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Linear Active Disturbance Rejection Controllers for PMSM Speed Regulation System Considering the Speed Measurement Noise

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Abstract—In the permanent magnet synchronous motor (PMSM) speed regulation system, the dynamic performance of the conventional active disturbance rejection control (ADRC) system will be deteriorated by using a low bandwidth speed filter. To solve this problem, two ADRC controllers considering the speed measurement noise are proposed in this paper. One proposed ADRC system is based on the extended state observer (ESO), the other proposed ADRC system is based on phase-locking loop observer (PLLO). In the proposed two ADRC systems, an integrator is used as the speed filter so that the measured position can be directly used for observing the speed and disturbance without speed calculation. Meanwhile, by using an integrator as the speed filter, the dynamic performance of the proposed ESO-based ADRC system is not affected since it is independent of the speed filter, and the proposed PLLO-based ADRC system has a better rejection ability for the low-frequency disturbance. Experimental results validate the proposed methods.

Index Terms—Active disturbance rejection control, extended state observer, measurement noise, phase-locking loop observer.

I. INTRODUCTION

AMONG many topologies of electric machines [1], surface-mounted permanent magnet synchronous motor (PMSM) has been widely used in servo applications because of its advantageous features including high efficiency, high-power density, large torque-to-inertia ratio, low noise, and free maintenance. Due to the various external disturbance, internal parameter uncertainties, and some unmodelled dynamics, high performance control of servo system, i.e., good tracking performance, excellent load-torque rejection ability, nice noise suppression property, and strong robustness to model uncertainties, are very challenging [2].

In speed control systems, good speed-reference tracking and excellent load-torque rejection ability cannot be obtained simultaneously by using a conventional one-degree-of-freedom controller. Therefore, many two-degree-of-freedom (TDOF) robust control methods are proposed, such as TDOF PI controller [3] and disturbance/uncertainty estimation and attenuation (DUEA) techniques [4]. Among many DUEA techniques, disturbance observer (DOB) and extended state observer (ESO) are most extensively investigated and applied. DOB was initiated by Ohnishi and his colleagues in 1980s to estimate the external disturbance of servo systems [5]. ESO is generally regarded as an essential part of the so-

called active disturbance rejection control (ADRC), which was first proposed by Han [6]. In the ADRC system, the lumped disturbance including both the external disturbance and the internal disturbance caused by the parameter variations or unmodelled dynamics is extended as a new state, which is estimated by ESO. Therefore, the ADRC provides insight for common methods to deal with both linear and nonlinear systems, time-varying and time-invariant systems. Nevertheless, to ensure the required performance by the feedforward compensation, the precise modeling of the plant system should be indispensable; a more precise plant model provides higher compensation in the feedforward manner as well as a more progressive design in the feedback controller [7].

Though ADRC and its variants have been tried in many fields [8], they have a notable drawback, complicated control structure with many tuning parameters [9]. To solve this problem, linear ADRC (LADRC) was introduced in [10], where only the bandwidth of the observer and the closed-loop system are tuned. However, considering the practical issues of LADRC, the well-known superiority of ADRC over the classical controllers is completely subjected to sensor specifications [11]. For example, it is a common practice to use a position sensor, such as the incremental encoder or resolver, to measure the rotor position, and the speed is calculated by the differentiation of the mechanical angle. Due to the quantization error caused by the position sensor, the speed is contaminated by the measurement noise [12]. In experiments, it is quite common to use a low-pass filter (LPF) to filter out the high-frequency components of noise. Compared with the first order LPF, the phase-locking loop (PLL) based LPF is more widely used due to its less phase lag [13]. However, the LPF causes a phase lag between the filtered speed and actual speed, which is suitable for systems where high dynamics is not required [14]. In [15], a first order LPF is adopted for suppressing the measurement noise, an ESO is constructed considering the LPF. The simulation results show that the ADRC system based on the ESO with LPF performs better than the conventional ADRC system under the existence of the measurement noise. The same technique is also applied successfully to the voltage control of the permanent magnet synchronous generator [16]. In [17], the effect of the LPF is analyzed. It is found that the bandwidth of the observer is restricted by the time constant of the LPF for consideration of stability. In [18], the LPF is used for filtering out the noise in the observation error instead of the output feedback. Although

the design principle of the LPF is given, it has not been applied in the motor control system.

The alternative method to suppress the measurement noise is to use the instantaneous speed observer (ISO), which is based on the Luenberger observer and proposed by R. D. Lorenz [19]. By using this technique, zero-phase lag between the observed speed and actual speed can be achieved [20] at the expense of knowing mechanical parameters, especially the moment of inertia [21]. Fortunately, many mechanical parameter identification methods have been proposed [22], [23]. As analyzed in [12], the speed measurement noise changes with the speed. The frequency of the measurement noise varies from zero to a specific value that decided by the sampling time and the pulses per round, periodically. Therefore, the bandwidth of the instantaneous speed observer has to be reduced in the lower speed range, leads to a deteriorated dynamic performance. To solve this problem, one method is to use the high-gain observer with nonlinear gain [24], [25], which however complicates the parameter setting process. Other methods include the use of two cascaded state filters [26]-[29]. In [26], two state filters are adopted, one is used for observing the instantaneous speed, while the other is used for observing the accelerator based on the observed instantaneous speed. By feedback the observed speed and accelerator, the measurement noise can be greatly suppressed, hence it is possible to use high-gain feedback controller to improve the system disturbance rejection ability. In [27] and [28], the second state filter is used for observing the disturbance torque. It is proved in [29] that the ADRC system using two cascaded observers performs better disturbance rejection property than the conventional ADRC system. Nevertheless, the two state filters complicate the algorithm. Moreover, the dynamic performance of different state filters has not been compared. As is known from [30], though the same second order observer is used, the ADRC system based on the phase-locking loop observer (PLL) and the one based on ESO have a totally different disturbance rejection property and noise suppression performance.

To sum up, despite the existence of valuable works concerning the measurement noise suppression of ADRC system, some practical issues in speed regulation system have not been properly addressed yet. The contribution of this paper can be concluded as

- 1) A general LPF is built to regard the integrator as a special case of LPF, and the effect of the LPF in the conventional ADRC system and modified ADRC system is analyzed.
- 2) The ESO and PLL considering the speed filter are proposed to have a good observation of the speed and disturbance.
- 3) The two proposed ADRC systems are compared with the conventional ADRC system and the modified ADRC system, some useful conclusions are recommended to be carefully considered by experts and practitioners of ADRC.

The remaining paper is organized as follows. In Section II, the effect of the LPF in conventional ADRC system and the modified ADRC system is analyzed. In Section III, two ADRC systems considering the LPF are proposed, and the effect of bandwidth of the LPF is analyzed. Meanwhile, the dynamic

performances of different ADRC systems are compared. In Section IV, experimental results are presented and analyzed. Finally, conclusions are drawn in Section V.

II. CONVENTIONAL ADRC SPEED REGULATION SYSTEM

The motion equation of the PMSM system can be given as

$$\dot{\Omega} = \frac{-B\Omega + T_e - T_L}{J} \quad (1)$$

where Ω is the mechanical angular velocity, B is the viscous friction torque coefficient, J is the moment of inertia, and T_L is the load torque.

A. Feedback Control Law

The feedback control law is designed to obtain the proper control quantity, i.e., the torque reference, so that the speed reference tracking error can attenuate as desired. As a result, the motion equation should be modified as a function of the torque reference T_e^*

$$\dot{\Omega} = \frac{T_e^* - (T_e - T_e + B\Omega + T_L)}{J} = bT_e^* + d \quad (2)$$

where $b=1/J$ is the control gain, $d=-(T_e^*-T_e+B\Omega+T_L)/J$ is the total disturbance caused by the electromagnetic torque tracking error, the viscous friction torque, and the load torque.

Defining the reference of mechanical angular velocity as Ω^* , then the tracking error of mechanical angular velocity can be expressed as $e_s = \Omega^* - \Omega$ and we have

$$\dot{e}_s = \dot{\Omega}^* - \dot{\Omega} = \dot{\Omega}^* - bT_e^* - d \quad (3)$$

Adopting a linear feedback control law

$$\dot{e}_s = -k_{ps} e_s \quad (4)$$

where k_{ps} is the proportional gain.

Combining (3) and (4) yields

$$T_e^* = \frac{\dot{\Omega}^* + k_{ps}(\Omega^* - \Omega) - d}{b} \quad (5)$$

In (5), the mechanical angular velocity Ω and the disturbance d are generally unknown and usually substituted by their estimated value $\hat{\Omega}$ and $\hat{d}$, thus the control quantity of the system is modified as

$$T_e^* = \frac{\dot{\Omega}^* + k_{ps}(\Omega^* - \hat{\Omega}) - \hat{d}}{b} \quad (6)$$

Considering that the actual system cannot provide infinite control output, saturation limit is usually applied as follows.

$$T_{esat}^* = \begin{cases} T_{e\max}^* \text{sgn}(T_e^*), & |T_e^*| > T_{e\max}^* \\ T_e^*, & |T_e^*| \leq T_{e\max}^* \end{cases} \quad (7)$$

where $T_{e\max}^*$ is the torque limit.

Without considering the saturation of the torque, the output speed when using the modified control quantity can be expressed as

$$\Omega(s) = \Omega^*(s) - \frac{k_{ps} [\hat{\Omega}(s) - \Omega(s)] + [\hat{d}(s) - d(s)]}{s + k_{ps}} \quad (8)$$

It shows that the output speed tracks the command speed perfectly if the estimated speed and the estimated disturbance converge to their actual value.

B. Conventional ESO

Generally, the rotor position is obtained from the position sensors such as the encoder or resolver, and the speed is calculated by the derivative of the position. Due to the quantization error in the measurement of the position, the speed calculated by the classical frequency method is contaminated by a lot of noise, especially in the low-speed range. Using δ_n as the speed measurement noise, thus the measured speed can be expressed as $\Omega^m = \Omega + \delta_n$. In order to attenuate the measurement noise, a small bandwidth low-pass filter (LPF) is usually adopted in the speed feedback.

Suppose the feedback speed is filtered by a first order LPF, whose transfer function can be expressed by

$$G_{\text{LPF}}(s) = \frac{1}{\tau s + 1} = \frac{\omega_f}{s + \omega_f} \quad (9)$$

where τ and ω_f are the time constant and the cut-off frequency of the LPF, respectively.

In the conventional design process of the ESO, the speed filter is not considered. A classical linear second order ESO for the system expressed by (21) can be constructed as

$$\begin{cases} \dot{\tilde{\Omega}} = \Omega^m - \hat{\Omega} = \Omega + \delta_n - \hat{\Omega} \\ \dot{\hat{\Omega}} = bT_e^* + \hat{d} + k_1 \tilde{\Omega} \\ \dot{\hat{d}} = k_2 \tilde{\Omega} \end{cases} \quad (10)$$

where $\tilde{\Omega}$ is the speed observation error, k_1 and k_2 are the observer gains.

The block diagram of the ADRC speed-regulation system is shown in Fig. 1, where Ω_f^m is the filtered speed, and there is

$$\Omega_f^m(s) = \frac{1}{\tau s + 1} \Omega^m(s) = G_{\text{LPF}}(s) \Omega^m(s) \quad (11)$$

From (10) and Fig. 1, the estimated speed and the estimated disturbance can be deduced as

$$\begin{cases} \hat{\Omega}(s) = \frac{\Delta_2 - s^2}{\Delta_2} \Omega_f^m(s) + \frac{s}{\Delta_2} bT_e^*(s) \\ \hat{d}(s) = \frac{k_2 s}{\Delta_2} \Omega_f^m(s) - \frac{k_2}{\Delta_2} bT_e^*(s) \end{cases} \quad (12)$$

where the characteristic polynomial $\Delta_2 = s^2 + k_1 s + k_2$ can be set as $\Delta_2 = (s + \omega_{o2})^2$, ω_{o2} is the undamped natural frequency of the second order ESO, then the gains in the ESO can be calculated by $k_1 = 2\omega_{o2}$, $k_2 = \omega_{o2}^2$.

C. Effect of the Bandwidth of LPF on the ESO and the System

Substituting (5) into (12) yields

$$\begin{cases} \hat{\Omega}(s) = \frac{\Delta_2 + \tau s^3}{\Delta_2(\tau s + 1)} [\Omega(s) + \delta_n(s)] - \frac{s}{\Delta_2} [d(s) + s\delta_n(s)] \\ \hat{d}(s) = -\frac{k_2 \tau s^2}{\Delta_2(\tau s + 1)} [\Omega(s) + \delta_n(s)] + \frac{k_2}{\Delta_2} [d(s) + s\delta_n(s)] \end{cases} \quad (13)$$

It shows that the adoption of the speed filter leads to the wrong estimation of speed. It can be noted that the estimated disturbance is affected by the actual speed. However, the effect is very small because it is proportional to the second derivative of the actual speed. When $\omega_{o2} = 800$ rad/s, the speed estimation performance ($\hat{\Omega}/\Omega$), the disturbance estimation performance ($\hat{d}/d$), and the noised suppression performance ($\hat{\Omega}/\delta_n$) of the conventional ESO are shown in Fig. 2. It can be seen that the noise suppression performance can be improved from -20 dB/dec to -40 dB/dec, while the speed estimation performance is deteriorated.

Similarly, by substituting (13) into (8), the transfer functions from the speed reference to the output speed, from the disturbance to the output speed, and from the measurement noise to the output speed can be deduced as

$$\begin{cases} \Phi_r(s) = \frac{\Omega(s)}{\Omega^*(s)} = \frac{\Delta_2(s + k_{ps})(\tau s + 1)}{\Delta_2^{\text{cl}}} \\ \Phi_d(s) = \frac{\Omega(s)}{d(s)} = \frac{s(s + k_1 + k_{ps})(\tau s + 1)}{\Delta_2^{\text{cl}}} \\ \Phi_n(s) = -\frac{\Omega(s)}{\delta_n(s)} = \frac{k_{ps}(\Delta_2 - s^2) + k_2 s}{\Delta_2^{\text{cl}}} \end{cases} \quad (14)$$

where Δ_2^{cl} is the characteristic polynomial of the closed-loop control system.

$$\Delta_2^{\text{cl}} = \tau s^3(s + k_1 + k_{ps}) + \Delta_2(s + k_{ps}) \quad (15)$$

For the smooth speed reference whose derivative is not so large, e.g., the sinusoidal speed reference or the parabolic speed reference, the derivative feedforward can work normally. Therefore, the transfer function from the speed reference to the output speed is written as $\Phi_{r_sine}(s)$. For the step speed reference, however, the derivative feedforward is blocked due to the saturation of the control quantity, thus the transfer function from the step speed reference to the output speed is deduced as

$$\Phi_{r_step}(s) = \frac{\Omega(s)}{\Omega_{step}^*(s)} = \frac{\Delta_2 k_{ps}(\tau s + 1)}{\Delta_2^{\text{cl}}} \quad (16)$$

When $k_{ps} = 300$ and $\omega_{o2} = 800$ rad/s, the speed tracking performance (Ω/Ω^*), the disturbance rejection property (Ω/d), and the noised suppression performance (Ω/δ_n) of the closed-loop control system are also shown in Fig. 2. It can be seen that the noise suppression performance of the system can be improved from -40 dB/dec to -60 dB/dec by adopting the speed filter, however, it leads to a deteriorated tracking performance for the medium-frequency command. As a result, the rejection property for the medium-frequency disturbance is weakened at the same time.

D. Modified ADRC

From (8) and (13), it is known that the poor dynamic performance is caused by the wrong estimation of the speed. Therefore, a modified ESO is proposed in [18]. Fig. 3 shows the block diagram of the modified ADRC speed-regulation system. In the system, the modified ESO is expressed as

$$\begin{cases} \tilde{\Omega} = \Omega^m - \hat{\Omega} \\ \dot{\tilde{\Omega}}_f = -\tilde{\Omega}_f + \tilde{\Omega} \\ \dot{\hat{\Omega}} = bT_e^* + \hat{d} + k_1 \tilde{\Omega}_f \\ \dot{\hat{d}} = k_2 \tilde{\Omega}_f \end{cases} \quad (17)$$

where $\tilde{\Omega}_f$ is the filtered observation error.

Similarly, the observed states can be deduced as

Fig. 1. Block diagram of the conventional ADRC speed-regulation system

Fig. 2. The effect of τ on the conventional ESO and the conventional ADRC system. (a) Speed estimation. (b) Disturbance estimation. (c) Noise suppression.

$$\begin{cases} \hat{\Omega}(s) = \Omega(s) + \delta_n(s) - \frac{s(\tau s + 1)}{\Delta_3} [d(s) + s\delta_n(s)] \\ \hat{d}(s) = \frac{k_2}{\Delta_3} [d(s) + s\delta_n(s)] \end{cases} \quad (18)$$

where the characteristic polynomial $\Delta_3 = \tau s^3 + s^2 + k_1 s + k_2$.

It shows that the speed can be properly estimated, whereas the disturbance observation is affected by the LPF. When $\omega_{o2} = 800$ rad/s, the speed estimation performance ($\hat{\Omega}/\Omega$), the disturbance estimation performance ($\hat{d}/d$), and the noised

Fig. 3. Block diagram of the modified ADRC speed-regulation system.

Fig. 4. The effect of τ on the modified ESO and the modified ADRC system. (a) Speed estimation. (b) Disturbance estimation. (c) Noise suppression.

suppression performance ($\hat{\Omega}/\delta_n$) of the conventional ESO are shown in Fig. 4. Similarly, the noise suppression performance can be improved from -20 dB/dec to -40 dB/dec.

By substituting (18) into (8), the transfer functions from the sinusoidal speed reference to the output speed, from the step speed reference to the output speed, from the disturbance to the output speed, and from the measurement noise to the output speed can be deduced as

$$\begin{cases} \Phi_{r_sine}(s) = \frac{\Omega(s)}{\Omega_{sine}^*(s)} = 1 \\ \Phi_{r_step}(s) = \frac{\Omega(s)}{\Omega_{step}^*(s)} = \frac{k_{ps}}{s + k_{ps}} \\ \Phi_d(s) = \frac{\Omega(s)}{d(s)} = \frac{k_{ps}s(\tau s + 1) + \Delta_3 - k_2}{(s + k_{ps})\Delta_3} \\ \Phi_n(s) = -\frac{\Omega(s)}{\delta_n(s)} = \frac{k_{ps}(k_1 s + k_2) + k_2 s}{(s + k_{ps})\Delta_3} \end{cases} \quad (19)$$

It can be seen that the tracking performance is not affected by the time constant of the LPF. When $k_{ps}=300$, $\omega_{o2}=800$ rad/s, the disturbance rejection property (Ω/d), and the noise suppression performance (Ω/δ_n) of the closed-loop control system are also shown in Fig. 4. It shows that the noise suppression performance is improved from -40 dB/dec to -60 dB/dec. However, it is found that the medium-frequency disturbance is amplified. The larger the time constant of the LPF is, the amplification is larger, which may lead to the instability of the system.

According to the Routh–Hurwitz stability criterion, the condition for a stable operation is $\tau < k_1/k_2 = 2/\omega_{o2}$, i.e., $\omega > \omega_{o2}/2$, as verified by the root locus shown in Fig. 5. Therefore, the bandwidth of the LPF and the bandwidth of the ESO should be carefully set to ensure a stable operation.

III. PROPOSED ADRC SPEED REGULATION SYSTEM CONSIDERING THE SPEED FILTER

In this paper, a novel ADRC speed regulation system considering the speed filter is proposed. In order to have a general discussion of the filter, the transfer function of the LPF is modified as

$$G_{LPF}(s) = \frac{\omega_{f2}}{s + \omega_{f1}} \quad (20)$$

Fig. 5. Root locus when τ varies ($\omega_{o2}=500$ rad/s).

TABLE I
GENERAL LPF UNDER DIFFERENT CONDITIONS

$\omega_{f1} = \omega_{f2} = \omega_f \neq 0$	$\omega_{f1} = 0, \omega_{f2} = 1$
$G_{LPF}(s) = \frac{\omega_f}{s + \omega_f}$	$G_{LPF}(s) = \frac{1}{s}$

where ω_{f1} and ω_{f2} are the gains. When $\omega_{f1}=\omega_{f2}=\omega_f$, the LPF is the same first order LPF as the one shown by (9). When $\omega_{f1}=0$ and $\omega_{f2}=1$, however, the LPF is equivalent to an integrator, which can be seen as a general low-pass filter, as shown in TABLE I.

Extending the filtered speed Ω_f as a new state, then the state equation of the filtered speed can be expressed as

$$\begin{cases} \dot{\Omega}_f = -\omega_{f1}\Omega_f + \omega_{f2}\Omega \\ \dot{\Omega} = bT_e^* + d \end{cases} \quad (21)$$

A. ESO-based ADRC System Considering the Speed Filter

A classical linear ESO for the system expressed by (21) can be constructed as

$$\begin{cases} \tilde{\Omega}_f = \Omega_f^m - \hat{\Omega}_f = \Omega_f + \delta_{nf} - \hat{\Omega}_f \\ \dot{\hat{\Omega}}_f = -\omega_{f1}\hat{\Omega}_f + \omega_{f2}(\hat{\Omega} + k_0\tilde{\Omega}_f) \\ \dot{\hat{\Omega}} = bT_e^* + \hat{d} + k_1\tilde{\Omega}_f \\ \dot{\hat{d}} = k_2\tilde{\Omega}_f \end{cases} \quad (22)$$

where $\Delta_3 = s^3 + \omega_{f1}s^2 + \omega_{f2}(k_0s^2 + k_1s + k_2)$ is the characteristic polynomial, k_0 , k_1 , and k_2 are the observer gains.

The block diagram of the proposed ADRC system considering the speed filter is shown in Fig. 6. It can be seen that the difference between the proposed ADRC system and the modified ADRC system is the adoption of the gain k_0 . When $k_0=0$, the proposed ADRC system is equivalent to the modified ADRC system since the two LPF can be moved to the observation error path. Hence, the modified ADRC system can be seen as a specific case of the proposed ESO-based ADRC system. It should be noted that there are two kinds of estimated speed when $k_0 \neq 0$, as will analyzed in the PLL-based ADRC system.

From (22) and Fig. 6, the estimated speed and the estimated disturbance can be deduced as

$$\begin{cases} \hat{\Omega}(s) = \Omega(s) + \delta_n(s) - \frac{s(s + \omega_{f1} + k_0\omega_{f2})}{\Delta_3} [d(s) + s\delta_n(s)] \\ \hat{d}(s) = \frac{k_2\omega_{f2}}{\Delta_3} [d(s) + s\delta_n(s)] \end{cases} \quad (23)$$

Substituting (23) into (6), the transfer functions from the sinusoidal speed reference to the output speed, from the step speed reference to the output speed, from the disturbance to the output speed, and from the measurement noise to the output speed can be deduced as

$$\begin{cases} \Phi_{r_sine}(s) = \frac{\Omega(s)}{\Omega_{sine}^*(s)} = 1 \\ \Phi_{r_step}(s) = \frac{\Omega(s)}{\Omega_{step}^*(s)} = \frac{k_{ps}}{s + k_{ps}} \\ \Phi_d(s) = \frac{\Omega(s)}{d(s)} = \frac{(s + k_{ps})s(s + \omega_{f1} + k_0\omega_{f2}) + k_1\omega_{f2}s}{(s + k_{ps})\Delta_3} \\ \Phi_n(s) = -\frac{\Omega(s)}{\delta_n(s)} = \frac{k_{ps}\omega_{f2}(k_1s + k_2) + k_2\omega_{f2}s}{(s + k_{ps})\Delta_3} \end{cases} \quad (24)$$

By setting the characteristic polynomial of the observer as $\Delta_3 = (s + \omega_{o3})^3$, the gain in the ESO can be obtained as

$$\begin{cases} k_0 = (3\omega_{o3} - \omega_{f1})/\omega_{f2} \\ k_1 = 3\omega_{o3}^2/\omega_{f2} \\ k_2 = \omega_{o3}^3/\omega_{f2} \end{cases} \quad (25)$$

where ω_{o3} is the undamped natural frequency of the observer.

Consequently, the system output when the reference

Fig. 6. Block diagram of the ESO-based ADRC system considering the speed filter.

Fig. 7. Bode diagram of the ESO-based ADRC system under different ω_n . (a) Disturbance rejection property. (b) Noise suppression performance.

derivative feedforward works normally can be simplified as

$$\Omega(s) = \Omega^*(s) - \delta_n(s) + \frac{(s + k_{ps})s(s + 3\omega_{o3}) + 3\omega_{o3}^2s}{(s + k_{ps})\Delta_3} [d(s) + s\delta_n(s)] \quad (26)$$

From (26), it can be seen that the command tracking performance, the disturbance rejection ability, and the measurement noise suppression performance are all independent of the bandwidth of LPF, which is verified by the bode diagrams shown in Fig. 7, where $k_{ps}=300$, $\omega_{o3}=800$ rad/s. From (25), it can also be found that the gain k_0 should vary with the bandwidth of the LPF. However, k_0 in the modified ADRC system is always kept as zero.

It should be pointed out that the filtered speed is actually the integration of the speed or the mechanical angle θ_m when $\omega_n=0$ rad/s and $\omega_n=1$ rad/s. Since the mechanical angle can be measured directly by using the position sensor, it is the best choice to set the LPF to be an integrator. In this case, the state equation of the PMSM system expressed by (21) is modified as

Fig. 8. Block diagram of the PLLO-based ADRC system considering the speed filter.

Fig. 9. Bode diagram of the PLLO-based ADRC system under different ω_n . (a) Disturbance rejection property. (b) Noise suppression performance.

$$\begin{cases} \dot{\theta}_m = \Omega \\ \dot{\Omega} = bT_e^* + d \end{cases} \quad (27)$$

and the gains in the ESO expressed by (25) is also simplified as

$$\begin{cases} k_1 = 3\omega_{o3} \\ k_2 = 3\omega_{o3}^2 \\ k_3 = \omega_{o3}^3 \end{cases} \quad (28)$$

B. PLLO-based ADRC System Considering the Speed Filter

It is known from [30] that the ADRC system has different dynamics when different combinations of the speed feedback and disturbance feedforward are employed. Similarly, PLLO considering the speed filter is designed for observing the speed and the disturbance and the ADRC system based on the PLLO is analyzed in this paper.

The block diagram of the PLLO-based ADRC system is shown in Fig. 8. The difference between the PLLO-based ADRC system and the ESO-based ADRC system is the combination of speed feedback and disturbance feedforward. The state equation of the PLLO for the system expressed by (21) can be expressed as

$$\begin{cases} \tilde{\Omega}_f = \Omega_f^m - \hat{\Omega}_f \\ \dot{\hat{\Omega}}_f = -\omega_{f1}\hat{\Omega}_f + \omega_{f2}\hat{\Omega} \\ \dot{\hat{\Omega}} = bT_e^* + \hat{d} + k_0\tilde{\Omega}_f \\ \dot{\hat{d}} = k_1\tilde{\Omega}_f + k_2\tilde{\Omega}_f \end{cases} \quad (29)$$

It can be easily proved that the enhanced ADRC system has the same characteristic polynomial as the proposed ADRC system. Therefore, the estimated speed and the estimated disturbance from the PLLO can be deduced as

$$\begin{cases} \hat{\Omega}(s) = \Omega(s) + \delta_n(s) - \frac{s(s + \omega_{f1})}{\Delta_3} [d(s) + s\delta_n(s)] \\ \hat{d}(s) = \frac{(k_1s + k_2)\omega_{f2}}{\Delta_3} [d(s) + s\delta_n(s)] \end{cases} \quad (30)$$

Substituting (30) into (6), the transfer functions from the

sinusoidal speed reference to the output speed, from the step speed reference to the output speed, from the disturbance to the output speed, and from the measurement noise to the output speed can be deduced as

$$\begin{cases} \Phi_{r_sine}(s) = \frac{\Omega(s)}{\Omega_{sine}^*(s)} = 1 \\ \Phi_{r_step}(s) = \frac{\Omega(s)}{\Omega_{step}^*(s)} = \frac{k_{ps}}{s + k_{ps}} \\ \Phi_d(s) = \frac{\Omega(s)}{d(s)} = \frac{(s + k_{ps})s(s + \omega_{f1}) + \omega_{f2}k_0s^2}{(s + k_{ps})\Delta_3} \\ \Phi_n(s) = -\frac{\Omega(s)}{\delta_n(s)} = \frac{k_{ps}[\Delta_3 - s^2(s + \omega_{f1})] + (k_1s + k_2)\omega_{f2}s}{(s + k_{ps})\Delta_3} \end{cases} \quad (31)$$

When $k_{ps}=300$, $\omega_{o3}=800$ rad/s, the bode diagrams of the enhanced ADRC system under different ω_{f1} is shown in Fig. 9. It can be seen that the system dynamic performance is affected by the bandwidth of the LPF. However, it is worth noting that the effect of the LPF in the PLLO-based ADRC system is different from that in the ESO-based ADRC system. In the PLLO-based ADRC system, lower bandwidth of LPF deteriorates the noise suppression performance but improves the disturbance rejection ability.

Specially, when $\omega_{f1}=0$ rad/s and $\omega_{f2}=1$ rad/s, the system ability for rejecting the low-frequency disturbance can be enhanced from 20 dB/dec to 40 dB/dec, i.e., the steady-state speed tracking error caused by the disturbance is proportional to the second order derivative of the disturbance, as can be seen from (31). As a result, it is also the best choice to set the LPF to an integrator in the PLLO-based ADRC system. In this case, the PLLO is equivalent to the ISO in [28].

C. Comparison of Different ADRC Systems

In this section, the four ADRC systems, i.e., the conventional ADRC system ($\omega_f=2000$ rad/s), the modified ADRC system ($\omega_f=2000$ rad/s), the proposed ESO-based ADRC system ($\omega_{f1}=0$ rad/s and $\omega_{f2}=1$ rad/s), and the proposed PLLO-based ADRC system ($\omega_{f1}=0$ rad/s and $\omega_{f2}=1$ rad/s), are compared. Comparing Fig. 9 with Fig. 7, it can be found that the proposed ESO-based

Fig. 10. Comparison of the four ADRC systems. (a) Tracking performance. (b) Disturbance rejection property. (c) Noise suppression performance.

ADRC system has a better measurement noise suppression performance than the proposed PLLO-based ADRC system, while the latter performs better low-frequency disturbance rejection ability than the former. Therefore, the ESO is usually set with a large bandwidth while the PLLO is set with a small bandwidth in practice.

When $k_{ps}=300$, the bandwidths of the observer in the four systems are set as 800 rad/s, 800 rad/s, 1000 rad/s, and 300 rad/s respectively, the bode diagram of the four ADRC systems are shown in Fig. 10. It can be seen that the conventional ADRC system, the modified ADRC system, and the proposed ESO-based ADRC system have nearly the same disturbance rejection property and the measurement noise suppression performance. Nevertheless, the conventional ADRC system has a poor tracking performance. The proposed PLLO-based ADRC system has the best rejection ability for the low-frequency disturbance but the worst suppression performance for the medium frequency disturbance and the high-frequency measurement noise.

Since the proposed ESO-based ADRC system and the PLLO-based ADRC system have their own advantages, the practitioners of ADRC can make a selection between these two ADRC systems according to the requirements of their applications.

TABLE II
SPECIFICATION OF THE PMSM

Symbol	Quantity	Symbol	Quantity
Rated power P_N	0.75 (kW)	Number of pole-pairs p_n	4
Rated voltage U_N	220 (V)	D axis inductance L_d	8.0 (mH)
Rated speed n_N	3000 (rpm)	Q axis inductance L_q	8.0 (mH)
Rated torque T_N	2.4 (N·m)	Torque constant K_t	0.553 (N·m/A)
Current Limit I_{smax}	10 (A)	Motor inertia J_M	1.62×10^{-4} (kg·m ²)
Stator resistance R_s	1.1 (Ω)	Motor system inertia J	4.44×10^{-4} (kg·m ²)

Fig. 11. The block diagram of the control system.

Fig. 12. Configuration of the test bench

IV. EXPERIMENTAL RESULTS

In order to verify the above analysis and the proposed methods, experimental results of the four ADRC systems are presented and analyzed in this section. The specification of the servo motor is shown in TABLE II. As shown in Fig. 11, the control strategy is based on space vector pulse width modulation (SVPWM) control with $i_d^*=0$. Decoupled PI regulators are employed in the current-loop to control i_d and i_q respectively, and different ADRC controllers are employed in the speed-loop. The DC bus voltage u_{dc} is 150V. The saturation limit of q -axis reference current is 9 A. The bandwidths of d -axis and q -axis current-loops are 2000 rad/s.

The configuration of the test bench is shown in Fig. 12. A programmable DC power supply is used to provide 150V DC Bus voltage. There are two same PMSMs, one acts as a driving motor and the other acts as a generator. The driving inverter consists of dSPACE DS1103, Mitsubishi intelligent power module (IPM), current and voltage hall sensors, filtering and protection circuit, etc. The field-oriented control is employed for driving the PMSM.

A. Effect of the LPF

Given a step speed reference from -1000 rpm to 1000 rpm, when $k_{ps}=300$, the bandwidth of the observer in conventional ADRC system and the modified ADRC are $\omega_{o2}=800$ rad/s, while the one in the ESO-based ADRC system and the PLLO-based ADRC system are 1000 rad/s and 300 rad/s, respectively. The bandwidth of the LPF for speed feedback is set as 2000 rad/s and 500 rad/s, respectively. The speed responses of the four ADRC systems when tracking step speed reference and rejecting the load disturbance are shown in Fig. 13 and Fig. 14, respectively. In order to have a clear understanding of the relationship between the observed speed and the actual speed, both the observed speed (in Blue) and the filtered speed (in Magenta) are given. Meanwhile, step loading and slope unloading are used to test the disturbance rejection ability under different kinds of load torque. The bandwidth of the LPF is set to 2000 rad/s so that the filtered speed can be treated as the actual speed in low dynamic situations.

It can be seen that with the decrease of the bandwidth of LPF, the conventional ADRC system is prone to oscillation, which verifies that the tracking performance of the conventional ADRC system is affected by the bandwidth of LPF; the modified ADRC system tends to be unstable; the estimated speed, the disturbance rejection property, and the measurement noise suppression of the proposed ESO-based ADRC system are all not affected by the bandwidth of the LPF. For the PLLO-based ADRC system, the tracking performance is not affected by the bandwidth of the LPF. However, lower bandwidth is good for the disturbance rejection while bad for the noise suppression. Meanwhile, the higher the bandwidth is, the larger deviation between the observed speed and the actual speed is. All these results are identical to the theoretical results.

B. Comparison of Different ADRC Systems

As analyzed before, the conventional ADRC system and the

Fig. 13. Step response of the four ADRC systems under different bandwidth of LPF.

Fig. 14. Disturbance rejection of the four ADRC systems under different bandwidth of LPF.

modified ADRC system also show a good dynamic performance when a high bandwidth ($\omega_f=2000$ rad/s) LPF is used. When $\omega_1=0$ rad/s and $\omega_2=1$ rad/s, the proposed ESO-based ADRC system and the PLLO-based ADRC system have their special merits. Therefore, these four systems are compared in the experiments. The speed responses when tracking the step speed reference, rejecting the load disturbance, and rejecting the torque ripples in the low-speed range (30 rpm and 100 rpm) are shown in Fig. 15.

It can be seen from Fig. 15 that all the four systems have a good tracking performance for step speed reference. The convention ADRC system, the modified ADRC system, and the proposed ESO-based ADRC system have nearly the same disturbance rejection property for all kinds of disturbances. When a slope load disturbance is activated, all these three ADRC systems have a constant steady-state tracking error, however, the PLLO-based ADRC system has no steady-state tracking error. Hence, the PLLO-based ADRC system shows a better rejection ability for the low-frequency disturbance than the other three systems. When a step load torque disturbance is activated, however, the PLLO-based ADRC system shows a worse disturbance rejection ability than the other three systems.

The PLLO-based ADRC system has a poor rejection ability for the medium-frequency disturbance, which can also be found

in the constant speed operation experiments. When the speed is only 30 rpm, the frequency of the main component of the torque ripple is only 18π rad/s (9 Hz), the four systems have a comparable smooth speed. However, when the speed is increased to 100 rpm, the frequency of the main component of the torque ripple is only 60π rad/s (30 Hz), the PLLO-based ADRC system has a larger speed pulsation. Moreover, in the low-speed range, the speed noise in the PLLO-based ADRC system is much larger than that in the other systems. Though smaller bandwidth PLLO can be used to achieve a better noise suppression, a larger speed pulsation can be expected.

Fig. 16 shows the comparison of the sinusoidal command tracking performance. Two kinds of sinusoidal speed references are used for the test. When $n_{ref}=100\sin 40\pi t$ rpm, the four ADRC systems have a comparable tracking performance. The tracking error is caused by the Coulomb friction torque. The effect of the Coulomb friction torque can be alleviated when a high frequency sinusoidal speed reference is used. When $n_{ref}=50\sin 200\pi t$ rpm, a speed response with larger amplitude can be found in the conventional ADRC system, which is identical to the theoretical results shown in Fig. 10(a). In Fig. 10, delays caused by the current control loop and the digital system is not considered. Actually, these delays will cause the phase delay in the speed response. This point will be studied in our future work. Anyway, it can be seen that the modified ADRC system, the

Fig. 15. Comparison of the four systems when tracking step speed reference and rejecting different kinds of disturbances.

Fig. 16. Comparison of the four systems when tracking sinusoidal speed reference.

proposed ESO-based ADRC system, and the proposed PLLO-based ADRC system show better tracking performance for the high-frequency sinusoidal speed reference than the conventional ADRC system, which is consistent with the theoretical results.

V. CONCLUSION

In this paper, the conventional speed ADRC system and the modified ADRC system considering the speed filter are analyzed. The conventional ADRC system is prone to

oscillation when the bandwidth of the LPF is decreased since the speed filter is not considered in the design of ESO. In the modified ADRC system, the LPF is used for filtering the observation error of ESO. Though the tracking performance is improved, the modified ADRC system tends to be unstable when a low bandwidth LPF is employed. To solve these problems, two ADRC systems considering the speed filter is proposed.

The proposed ESO-based ADRC system is independent of the bandwidth of the LPF, thus it is the best choice to use the integrator as the LPF. In this case, the position can be directly taken as an input of the observer without calculating the speed first. The proposed ESO-based ADRC system has comparable disturbance rejection ability and measurement noise suppression performance with the conventional ADRC system and the modified ADRC system.

The proposed PLLO-based ADRC system is affected by the bandwidth of the LPF. When an integrator is used as the LPF, it has a better rejection ability for the low-frequency disturbance than the other three ADRC systems at the expense of deteriorated noise suppression performance. Similarly, no speed calculation is needed for observing disturbance when compared with the PLLO-based ADRC system proposed in [30].

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