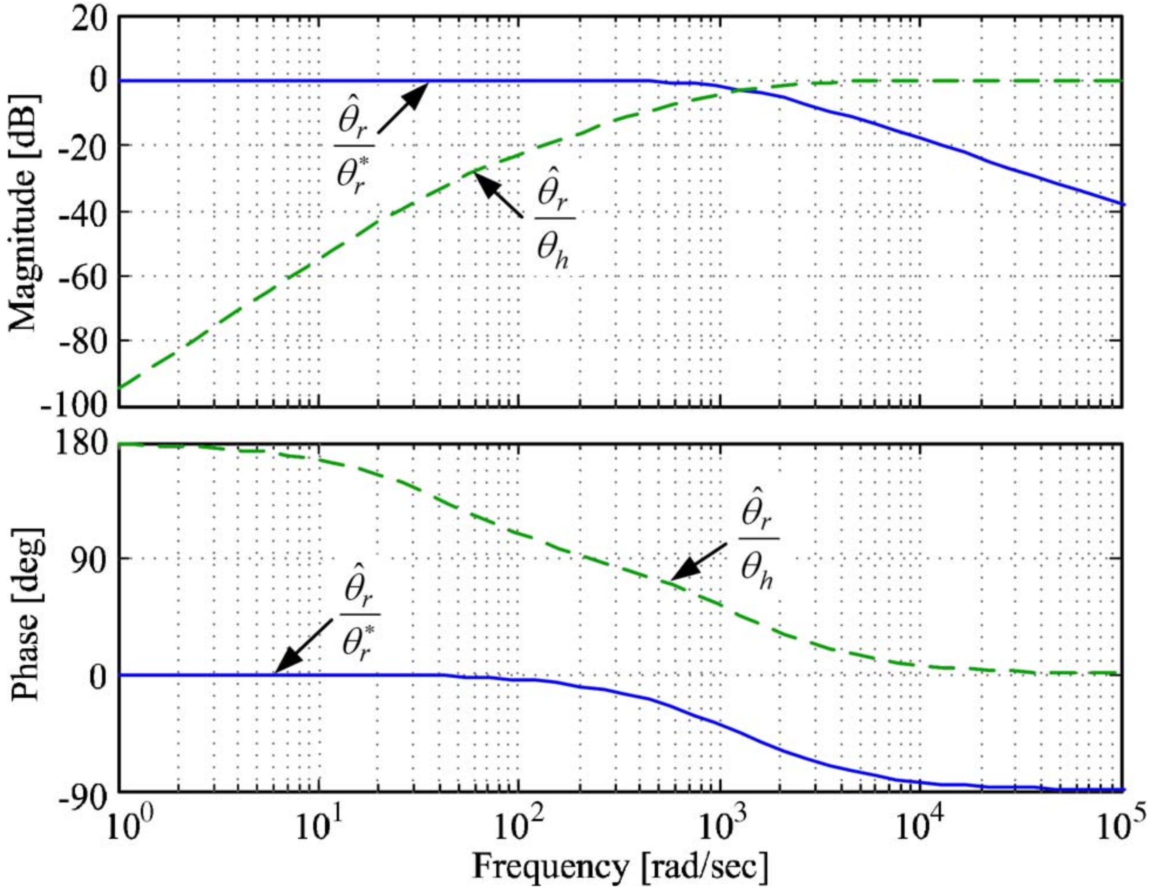

Fig. 6. Frequency response of the proposed observer ($k_p = 1268$ and $k_i = 54\,289$).

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D. Improvement in the Load Change Condition

In pulsewidth-modulation (PWM) voltage-source inverter drives, the errors between the reference voltages and actual voltages of the machine tend to increase as the motor load increases due to the inverter nonlinearity effects [20], [21]. In (5) and (6), it is shown that the reference voltages, instead of the actual voltages, are used to estimate the back-EMF vector. Therefore, the back-EMF estimation error increases according to the motor load. This leads to an error in the speed correction ω_{corr} by (9) and then causes an error in the position estimate $\widehat{\theta}_r$ inevitably.

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The instantaneous load torques can be used to compensate this error, where the load torques can be calculated in the reference and estimated synchronous frames shown in Fig. 7. The reference and estimated torques T_e^* and $\widehat{T}_e$ are respectively represented as

$$T_e^* = K_T i_{qs}^* \tag{13}$$

$$\widehat{T}_e = K_T i_{qs} \tag{14}$$

where K_T is the torque constant. i_{qs}^* represents the reference q-axis current generated by the speed controller. i_{qs} implies the estimated q-axis current, which is the transformed value with respect to the estimated position, i.e., $i_{qs} = i_\beta \cos\widehat{\theta}_r - i_\alpha \sin\widehat{\theta}_r$.

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As can be seen in Fig. 7, if the load angle δ, which implies the position error induced by the motor load, is a very small value, (13) can be transformed as

$$T_e^* = K_T(i_{ds}\sin\delta + i_{qs}\cos\delta)$$

$$\simeq K_T(i_{ds}\delta + i_{qs}) \tag{15}$$

where i_{ds} $(= i_\alpha \cos\widehat{\theta}_r + i_\beta \sin\widehat{\theta}_r)$ represents the estimated d-axis current.

From (14) and (15), the instantaneous torque error is given as

$$T_e^* - \widehat{T}_e = K_T i_{ds}\delta. \tag{16}$$

Therefore

$$i_{ds}\delta = i_{qs}^* - i_{qs}. \tag{17}$$

In (17), it is clear that the right side $(i_{qs}^* - i_{qs})$ reflects the information of the load angle δ. Therefore, the load angle can be estimated by the following PI-typed controller:

$$\widehat{\delta} = \left(k_{pc} + \frac{k_{ic}\cdot T_s}{1+z^{-1}}\right)(i_{qs}^* - i_{qs}) \tag{18}$$

where $\widehat{\delta}$ is the estimated load angle. k_{pc} and k_{ic} are the proportional and integral gains, respectively.

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The total position estimation error can be considered as the sum of the position error of (8) and the estimated load angle of (18). As in (19), if the estimated load angle of (18) is added to the speed correction of (9), the proposed vector-tracking observer may enhance the position estimation performance by correcting the load-induced error

$$\omega_{\text{corr}} = -\left(k_p + \frac{k_i \cdot T_s}{1 + z^{-1}}\right) \cdot \left[\|\overline{e}^* \times \widehat{\overline{e}}\| \cdot \text{sgn}(\widehat{\omega}_r) \boxed{- \widehat{\delta}}\right]. \quad (19)$$