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An Adaptive Quasi-Sliding-Mode Rotor Position Observer-Based Sensorless Control for Interior Permanent Magnet Synchronous Machines

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Abstract—Advantages such as parameter insensitivity and high robustness to system structure uncertainty make the sliding-mode observer (SMO) a promising solution for sensorless control of interior permanent magnet synchronous machines (IPMSMs). In practical industry applications, in order to utilize digital controllers and achieve comparable performance under a lower sampling frequency, a discrete-time or quasi-SMO (QSMO) is commonly used. However, because of the saliency of an IPMSM, the magnitude of the extended electromotive force (EMF) will change with load (torque and/or speed) variations, which makes it challenging for the QSMO to estimate the extended EMF accurately. Without proper observer parameters, a phase shift will be observed in the QSMO-estimated rotor position when the load changes. In order to overcome this problem, an adaptive QSMO using an online parameter adaption scheme is proposed to estimate the extended EMF components in an IPMSM, which are then used to estimate the rotor position of the IPMSM. The resulting position estimation has zero phase lags and is highly robust to load variations. The proposed adaptive QSMO is implemented on a 150-kW IPMSM drive system used in heavy-duty, off-road, hybrid electric vehicles. Testing results for ramp torque changes, four-quadrant operations, and complete torque reversals between full motoring and full braking modes are presented to verify the effectiveness of the proposed sensorless control algorithm.

Index Terms—Adaptive observer, interior permanent magnet synchronous machine (IPMSM), position estimation, quasi-sliding-mode observer (QSMO), sensorless control.

I. INTRODUCTION

INTERIOR permanent magnet synchronous machines (IPMSMs) are widely used in electric and hybrid electric vehicle systems due to their distinctive advantages, such as high efficiency, high power density and wide constant power region. In traditional IPMSM drives, electromechanical position sensors, e.g., resolvers, optical encoders, and hall-effect sensors, are commonly used to obtain the accurate information of

rotor position in order to achieve high-performance control for the IPMSMs. The use of these sensors increases the cost, size, weight, and hardware wiring complexity of the IPMSM drive systems. From the viewpoint of system reliability, mounting electromechanical sensors on rotor shafts will degrade mechanical robustness of the electric machines. The electromagnetic interference (EMI) noise in wiring harness due to switching events and broken wires may be fatal to the controller operation. Moreover, sensors are often subject to high failure rates in harsh environments, such as excessive ambient temperature, superhigh-speed operation, and other adverse or heavy-load conditions [1]. To overcome these drawbacks, much research effort has gone into the development of sensorless drives that have comparable dynamic performance to sensor-based drives during the last decades.

Among many different rotor position estimation schemes proposed for sensorless control of permanent magnet synchronous machines (PMSMs) [2]–[4], [14]–[22], the electromotive force (EMF)-based position estimation methods [2], [3] are one of the major techniques for medium and high-speed applications. However, due to the saliency of IPMSMs, the magnitudes of the back EMF or extended EMF [2] components depend on both operating conditions (e.g., rotor speed and load) and machine parameters (i.e., inductances and stator resistance). In some of the existing observer design methods, it is assumed that the speed-related back EMF term is relatively dominant at a high speed and the change in the magnitude of the back EMF is relatively small when the load changes. However, for heavy-load applications, e.g., traction motors in electric vehicles, the effect due to load changes cannot be ignored. Therefore, an advanced rotor position observer that is robust to large load variations is needed.

Among different types of rotor position observers, the sliding-mode observer (SMO) is a promising candidate. Generally speaking, an SMO is an observer whose inputs are discontinuous functions of the errors between the estimated and measured outputs [7]. If a sliding-mode manifold is well designed, when the trajectories of the desired states reaches the designed manifold, the sliding mode will be enforced. The dynamics of the system states of interest under the sliding mode depend only on the surfaces chosen in the state space and are not affected by system structure or parameter uncertainties. These features are especially attractive for IPMSM applications since the parameters of an IPMSM often vary with operating conditions.

The use of sliding-mode principles for digital control systems has become more and more popular over the last few years

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owing to the widespread use of digital controllers [7]. Fast control-loop frequencies that typically occur in a continuous-time SMO require a very small sampling period to make the observer work properly. In a discrete-time sliding-mode controller or observer, to facilitate digital signal processor or microchip-based applications, a finite sampling frequency is used; and the controller or observer inputs are calculated once per sampling period and held constant during that interval. Under such a circumstance, the trajectories of the system states of interest are unable to precisely move along the sliding surface, which will lead to a quasi-sliding-mode motion only [8], [9]. Recently, discrete-time SMOs (DSMO) have received more and more attention since discretized reaching laws were proposed [8]–[13], which can be used for nonlinear dynamical models with various model/parameter uncertainties or disturbances.

The SMO has been extensively applied to sensorless control of surface-mounted PMSMs [14]–[17]. In PMSM applications, the inductances L_d and L_q in the d – q rotating reference frame are assumed to be equal; and the magnitude of the back EMF is only a function of rotor speed and is not affected by load variations. Although the extended EMF-based IPMSM model [2] has a similar structure to the surface-mounted PMSM model, the magnitude of the extended EMF is a function of both rotor speed and stator currents, which depend on load conditions. Because of this, the rotor position estimated by an SMO will have a variable phase shift during load changes in the IPMSM applications [18].

This paper proposes an extended EMF-based adaptive quasi-sliding-mode position observer (QSMO) for sensorless IPMSM drives operating under medium- and high-speed conditions. The switching function of the adaptive QSMO is well designed to guarantee global stability of the position observer and bounded motion of the state trajectory within a limited boundary layer. Since the extended EMF depends on both speed and torque, an adaptive parameter selection methodology is proposed for the QSMO based on a discretized reaching law, which is easy to implement in practical IPMSM drives. The proposed adaptive QSMO is validated by experimental results on a 150-kW IPMSM drive system used for heavy-duty, off-road, hybrid electric vehicles. The results on ramp torque tests with different slew rates, four-quadrant operations, and complete torque reversals between full motoring and full braking operations are presented to verify the effectiveness of the proposed sensorless control system.

II. QSMO DESIGN FOR IPMSMS

A. Extended EMF-Based IPMSM Model

The dynamics of an IPMSM can be modeled in the α – β stationary reference frame as

$$\begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix} = p \begin{bmatrix} L + \Delta L \cos(2\theta_{re}) & \Delta L \sin \theta_{re} \\ \Delta L \sin \theta_{re} & L - \Delta L \cos(2\theta_{re}) \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + R \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \omega_{re} \psi_m \begin{bmatrix} -\sin \theta_{re} \\ \cos \theta_{re} \end{bmatrix} \quad (1)$$

where θ_{re} is the rotor position angle; p is the derivative operator; v_α and v_β are the stator voltages; i_α and i_β are the stator currents; ω_{re} is the rotor electrical speed; R is the stator resistance; $L = (L_d + L_q)/2$; $\Delta L = (L_d - L_q)/2$; and L_d and L_q are the d -axis and q -axis inductances, respectively. Due to the saliency of the IPMSM (i.e., $L_d \neq L_q$), both the back EMF components and the inductance matrix contain the information of the rotor position angle. Moreover, since (1) contains both $2\theta_{re}$ and θ_{re} terms, it is not easy to obtain the rotor position from the back EMF components directly. To facilitate the rotor position observation, an extended EMF-based IPMSM model is proposed in [2] as follows:

$$\begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix} = \begin{bmatrix} R + pL_d & \omega_{re}(L_d - L_q) \\ \omega_{re}(L_q - L_d) & R + pL_d \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \underbrace{[(L_d - L_q)(\omega_{re}i_d - pi_q) + \omega_{re}\psi_m]}_{\text{extended EMF}} \begin{bmatrix} -\sin \theta_{re} \\ \cos \theta_{re} \end{bmatrix}. \quad (2)$$

In (2), only the extended EMF components contain the information of the rotor position. If the extended EMF components can be estimated, the rotor position can be obtained directly.

However, the expression of the extended EMF of an IPMSM is much more complex than the back EMF expression of a surface-mounted PMSM. For a surface-mounted PMSM, the back EMF is only a function of rotor speed. If the PMSM accelerates, the magnitude of the back EMF will increase and vice versa. While for an IPMSM, the magnitude of the extended EMF is a function of rotor speed, stator current, and derivative of stator current, which means that the load condition will affect the magnitude of the extended EMF, and this effect is nonlinear. For high-speed and heavy-load applications, the currents will have large changes in a short period during a state transient and the current derivatives will become very large, which will cause significant distortions of the waveforms of the extended EMF components. To overcome this problem, a rotor position observer that is highly robust to load variations is needed for high-performance sensorless control of IPMSMs.

B. State-Based QSMO Design

Let η denote the $(L_d - L_q)(\omega_{re}i_d - pi_q) + \omega_{re}\psi_m$ term, which is the magnitude of the extended EMF components, the dynamic current equations of an IPMSM can be expressed as

$$\begin{cases} \frac{di_\alpha}{dt} = \frac{v_\alpha}{L_d} - \frac{R}{L_d}i_\alpha + \omega_{re}\frac{L_q - L_d}{L_d}i_\beta + \frac{\eta}{L_d}\sin \theta_{re} \\ \frac{di_\beta}{dt} = \frac{v_\beta}{L_d} - \frac{R}{L_d}i_\beta - \omega_{re}\frac{L_q - L_d}{L_d}i_\alpha - \frac{\eta}{L_d}\cos \theta_{re}. \end{cases} \quad (3)$$

In order to transform the continuous-system model (3) into a discrete-time model, the following first-order Euler method is used to represent the derivative terms:

$$\frac{d\mathbf{X}(t)}{dt} \approx \frac{\mathbf{X}[k+1] - \mathbf{X}[k]}{T_s} \quad (4)$$

where T_s is the sampling interval. Then, the discrete version of (3) at the $(k+1)$ th time step can be expressed as

$$\begin{cases} \frac{i_\alpha[k+1]}{T_s} = \frac{v_\alpha}{L_d} - \omega_{re} \frac{2\Delta L}{L_d} i_\beta[k] + E_\alpha + \left(\frac{L_d - T_s R}{T_s L_d} \right) i_\alpha[k] \\ \frac{i_\beta[k+1]}{T_s} = \frac{v_\beta}{L_d} + \omega_{re} \frac{2\Delta L}{L_d} i_\alpha[k] + E_\beta + \left(\frac{L_d - T_s R}{T_s L_d} \right) i_\beta[k] \end{cases} \quad (5)$$

where $E_\alpha = \eta \sin \theta_{re} / L_d$ and $E_\beta = -\eta \cos \theta_{re} / L_d$. A current estimator which has the same structure as the current model (5) of the IPMSM can be designed as follows:

$$\begin{cases} \frac{\hat{i}_\alpha[k+1]}{T_s} = \frac{v_\alpha^*}{L_d} - \omega_{re} \frac{2\Delta L}{L_d} \hat{i}_\beta[k] + l Z_\alpha + \left(\frac{L_d - T_s R}{T_s L_d} \right) \hat{i}_\alpha[k] \\ \frac{\hat{i}_\beta[k+1]}{T_s} = \frac{v_\beta^*}{L_d} + \omega_{re} \frac{2\Delta L}{L_d} \hat{i}_\alpha[k] + l Z_\beta + \left(\frac{L_d - T_s R}{T_s L_d} \right) \hat{i}_\beta[k] \end{cases} \quad (6)$$

where Z_α and Z_β are the outputs of a switching function, which is a saturation function in this paper; and l is the observer gain. In (6), the command voltage values v_α^* and v_β^* are used, which are obtained from the current-regulated vector control of the IPMSM, such that the terminal voltages do not need to be measured. If the insulated-gate bipolar transistor (IGBT) dead-time effect is well compensated, the command voltages will be equal to the terminal voltages and the effect of the IGBT dead time on the current estimation can be neglected [23].

Let $\varepsilon[k]^T = [i_\alpha[k] - \hat{i}_\alpha[k] \ i_\beta[k] - \hat{i}_\beta[k]]$ be the vector of the current tracking errors, and the equations of the current tracking error dynamics can be obtained by subtracting (6) from (5)

$$\begin{cases} \varepsilon_\alpha[k+1] = \left(1 - \frac{T_s R}{L_d} \right) \varepsilon_\alpha[k] + T_s E_\alpha[k] - T_s l Z_\alpha[k] \\ \varepsilon_\beta[k+1] = \left(1 - \frac{T_s R}{L_d} \right) \varepsilon_\beta[k] + T_s E_\beta[k] - T_s l Z_\beta[k]. \end{cases} \quad (7)$$

The sliding surface is designed as $s[k] = \varepsilon[k] = 0$. A variable switching function is defined as follows for the QSMO:

$$Z_{\alpha\beta} = \begin{cases} Z_0 & \varepsilon[k] \geq Z_0 \\ \varepsilon[k] & -Z_0 < \varepsilon[k] < Z_0 \\ -Z_0 & \varepsilon[k] \leq -Z_0 \end{cases} \quad (8)$$

where Z_0 is the width of the boundary layer, and $Z_0 > 0$. The switching function will change its output according to the movement of the state (i.e., the current tracking error) trajectory and force the state trajectory to move toward the sliding surface and remain in a quasi-sliding mode. If the quasi-sliding mode is enforced, the current tracking error will be limited within a certain boundary; the output of the switching function will be equal to the extended EMF with harmonics, where the harmonics can be removed by using the filters proposed in [24].

III. PARAMETER ADAPTION SCHEME

The two parameters, i.e., the observer gain l and the width Z_0 of the boundary layer of the saturation function, are critical to the performance of the QSMO. In this section, an online parameter adaption methodology is proposed for the QSMO. The proposed method is originated from system stability verification.

Fig. 1. Illustration of the state trajectory for Condition 1.

Fig. 2. Illustration of the state trajectory for Condition 2.

A. Stability Analysis

A stability analysis is provided to verify that, if the parameters are selected properly, the QSMO will exhibit a quasi-sliding-mode behavior after a finite time step. In order to force the state trajectory to move from the initial state to the sliding surface, the following two conditions should be satisfied simultaneously, and the corresponding schematic diagrams are shown in Figs. 1 and 2, respectively.

- 1) The state trajectory should move toward the sliding surface when the state magnitude is larger than the width of the boundary layer (i.e., $|\varepsilon[k]| > Z_0$), which means a) when $\varepsilon[k] > Z_0$, $\varepsilon[k+1] < \varepsilon[k]$; b) while when $\varepsilon[k] < -Z_0$, $\varepsilon[k+1] > \varepsilon[k]$.
- 2) The state trajectory should not move too far in the approaching direction in each step. In order to limit the change of the state trajectory between the k th and $(k+1)$ th steps, the following condition should be satisfied: a) when $\varepsilon[k] > Z_0$, $\varepsilon[k+1] + \varepsilon[k] > 0$; b) while when $\varepsilon[k] < -Z_0$, $\varepsilon[k+1] + \varepsilon[k] < 0$.

If both conditions are satisfied, not only the discretized convergence but also the stability of the observer can be guaranteed, where the discretized stability criterion can be expressed as $\frac{1}{2}(\varepsilon[k+1] - \varepsilon[k]) \cdot \varepsilon[k] < 0$. In order to satisfy these two conditions, the following constraints for the parameters of the QSMO can be obtained:

$$|E_\alpha[k]| < l Z_0 < \left(\frac{2}{T_s} - \frac{R}{L_d} \right) Z_0 - |E_\alpha[k]| \quad (9-I)$$

$$Z_0 > \frac{2|\eta|}{2L_d f_s - R}. \quad (9-II)$$

The derivation of (9-I) and (9-II) is provided in the Appendix. Since the magnitudes and frequencies of E_α and E_β are identical except that they have 90° phase shift, (9-I) is also applicable to E_β . In (9-I), the inequality on the left-hand side indicates that lZ_0 should be larger than the magnitude of the extended EMF. If this inequality is satisfied, Condition 1 can be guaranteed. This requirement has been mentioned in the previous work [14]–[17]. However, the inequality on the right-hand side of (9-I) should be satisfied simultaneously, which is derived from Condition 2 and indicates that lZ_0 should also have an upper boundary. Otherwise, a phase shift will present in the rotor position estimated from the QSMO when the load changes. Furthermore, without proper parameters, a discretized chattering problem or even system instability will occur.

In order to guarantee the existence of l and Z_0 , the upper boundary in (9-I) should be always greater than the lower boundary, which is used to derive (9-II). It is known that the SMO has a high-gain effect, i.e., a large observer gain can help suppress the tracking error caused by disturbances. Therefore, in this paper, theoretically the tracking error ε can be reduced by increasing the observer gain l . However, as shown in (9-II), for a discrete-time system, the tracking error cannot be reduced by arbitrarily increasing the observer gain l , because the minimum tracking error depends on the sampling frequency f_s .

B. Parameter Adaption Scheme

Let $Z_{\min}$ denote the minimum value of Z_0 . According to (9-II), $Z_{\min}$ is defined as

$$Z_{\min} = \frac{2|\eta|}{2L_d f_s - R}. \quad (10)$$

If a constant pulse width modulation (PWM) frequency is adopted and currents are sampled once per PWM cycle, the sampling frequency f_s can be viewed as a constant. Assume that the machine parameters have no large variations. Therefore, $Z_{\min}$ is a function of η . Under low-speed or light-load operating conditions, η will be small, and therefore, $Z_{\min}$ will be relatively small. On the other hand, under high-speed or heavy-load conditions, η will be large and $Z_{\min}$ will also be relatively large. In order to satisfy both (9-I) and (9-II), Z_0 should be larger than the maximum value of $Z_{\min}$ corresponding to the highest speed and maximum torque condition. However, for low-speed and light-load conditions, a small Z_0 is desired to ensure good current tracking performance. The best method to solve this dilemma is using an adaptive Z_0 to not only satisfy (9) but also guarantee the best current tracking performance for each load condition.

Consider again the magnitude of the extended EMF $\eta = (L_d - L_q)(\omega_{re} i_d - di_q/dt) + \omega_{re} \psi_m$. In steady state, di_q/dt can be assumed to be 0. Thus, if the values of i_d and ω_{re} are known, the value of η can be determined. In practice, the value of i_d can be obtained from the electromagnetic torque command. For an IPMSM, the electromagnetic torque T_e can be expressed as

$$T_e = \frac{3}{2} p_o i_q [(L_d - L_q) i_d + \psi_m] \quad (11)$$

where p_o is the number of magnetic pole pairs of the IPMSM. The relationship between i_d and i_q depends on the control algorithm used for the IPMSM. For example, if the maximum torque per ampere (MTPA) control is used, a simplified relationship between i_d and i_q can be obtained by taking Taylor's series expansion [18] as follows:

$$i_d^* = \frac{(L_d - L_q)}{\psi_m} i_q^2. \quad (12)$$

According to (11) and (12), once the torque command is given, the values of i_d and i_q can be uniquely determined. In practice, the relationship between the command torque T_e^* and currents i_d and i_q can be implemented by using lookup tables or high-order polynomials.

According to previous analysis, η can be expressed as a function of the electromagnetic torque T_e and speed ω_{re} of the IPMSM, i.e., $\eta = \eta(T_e, \omega_{re})$. If both the speed and torque commands (ω_{re}^* and T_e^*) are given, the value of η can be uniquely determined. Then, $Z_{\min}$ can be calculated by using (10) and Z_0 can be simply set to be equal to $Z_{\min}$. However, the method of directly setting $Z_0 = Z_{\min}$ has some limitations. First, since the machine parameters L_d and R may change significantly with operating conditions, it will require extra effort to obtain the accurate information of these parameters for determining $Z_{\min}$ using (10). Second, (9-I), (9-II), and (10) are derived for steady-state operating conditions. During transient conditions, the exact value of the current derivative term di_q/dt is difficult to obtain. Considering these two uncertainties, this study proposes the following methods to ensure that the QSMO is robust to both load transients and machine parameter variations.

First, in industrial drives, the maximum slew rate limit of the current change is usually set in the controller. Thus, the current derivative is a bounded value. To handle current transients during load variations, the values of l and Z_0 are adaptively determined from $Z_{\min}$ online as follows:

$$Z_0 = \alpha Z_{\min} \quad (13)$$

where α is a new coefficient, which is always greater than 1. The method to determine α based on the slew rate limit of the current change will be discussed later. Furthermore, according to (9-I), lZ_0 should be greater than the magnitude of the extended EMF, which can be guaranteed if $l \cdot Z_{\min}$ is set to be equal to the magnitude of the estimated extended EMF. Therefore

$$l = \hat{\eta} / Z_{\min}. \quad (14)$$

To guarantee Z_0 determined by (13) will always satisfy (9-I) and (9-II) in the transient, a sufficiently large α should be selected, i.e., $\alpha - 1 \geq \left| \frac{(L_d - L_q) p i_q}{(L_d - L_q)(\omega_{re} i_d) + \omega_{re} \psi_m} \right|$, where the sum of the numerator and denominator is the magnitude of the extended EMF η in (2), and the denominator is the value of η at steady state. Therefore, during a large load transition, the value of $(\alpha - 1)$ indicates the maximum percentage of the uncertainty in η with respect to its steady-state value caused by the current

Fig. 3. Block diagram of the adaptive QSMO.

transient term pi_q , which can be further written as follows:

$$|pi_q| \leq (\alpha - 1) \underbrace{\left| \omega_{re} \left(i_d + \frac{\psi_m}{L_d - L_q} \right) \right|}_{\beta}. \quad (15)$$

In (15), the maximum value of the current derivative can be determined from the slew rate limit of the current. Then, the relationship between the magnitudes of the actual and estimated extended EMF will be

$$\begin{aligned} |\eta| &= |(L_d - L_q)(\omega_{re} i_d) + \omega_{re} \psi_m - (L_d - L_q)pi_q| \\ &\leq |(L_d - L_q)(\omega_{re} i_d) + \omega_{re} \psi_m| + |(L_d - L_q)pi_q| \\ &\leq \alpha |(L_d - L_q)(\omega_{re} i_d) + \omega_{re} \psi_m| \\ &= \alpha |\hat{\eta}|. \end{aligned} \quad (16)$$

Therefore, with a sufficiently large α , the observer parameters calculated by (13) and (14) will always satisfy (9-I) and (9-II). In (15), the value of β can be calculated by using the steady-state values of i_d and ω_{re} . In normal cases, there is $L_d < L_q$ and i_d is always negative for flux weakening or i_d is equal to zero otherwise. Thus, i_d and $\psi_m / (L_d - L_q)$ have the same sign. To ensure that (15) is always valid for all the current conditions, a large value is obtained for α by using the minimum value of β when $i_d = 0$. Therefore, α can be determined as follows:

$$\alpha = 1 + \frac{(L_d - L_q) |pi_q|_{\max}}{\omega_{re} \psi_m} \quad (17)$$

where $|pi_q|_{\max}$ is the maximum slew rate limit of the current derivative, and $(\alpha - 1)$ is inversely proportional to the rotor electrical speed. The block diagram of the proposed parameter adaption scheme and the resulting adaptive QSMO are shown in Fig. 3.

Second, machine parameter variations are always one of the most critical issues in the IPMSM position estimation. In high-power applications, the machine parameters, e.g., stator resistance R and inductances L_d and L_q , will have large variations when the operating point changes. In the denominator of (10), R is much smaller than the term $2L_d f_s$. Therefore, the variation of R has little influence to the observer performance, especially under medium- and high-speed conditions. To consider the effect of L_d and L_q variations on the QSMO performance, lookup tables are utilized to obtain their values in real time according

Fig. 4. L_q lookup table generated by a FEA method.

to the load conditions. For example, a finite-element analysis (FEA)-based method can be used to find the relationships between the inductances and the stator currents and gamma angle, which is defined as the angle between the phase current vector and i_d vector. Such relationships can be expressed by lookup tables, as shown in Fig. 4 for L_q of the IPMSM used in this paper. The lookup tables can then be used to calculate the QSMO parameters based on (10). By using the coefficient α and the inductance lookup tables, the proposed adaptive QSMO is robust to both machine parameter variations and load transients.

C. Overall Sensorless Control System

The proposed adaptive QSMO is integrated in the current-regulated space vector control of the IPMSM, leading to a sensorless control system for the IPMSM, as shown in Fig. 5. The rotor position is obtained from the proposed QSMO; the rotor speed is then calculated by using the estimated rotor position. A proportional-integral (PI) speed regulator is used to generate the torque command from the speed tracking error. If the IPMSM is operated in the torque control mode, the torque percentage can be directly commanded instead of being generated from the outer-loop speed control. The base torque is the maximum torque at each speed point and is obtained by using a 2-D lookup table. Since the inverter dc-link voltage also affects the current command, a speed-voltage ratio is used. The current commands are generated by two lookup tables based on torque percentage and speed-voltage ratio. Other modules of the control system include current PI regulators with feedforward voltage compensation, Park transformation, space-vector PWM generator, etc.

IV. EXPERIMENT RESULTS

A. Test Stand Setup

An experimental stand is designed to further validate the proposed adaptive QSMO, as shown in Fig. 6. In the test stand, a prime mover machine and an IPMSM are connected back to back sharing a common dc bus from a power supply. The dc-bus voltage is 700 V. The prime mover machine maintains the shaft speed while the IPMSM works in the torque control mode.

Fig. 5. Block diagram of the proposed adaptive QSMO-based sensorless control scheme for an IPMSM.

Fig. 6. Schematic of the test stand for the IPMSM drive.

 TABLE I
SPECIFICATION OF THE IPMSM

Maximum power	150 kW	Stator resistance	0.01 Ω
Maximum torque	300 Nm	Base speed	5000 RPM
Flux linkage	0.095 Wb	Pole pairs number	4
Average d -axis inductance	0.2 mH	Average q -axis inductance	0.55 mH

The parameters of the IPMSM are listed in Table I. Considering current regulation quality, switching losses, system efficiency, switching noise, and EMI issues, the PWM frequency is selected as 6 kHz. The sampling frequency for the currents is the same as the PWM switching frequency. The QSMO is executed once per PWM cycle. Since the command voltages instead of the measured IPMSM terminal voltages are used in the QSMO, the IGBT dead-time effect will cause a phase shift between the estimated and measured rotor positions. In the test stand, the IGBT dead-time effect of the inverter has been fully compensated. Therefore, using the command voltages is the same as using the measured IPMSM terminal voltages [23].

According to the parameters listed in Table I, a suitable value is determined for α . Suppose that the highest torque slew rate for the IPMSM drive system is 20 000 N·m/s at the base speed. When the command torque increases with the maximum slew rate from 0 N·m to the full load of 300 N·m within 15 ms, and i_q^* correspondingly increases from 0 to 350 A, then $pi_q = 23$ kA/s. If $i_d = 0$, then $\beta = |\omega_{re}\psi_m / (L_d - L_q)| = 142$ kA/s.

Therefore, α is calculated to be 1.16 and is chosen to be 1.2 in the experiments.

To fully evaluate the performance of the sensorless drive system using the proposed adaptive QSMO, four groups of testing results are presented: 1) System steady-state performance: verify the zero-phase-lag (between the estimated and measured positions) behavior for different load levels at the base speed, where zero phase lag means that the average position estimation error is zero [3]; 2) system dynamic performance under ramp load changes with different slew rates, including 400 N·m/s, 2000 N·m/s, and 4000 N·m/s; 3) system steady-state and dynamic performance in four quadrants of operation: verify the symmetrical operation characteristics of the sensorless drive system; and 4) system dynamic performance under complete torque reversals: verify the ride through capability of the sensorless drive system during large load variations. Furthermore, experimental results for the sensorless drive system using the conventional DSMO (i.e., using a conventional discretized reaching law [25] and without the parameter adaption scheme in Fig. 3) under a torque ramp change and complete torque reversal with the highest slew rate of 4000 N·m/s are provided at the end of this section to further demonstrate the steady state and dynamic performance and stability improvement of the sensorless drive system using the proposed adaptive QSMO.

B. Steady-State Performance

In order to evaluate the zero-phase-lag behavior of the adaptive QSMO over the full load range, a set of torque ramp change tests are performed at the base speed by increasing the torque command linearly with the same slew rate of 400 N·m/s from zero to different steady-state values, as shown in Fig. 7. First, the parameters of the QSMO, l and Z_0 , are designed for the free-shaft condition. The QSMO with the fixed parameters (i.e., without the parameter adaption scheme) is used for sensorless control of the IPMSM for each torque ramp change test; the resulting position estimation errors are shown in Fig. 7 as well. The QSMO without the parameter adaption scheme can guarantee a zero phase lag under the free-shaft condition where the parameters are designed. However, phase lags (i.e., negative position

Fig. 7. Phase lags at different steady-state torque levels using the QSMO without the parameter adaption scheme.

Fig. 8. Position estimation errors showing zero-phase-lag behavior in torque ramp change tests using the adaptive QSMO.

estimation errors) are observed under other loading conditions. As shown in Fig. 7, the phase lag increases nonlinearly with the steady-state torque level. At the maximum torque, the phase lag reaches 50 electric degrees. As a comparison, the adaptive QSMO is also applied for sensorless control of the IPMSM for each torque ramp change test; and the resulting position estimation errors are shown in Fig. 8. The position estimation error always oscillates within $\pm 5^\circ$ around 0° and has no phase lags in any torque ramp change case.

Fig. 9 shows the performance of the proposed adaptive QSMO under free shaft under different speed conditions. The PWM frequency is maintained at 6 kHz to evaluate the impact of speed variations on the QSMO performance at a constant sampling frequency. Since the fundamental frequency of the extended EMF components increases proportionally with the speed but the sampling frequency is the same for different speed conditions, the number of calculation points per electrical revolution of the QSMO for the highest speed case (4500 r/min) in Fig. 9(d) is only 1/9 of that for the lowest speed case (500 r/min) in Fig. 9(a). Therefore, the estimated extended EMF components become more discontinuous when the speed increases. However, by using the parameter adaption scheme, the performance of the QSMO, as demonstrated by the position estimation errors in Fig. 9, has no degradation from low speed to high speed. It

Fig. 9. Experimental results of the estimated extended EMF components, estimated and measured rotor positions, and position estimation errors under different speeds when $f_s = 6$ kHz: (a) 500 r/min, (b) 1500 r/min, (c) 3000 r/min, and (d) 4500 r/min.

should be pointed out that the sampling frequency for the QSMO should be high enough to ensure accurate position estimation, but should not be a very large value for the sake of algorithm implementation. In practice, a reasonable sampling ratio between 15 and 20 can ensure acceptable position estimation accuracy, e.g., position estimation errors less than 4 electric degrees, for the QSMO, where the sampling ratio is defined to be the number of sampling instants per electrical revolution. This can be obtained from the testing results shown in Fig. 9.

C. Dynamic Performance Under Torque Ramp Changes

In this set of tests, positive ramp changes from zero to the maximum value of 300 N·m with different slew rates are applied to the torque command. Since the prime mover machine maintains a negative speed, i.e., $d\theta_{re}/dt < 0$, when the torque is positive, the IPMSM works in the braking mode as a generator. The current tracking performance, including the trajectories of the current commands i_d^* and i_q^* , as well as the actual currents i_d and i_q , is shown in Fig. 10 for three torque ramp change cases with the slew rate of 400, 2000, and 4000 N·m/s, respectively. In all scenarios, the sensorless drive system presents consistent

Fig. 10. Current tracking performance under three torque ramp change scenarios.

Fig. 11. Current trajectories for three torque ramp change scenarios.

steady-state current tracking errors. To observe the dynamic performance clearly, the trajectories of the actual currents corresponding to three different cases and the trajectory of the current command are shown in Fig. 11. Since the same PI gains are used for the feedforward current regulators in Fig. 5 for all cases, the system has a relatively larger current tracking error at the beginning for the torque ramp change case with a higher slew rate. However, all three current trajectories converge toward the command current trajectory and track the current command precisely. The initial stage, which is the area in the blue-dashed-line circle in Fig. 11, is critical to the sensorless drive, especially under fast changing load conditions [25]. In this region, the current regulation experiences a transient stage, which will further introduce oscillating error to the position estimation and may cause instability of the system. The proposed parameter adaptation scheme makes the QSMO to have zero-phase-lag behavior at different load levels. This ensures that the QSMO works in

the desired sliding surface regardless of the load conditions and system stability [25].

D. Four-Quadrant Operations

As shown in Fig. 5, lookup tables are used to generate command currents from torque command. The lookup tables are first generated from the MTPA profile and then tuned on the test stand to guarantee right operating points. For the test stand used in this study, the operating points for the motoring mode and braking mode are symmetrical in the lookup tables. Therefore, the sensorless drive system is expected to have symmetrical behavior under four-quadrant operations, where the four-quadrant operating conditions are defined as: Q1) Motoring with positive speed and positive torque; Q2) braking with negative speed and positive torque; Q3) motoring with negative speed and negative torque; Q4) braking with positive speed and negative torque.

In this set of tests, a ramp change with a slew rate of -4000 or 4000 N·m/s is applied to the torque command for each quadrant of operation. As Fig. 12 shows, the sensorless drive system is always stable, and the errors between the estimated and measured rotor positions have no steady-state offset for all the cases. The position estimation error is also in an acceptable range during the load transient of each case. The responses (i.e., speed and position estimation error) of the system in the two motoring modes (Q1 and Q3) and two braking modes (Q2 and Q4) are symmetrical with each other. However, the transient stages, i.e., the position tracking settling time of the QSMO, of the motoring modes are slightly longer than those of the braking modes. This is caused by the dc-bus voltage. In the braking modes, the dc-bus voltage is higher than 700 V. However, in the motoring modes, the prime mover machine does not have fast enough dynamic response to supply power needed for IPMSM motoring, which results in both relatively larger speed oscillations and dc-bus voltage drops. The dc-bus voltage drops will further affect the transient performance of the sensorless IPMSM drive in the motoring modes.

Fig. 12. Performance of the sensorless drive under four-quadrant operations.

E. Complete Torque Reversal

Complete torque reversal is always one of the toughest tests for evaluating the ride through capability of a sensorless drive system under a large load transient. In a complete torque reversal test, the fast changing load, the cross zero of torque, and sudden shaft speed change will significantly affect the performance of the sensorless control system. What is worse, if the IPMSM transmits from the full (i.e., maximum torque and base speed) braking mode to the full motoring mode, it will always consume dc power, which will cause a larger dc voltage drop than the case when the IPMSM transmits from the full motoring mode to the full braking mode. As discussed in Section V-D, this will introduce disturbances into the drive system and result in a relatively longer transient stage. If the sensorless drive system is not robust enough, instability will occur, which will easily trigger over current faults.

Fig. 13 shows the testing results for two cases of complete torque reversal, i.e., (a) from full motoring to full braking and (b) from full braking to full motoring, where the slew rate of the torque changes is $4000 \text{ N}\cdot\text{m/s}$. Because of the sudden change in the torque command, the shaft speed increases/drops 450 r/min in both cases. However, the position and speed estimations exhibit good ride through performance under complete torque reversals. Although the position estimation has a relatively large maximum error of 10 electric degrees in the transient, the estimated position converges toward the measured position quickly.

The dc-bus voltage in the case of full motoring to full braking transition is shown in Fig. 14. From 1.95 to 2.1 s, the command

Fig. 13. Performance of the sensorless drive under complete torque reversals: (a) from full motoring to full braking and (b) from full braking to full motoring.

Fig. 14. DC-bus voltage in the case of full motoring to full braking transition.

torque increases from -300 to $300 \text{ N}\cdot\text{m}$. Prior to 1.95 s, the dc-bus voltage is around 700 V , and then increases because electric power is fed back to the dc bus when the IPMSM is in the braking mode. When the dc-bus voltage reaches 750 V , the dc chopper turns ON and the dc voltage begins to drop. The dc chopper turns OFF when the dc voltage is below 725 V . This explains why the dc-bus voltage increases and decreases

Fig. 15. Performance of conventional DSMO-based sensorless drive under (a) torque ramp change and (b) complete torque reversal.

back and forth several times during the torque reversal. When the torque reaches steady state, the dc-bus voltage will drop to 700 V again.

F. System Performance Using Conventional DSMO

As a comparison, similar experiments, i.e., torque ramp change and complete torque reversal, are performed for the sensorless drive system using a conventional DSMO without the proposed parameter adaption scheme in Fig. 3. The torque command profiles and the resultant position estimation errors of the system for the torque ramp change and complete torque reversal tests under base speed are shown in Fig. 15(a) and (b), respectively. In the torque ramp change test, the torque command is increased linearly from 0 to 120 N·m with a constant slew rate of 4000 N·m/s. The parameters of the DSMO are tuned to guarantee a zero phase lag between the estimated and measured positions under the zero-torque condition. As shown in Fig. 15(a), the position estimation error has large oscillations during the torque transient stage, and phase lags are obvious. Although the position estimation error settles down after the torque command reaches the new steady-state value, there is an obvious phase lag around 10 electric degrees between the estimated and measured positions. In this case, due to the saliency of the IPMSM, without proper observer parameter adaption, a phase difference will present between the estimated and measured positions. If the torque is ramp changed to a higher value, e.g., 200 N·m, the system will lose stability due to a large phase lag. In the complete torque reversal test, the torque command is reduced linearly from 300 to −300 N·m with a constant slew rate

of 4000 N·m/s. The parameters of the DSMO are tuned to guarantee a zero phase lag between the estimated and measured positions when the torque is 300 N·m. The measured rotor position is first used in the drive system (i.e., a sensor-based drive system) to increase the output torque of the IPMSM to 300 N·m. Then, when the estimated rotor position is aligned with the measured rotor position, the drive system is switched to the closed-loop sensorless control. With fixed observer parameters, the sensorless drive system is able to produce 300 N·m torque at steady state. However, when the torque reversal occurs, instability is observed. The position estimation error diverges quickly, which triggered an overcurrent fault on the test stand.

V. CONCLUSION

An adaptive QSMO has been proposed for sensorless control of IPMSMs operating under medium- and high-speed conditions. The adaptive QSMO is robust to load variations and allows the state trajectory of the SMO to fast converge into the designed boundary layer around the sliding surface. The global stability and quasi-sliding-mode motion have been guaranteed using the proposed adaptive switching function. Experimental results have verified that the proposed QSMO with the linear parameter adaption schemes has good steady-state and transient performance over a wide speed and load ranges. The performance of the adaptive QSMO has no degradation even using relatively low sampling frequencies (e.g., 6 kHz) under high-speed and heavy-load conditions. As shown in the experimental results, the sensorless drive using the proposed adaptive QSMO presents excellent performance under ramp torque changes with different slew rates, symmetrical performance for four-quadrant operations, and excellent ride through capability under complete torque reversals. These capabilities however could not be achieved by using the conventional DSMO without the parameter adaption scheme. In practical applications, if a faster execution rate is used for the QSMO, e.g., using a field-programmable gate array-based controller, the performance of the sensorless drive system can be further improved.

APPENDIX

A. Inequality Derived From Stability Condition 1

According to (7) and (8), if $\varepsilon_\alpha[k] > Z_0$, then $Z_\alpha = Z_0$. Under this condition, $\varepsilon_\alpha[k+1] < \varepsilon_\alpha[k]$ needs to be satisfied. Thus

$$\varepsilon_\alpha[k+1] - \varepsilon_\alpha[k] = -\frac{T_s R}{L_d} \varepsilon_\alpha[k] + T_s E_\alpha[k] - T_s l Z_0 < 0 \quad (\text{A-1})$$

which is equivalent to $l Z_0 + \frac{R}{L_d} \varepsilon_\alpha[k] > E_\alpha[k]$.

Since $\varepsilon_\alpha[k] > Z_0 > 0$, $\frac{R}{L_d} \varepsilon_\alpha[k] > \frac{R}{L_d} Z_0$. Thus, if the following inequality is satisfied, (A-1) will also be satisfied

$$\left(l + \frac{R}{L_d}\right) Z_0 > E_\alpha[k]. \quad (\text{A-2})$$

If $\varepsilon_\alpha[k] < -Z_0$, then $Z_\alpha = -Z_0$. In this case, $\varepsilon_\alpha[k+1] > \varepsilon_\alpha[k]$ needs to be satisfied. Thus

$$\varepsilon_\alpha[k+1] - \varepsilon_\alpha[k] = -\frac{T_s R}{L_d} \varepsilon_\alpha[k] + T_s E_\alpha[k] + T_s l Z_0 > 0 \quad (\text{A-3})$$

which is equivalent to $l Z_0 - \frac{R}{L_d} \varepsilon_\alpha[k] > -E_\alpha[k]$.

Since $\varepsilon_\alpha[k] < -Z_0 < 0$, $-\frac{R}{L_d} \varepsilon_\alpha[k] > \frac{R}{L_d} Z_0$. Thus, if the following inequality is satisfied, (A-3) will also be satisfied

$$\left(l + \frac{R}{L_d}\right) Z_0 > -E_\alpha[k]. \quad (\text{A-4})$$

According to (A-2) and (A-4), since R/L_d is positive, a stronger condition can be obtained as

$$l Z_0 > |E_\alpha[k]| \quad (\text{A-5})$$

so that both (A-2) and (A-4) are satisfied, so as Condition 1.

B. Inequality Derived From Stability Condition 2

If $\varepsilon_\alpha[k] > Z_0$, then $Z_\alpha = Z_0$. In this condition, $\varepsilon[k+1] + \varepsilon[k] > 0$ needs to be satisfied. Thus

$$\varepsilon_\alpha[k+1] + \varepsilon_\alpha[k] = \left(2 - \frac{T_s R}{L_d}\right) \varepsilon_\alpha[k] + T_s E_\alpha[k] - T_s l Z_0 > 0 \quad (\text{A-6})$$

which is equivalent to $l Z_0 < \left(\frac{2}{T_s} - \frac{R}{L_d}\right) \varepsilon_\alpha[k] + E_\alpha[k]$.

Since $\varepsilon_\alpha[k] > Z_0 > 0$, $\left(\frac{2}{T_s} - \frac{R}{L_d}\right) Z_0 < \left(\frac{2}{T_s} - \frac{R}{L_d}\right) \varepsilon_\alpha[k]$. Thus, if the following inequality is satisfied, (A-6) will also be satisfied:

$$l Z_0 < \left(\frac{2}{T_s} - \frac{R}{L_d}\right) Z_0 + E_\alpha[k]. \quad (\text{A-7})$$

If $\varepsilon_\alpha[k] < -Z_0$, then $Z_\alpha = -Z_0$. In this condition, $\varepsilon[k+1] + \varepsilon[k] < 0$ needs to be satisfied. Thus

$$\varepsilon_\alpha[k+1] + \varepsilon_\alpha[k] = \left(2 - \frac{T_s R}{L_d}\right) \varepsilon_\alpha[k] + T_s E_\alpha[k] + T_s l Z_0 < 0 \quad (\text{A-8})$$

which can be formulated as: $l Z_0 < -\left(\frac{2}{T_s} - \frac{R}{L_d}\right) \varepsilon_\alpha[k] - E_\alpha[k]$.

Since $\varepsilon_\alpha[k] < -Z_0 < 0$, $\left(\frac{2}{T_s} - \frac{R}{L_d}\right) Z_0 < -\left(\frac{2}{T_s} - \frac{R}{L_d}\right) \varepsilon_\alpha[k]$. Thus, if the following inequality is satisfied, (A-8) will also be satisfied:

$$l Z_0 < \left(\frac{2}{T_s} - \frac{R}{L_d}\right) Z_0 - E_\alpha[k]. \quad (\text{A-9})$$

According to (A-7) and (A-9), a stronger condition can be obtained as

$$l Z_0 < \left(\frac{2}{T_s} - \frac{R}{L_d}\right) Z_0 - |E_\alpha[k]| \quad (\text{A-10})$$

so that both (A-7) and (A-9) are satisfied, so as Condition 2.

In order to satisfy both Conditions 1 and 2, both (A-5) and (A-10) should be used, which draws (9-I). To guarantee the existence of $l Z_0$, the upper boundary in (A-10) should be larger than the lower boundary in (A-5), which draws (9-II).

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