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Robust Observer with Higher-order Sliding Mode for Sensorless Speed Estimation of a PMSM

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Abstract— **This paper addresses the problem of sensorless speed estimation for the permanent magnet synchronous motor (PMSM). A higher-order sliding mode (HSM) observer is developed to provide estimation of back electro motive forces (EMFs) that are treated as unknown inputs without the use of lowpass filter. A finite-time smooth estimation is obtained and the chattering phenomenon is eliminated. An accurate speed estimate of PMSM can be algebraically computed with the estimated back EMFs. Experimental results show the effectiveness of the proposed method.**

Index Terms— **permanent magnet synchronous motor (PMSM), higher-order sliding mode (HSM), sensorless control, speed estimation.**

NOMENCLATURE

ω_e	Rotor electrical speed
i_α, i_β	Currents in stationary reference frame
V_α, V_β	Voltages in stationary reference frame
e_α, e_β	Back EMFs in stationary reference frame
R	Stator resistance
L	Synchronous inductance
K_E	EMF constant
θ_e	Rotor position angle

I. INTRODUCTION

The permanent magnet synchronous motor (PMSM) is very popular in ac motor applications for various speed controls. It has a more complex control system than a direct-current (dc) motor. A high performance PMSM drive relies on field oriented or vector control and this requires precise rotor position [2]–[4]. In most variable-speed drive systems, the rotor position is measured by an optical encoder mounted on the shaft or a resolver. However, the usage of this sensor creates several disadvantages from the standpoint of drive cost, weight, reliability, noise, and is very sensitive to environmental constraints such as vibration and temperature [5]. To overcome these disadvantages, a sensorless control method has been developed for control of the motor using the estimated values of the position and speed of the rotor [2]–[4], [6].

Currently, there are several sensorless methods available in the literature [6]–[8]. There are two kinds of approaches depending on the speed operating range required by the application: magnetic saliency methods and estimation of variables using state observers. In magnetic saliency methods, the rotor position estimation is done through the injection of proper test signals [9]–[11]. These methods are relatively difficult to implement but they offer a proper solution for both standstill and low speed operation. State observers require the measured electrical quantities (applied voltages and currents) to estimate the rotor position and/or speed and are preferred for medium/high speed operation.

Sliding mode observers (SMO) have been developed to robustly estimate the system states through the concepts of sliding surface design and equivalent control [7]. SMO's have been successful in the estimation of unknown inputs, faults [12]–[14]. Several first-order sliding mode based methods for the rotor position and speed estimation in industrial drives are available in the literature [1], [15], [16]. In conventional first-order sliding mode observer, a low-pass filter and an additional position compensation for the rotor are required to reduce the chattering phenomenon due to the usage of signum function [14], [15], [17]. A survey was recently conducted in [18] for the implementation of the sliding mode control. In [15], the cross-coupling terms of the $d-q$ current dynamics are treated as unknown disturbances. Decoupling terms with improper parameters can slightly degrade the system's performance. Integral sliding mode (ISM) controller with a switching output was proposed in [17] to overcome these disturbances. As a result, ISM can guarantee the robustness of the system starting from the initial time instance. However, the speed estimation based on ISM current control requires an additional low-pass filter, which introduces the delay, and which in turn reduces the system's phase margin and can cause instability. Recently, a high-speed SM observer [19] was proposed for sensorless speed estimation in a PMSM. A sigmoid function is employed instead of the switching function to avoid the low-pass filter. However, the selection of sliding mode gains and boundary layer are not straight forward as they are dependent on the rotor speed. The method is more suitable for constant speed applications.

Recently, higher-order SM observers have been developed to overcome the disadvantages of first-order SM observer [20]–[26]. The super–twisting algorithm (STA) (see [24]) provides finite time and exact convergence, even in the presence of bounded perturbations. To analyze the robustness of the STA for a wider class of disturbances, strict Lyapunov functions are developed in [20]. This Lyapunov function makes some additional modifications of the super-twisting algorithm (STA) by including terms which improves its robustness and convergence properties [20], [21]. Also, it can reduce the well-known chattering phenomenon that is presented in the neighborhood of the sliding manifold.

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In this paper, a modification of STA is used to design an appropriate HSM observer. The observer enables the estimation of the rotor position and speed of the PMSM in real-time while removing the well-known chattering phenomenon. In HSM scheme, the unknown back EMFs are estimated using the robust terms designed by the sliding mode observer from current dynamics of the PMSM. With the back EMFs accurately estimated, the rotor position and speed can be obtained algebraically. The proposed technique does not require any low-pass filtering and hence has no delay in the estimation. The proposed method is experimentally verified on a PMSM.

The rest of the paper is organized as follows. Modeling of PMSM is discussed in Section 2. Section 3 designs the higher-order sliding mode observer based on super-twisting algorithm for speed estimation. Section 4 presents the experimental results. Section 5 concludes the paper.

Through out this paper, $\lambda_{max}(A)$ denotes the maximum eigenvalue of a matrix A, $\|A\|$ denotes the 2-norm $\sqrt{\lambda_{max}(A^TA)}$ of A. $\lambda_{min}(A)$ represents its minimum singular value.

II. PMSM MODELING AND PROBLEM STATEMENT

The model of the PMSM, in the stationary reference frame, can be described by the following system:

$$\left\{ \begin{array}{lcl} \frac{di_\alpha}{dt} &=& \frac{-R}{L}i_\alpha - \frac{1}{L}e_\alpha + \frac{1}{L}V_\alpha \\ \frac{di_\beta}{dt} &=& \frac{-R}{L}i_\beta - \frac{1}{L}e_\beta + \frac{1}{L}V_\beta \end{array} \right. \tag{1}$$

with

$$\left\{ \begin{array}{lcl} e_\alpha &=& -K_E\omega_e \sin\theta_e \\ e_\beta &=& K_E\omega_e\cos\theta_e \end{array} \right. \tag{2}$$

Assuming that the motor speed changes slowly, we have $\dot{\omega}_e \approx 0$. The state equations of the PMSM in the stationary reference frame together with dynamics of back EMFs can be obtained as:

$$\left\{ \begin{array}{lcl} \frac{di_\alpha}{dt} &=& \frac{-R}{L}i_\alpha - \frac{1}{L}e_\alpha + \frac{1}{L}V_\alpha \\ \frac{di_\beta}{dt} &=& \frac{-R}{L}i_\beta - \frac{1}{L}e_\beta + \frac{1}{L}V_\beta \\ \frac{de_\alpha}{dt} &=& -\omega_e e_\beta \\ \frac{de_\beta}{dt} &=& \omega_e e_\alpha \end{array} \right. \tag{3}$$

The voltages V_α, V_β and currents i_α, i_β are the known quantities. The objective is to design an observer to estimate the back EMFs using the available measurements.

III. HIGH ORDER SLIDING MODE OBSERVER

Let us note that the extra terms (e_α, e_β) of the plant dynamics in (3) act like unknown inputs. The idea is to use sliding modes on both α and β axes to reconstruct these unknown inputs. In the following, a HSM observer will be designed to estimate the unknown inputs.

Applying the same design principles as for variable structure control, the observer trajectories are constrained to evolve after a finite time on a suitable sliding manifold by the use of a discontinuous output injection signal (the sliding manifold is usually given by the difference between the observer and the system output). Hence, the sliding motion provides an estimate (asymptotically or in finite time) of the system states.

A. Generalities

We denote by $t \to e(t, e_0)$ a solution of the system:

$$\left\{ \begin{array}{lcl} \dot{e}(t) &=& g(e(t)) \\ e(t_0) &=& e_0 \end{array} \right. \tag{4}$$

where $e = \left[\begin{array}{ccc} e_1 & \cdots & e_n \end{array} \right]^T \in \mathcal{R}^n$, $g : \mathcal{M} \mapsto \mathcal{R}^n$ is continuous on an open neighborhood $\mathcal{M}$ of the origin $e = 0$.

The following result gives a sufficient condition for system (4) to be finite time stable:

Lemma 1: ([27]) Let the origin be an equilibrium point for system (4). If there is a Lyapunov function $V : \mathcal{M} \to \mathcal{R}^+$ such that the time derivative along the solution of (4) satisfies

$$\dot{V}(e) \leq -cV^a(e) \tag{5}$$

with $a \in (0, 1)$ and $c > 0$. Then, the origin is finite time stable.

Lemma 2: ([27]) Let us consider the following system:

$$\left\{ \begin{array}{lcl} \dot{e}(t) &=& -\lceil e(t) \rfloor^\alpha \\ e(0) &=& e_0 \end{array} \right. \tag{6}$$

where $\text{sign}(e) = [\text{sign}(e_1), ..., \text{sign}(e_n)]^T$, $|e|^\alpha = \text{diag}(|e_1|^\alpha, ..., |e_n|^\alpha)$ and $\lceil e \rfloor^\alpha = |e|^\alpha \text{sign}(e)$ with $\alpha \in (0, 1)$, system (6) is continuous everywhere and locally Lipschitzian everywhere except at the origin. The solutions converge to zero in finite time.

B. Finite-time stability

For ease of exposition, consider the following system:

$$\left\{ \begin{array}{lcl} \dot{e}(t) &=& -ae(t) + \nu(t) + \delta(e,t) \\ e(t_0) &=& e_0 \end{array} \right. \tag{7}$$

where $e \in \mathcal{R}$ and a is a known positive constant, $\delta(e, t)$ is the unknown input/perturbation and

$$\nu(t) = -K_1\phi_1(e(t)) - K_2 \int_0^t \phi_2(e(t))dt \tag{8}$$

and

$$\phi_1(e(t)) = e(t) + K_3|e(t)|^{\frac{1}{2}} \text{ sign}(e(t)) \tag{9}$$

$$\begin{aligned} \phi_2(e(t)) &= e(t) + \frac{K_4^2}{2}\text{sign}(e(t)) \\ &+ \frac{3}{2}K_4|e(t)|^{\frac{1}{2}} \text{ sign}(e(t)) \end{aligned} \tag{10}$$

K_1, K_2, K_3 and K_4 are appropriately designed positive constants.

Assumption 1: The time derivative of the unknown input/perturbation is upper bounded as follows:

$$\|\dot{\delta}(e, t)\| \leq \rho \tag{11}$$

for a positive constant ρ.

Proposition 1: Under Assumption 1, the origin of system (7) is a finite time stable equilibrium point. Further, the finite time smooth estimation of the unknown input/perturbation $\delta(e,t)$ is given by $K_2 \int_0^t \phi_2(e(t))dt$.

Proof: Since $\|\phi_2(e)\| \geq \frac{K_4^2}{2}$, one gets:

$$\|\dot{\delta}(e,t)\| \leq \|\phi_2(e)\| \tag{12}$$

if

$$K_4 \quad \geq \quad \sqrt{2\rho} \tag{13}$$

Let us select a Hurwitz matrix A_0:

$$A_0 = \begin{bmatrix} -(K_1+a) & 1 \\ -K_2 & 0 \end{bmatrix}$$

where $K_1 > 0$ and $K_2 > 0$.

The system (7), (8) can be equivalently represented by the system of two first order equations:

$$\begin{cases} \dot{e}_1 & = & e_2 - (K_1+a)\left(e_1 + K_4\lceil e_1 \rfloor^{\frac{1}{2}}\right) \\ \dot{e}_2 & = & -K_2\left(e_1 + \frac{K_4^2}{2}\text{sign}(e_1) + \frac{3}{2}K_4\lceil e_1\rfloor^{\frac{1}{2}}\right) + \dot{\delta} \end{cases} \tag{14}$$

with $e_1 = e$, $e_2 = \delta - K_2\int_0^t \phi_2(e_1)dt$ and

$$K_4 = \frac{K_1 K_3}{K_1 + a}$$

The solutions of the discontinuous differential equations and inclusions are understood in the sense of Filippov..
Let us consider the new state vector:

$$\xi = \begin{bmatrix} \xi_1 \\ \xi_2 \end{bmatrix} = \begin{bmatrix} \phi_1(e_1) \\ e_2 \end{bmatrix} \tag{15}$$

Using the differential equations inclusion theory, the derivative of Eq. (15) is derived (see [20]). System (14) can be rewritten as:

$$\begin{aligned} \dot{\xi} &= \begin{bmatrix} \left(1+\frac{K_4}{2}|e_1|^{\frac{-1}{2}}\right)(e_2 - (K_1+a)\phi_1(e_1)) \\ -K_2\phi_2(e_1) + \dot{\delta} \end{bmatrix} \\ &= \left(1+\frac{K_4}{2}|e_1|^{\frac{-1}{2}}\right)\begin{bmatrix} \xi_2 - (K_1+a)\xi_1 \\ -K_2\xi_1 \end{bmatrix} + \begin{bmatrix} 0 \\ \dot{\delta} \end{bmatrix} \\ &= \left(1+\frac{K_4}{2}|e_1|^{\frac{-1}{2}}\right)(A_0\xi) + \begin{bmatrix} 0 \\ \dot{\delta} \end{bmatrix} \end{aligned}$$

The stability analysis of system (14) is performed using the following candidate Lyapunov function [28]:

$$V(\xi) = \xi^T P \xi \tag{16}$$

with $P = P^T = \begin{bmatrix} \lambda + 4\epsilon^2 & -2\epsilon \\ -2\epsilon & 1 \end{bmatrix}$, $\lambda > 0$ and $\epsilon > 0$. It is worth noting that the matrix P is positive definite if λ and ϵ are any real number.

Remark 1: Since $V(\xi)$ is continuous but not locally Lipschitz, the usual versions of Lyapunov Theorem cannot be used. However, it is possible to show that $V(\xi)$ is an absolutely continuous function along the system trajectories. It implies that it is differentiable almost everywhere, and on those points the derivative can be calculated in the usual way. Moreover, if the derivative $\dot{V}$ is negative definite almost everywhere, then V is monotone decreasing and converges to zero, what is the condition required by Zubov's Theorem [29].

Its time derivative along the solutions of the system is given by:

$$\dot{V} \quad = \quad \left(1+\frac{K_4}{2}|e_1|^{\frac{-1}{2}}\right)\xi^T\left(A_0^T P + P A_0\right)\xi + 2\xi^T P\begin{bmatrix} 0 \\ \dot{\delta} \end{bmatrix}$$

From Eq. (12), one can note that:

$$\left\|2\xi^T P\begin{bmatrix} 0 \\ \dot{\delta} \end{bmatrix}\right\| \leq \left\|\left(1+\frac{K_4}{2}|e_1|^{\frac{-1}{2}}\right)2\xi^T P\begin{bmatrix} 0 \\ \xi_1 \end{bmatrix}\right\|$$

It can be shown that:

$$\begin{aligned} \dot{V} &\leq \left(1+\frac{K_4}{2}|e_1|^{\frac{-1}{2}}\right)\left(\xi^T P((A_0^T + A_0)\xi + 2\begin{bmatrix} 0 \\ \xi_1 \end{bmatrix})\right) \\ &\leq -\left(1+\frac{K_4}{2}|e_1|^{\frac{-1}{2}}\right)\xi^T Q \xi \\ &\leq -\left(1+\frac{K_4}{2}|e_1|^{\frac{-1}{2}}\right)\lambda_{min}(Q)\|\xi\|^2 \end{aligned}$$

with

$$Q = \begin{bmatrix} Q_1 & Q_2 \\ Q_2 & Q_3 \end{bmatrix} \tag{17}$$

$$\begin{cases} Q_1 & = & 2(K_1+a)(\lambda+4\epsilon^2) - 4\epsilon(K_2-1) \\ Q_2 & = & -2\epsilon(K_1+a) + (K_2+1) - (\lambda+4\epsilon^2) \\ Q_3 & = & 4\epsilon \end{cases} \tag{18}$$

In order to guarantee the positive definiteness of matrix Q, one chooses:

$$K_2 = \lambda + 4\epsilon^2 + 2\epsilon(K_1+a) \tag{19}$$

The matrix Q is positive definite if:

$$K_1 > -a + \frac{4\epsilon + 2\epsilon\lambda + 8\epsilon^3}{\lambda} + \frac{1}{4\epsilon\lambda} \tag{20}$$

From Eq. (15), one can deduce that:

$$\begin{aligned} ||\xi||^2 &= \xi_1^2 + \xi_2^2 \\ &= e_1^2 + 2K_3|e_1|^{\frac{3}{2}} + K_3^2|e_1| + \xi_2^2 \\ &\geq K_3^2|e_1| \end{aligned}$$

Since $K_3 > 0$

$$-\frac{K_3}{||\xi||} \quad \geq \quad -|e_1|^{-\frac{1}{2}}$$

It implies that:

$$\dot{V} \leq -\frac{\lambda_{min}(Q)}{\lambda_{max}^{\frac{1}{2}}(P)}\frac{K_3^2}{2}V^{\frac{1}{2}} - \frac{\lambda_{min}(Q)}{\lambda_{max}(P)}V$$

The closed-loop system (14) is stabilized in finite time. Since $\|\xi\|$ converges to zero in finite time, e_1 and e_2 converge to 0. Therefore, the term $K_2\int_0^t \phi_2(e(t))dt$ gives in finite time a smooth estimation of the unknown perturbation $\delta(e,t)$. ■

C. High order sliding observer design

Let us define the sliding mode observer for PMSM currents as follows:

$$\begin{cases} \frac{d\hat{i}_\alpha}{dt} &= \frac{-R}{L}\hat{i}_\alpha + \frac{1}{L}V_\alpha - \frac{1}{L}\nu_1(t) \\ \frac{d\hat{i}_\beta}{dt} &= \frac{-R}{L}\hat{i}_\beta + \frac{1}{L}V_\beta - \frac{1}{L}\nu_2(t) \end{cases} \tag{21}$$

The robust terms are given by

$$\begin{cases} \nu_1(t) &= -K_1\phi_1(s_\alpha(t)) - K_2\int_0^t \phi_2(s_\alpha(t))dt \\ \nu_2(t) &= -K_1\phi_1(s_\beta(t)) - K_2\int_0^t \phi_2(s_\beta(t))dt \end{cases} \tag{22}$$

where the sliding surfaces are defined as $s_\alpha(t) = \hat{i}_\alpha - i_\alpha$, $s_\beta(t) = \hat{i}_\beta - i_\beta$ and $\phi_1(\cdot), \phi_2(\cdot)$ are defined in (9) $-$ (10).

Assumption 2: At least locally, there are positive constants ρ_1 and ρ_2 such that the following terms are bounded as follows:

$$\begin{cases} \|\omega_e e_\alpha\| &\leq \rho_1 \\ \|\omega_e e_\beta\| &\leq \rho_2 \end{cases} \tag{23}$$

for some positive constant ρ_1 and ρ_2.
Assumption 2 is not restrictive since w_e, e_α and e_β are continuous on a compact set.

Proposition 2: Under Assumption 2, the estimating errors s_α and s_β are stabilized towards zero in finite time. A smooth estimation of the unknown back EMFs is given as follows:

$$\begin{cases} \hat{e}_\alpha &= K_2\int_0^t \phi_2(s_\alpha(t))dt \\ \hat{e}_\beta &= K_2\int_0^t \phi_2(s_\beta(t))dt \end{cases} \tag{24}$$

Proof: The time derivatives of the sliding surface are as follows:

$$\begin{cases} \dot{s}_\alpha &= \frac{-R}{L}s_\alpha - \frac{1}{L}V_1(t) + \frac{1}{L}e_\alpha \\ \dot{s}_\beta &= \frac{-R}{L}s_\beta - \frac{1}{L}V_2(t) + \frac{1}{L}e_\beta \end{cases} \tag{25}$$

According to Prop. 1, the origin of system (25) is a finite time stable equilibrium. In sliding mode, $s_\alpha = \dot{s}_\alpha = 0$ and $s_\beta = \dot{s}_\beta = 0$.

Hence, the reduced order dynamics of system (25) becomes:

$$\begin{cases} 0 &= -K_2\int_0^t \phi_2(s_\alpha(t))dt + e_\alpha \\ 0 &= -K_2\int_0^t \phi_2(s_\beta(t))dt + e_\beta \end{cases} \tag{26}$$

■

Using the estimated back EMFs, the position of the rotor can be calculated as

$$\hat{\theta}_e = -\tan^{-1}\left(\frac{\hat{e}_\alpha}{\hat{e}_\beta}\right) \tag{27}$$

Also, with estimated back EMFs, using equation (2), the speed can be computed algebraically as

$$\hat{\omega}_e = \frac{1}{K_E}\sqrt{\hat{e}_\alpha^2 + \hat{e}_\beta^2} \tag{28}$$

The speed estimate uses the EMF constant K_E and the estimated back EMFs $\hat{e}_\alpha$, $\hat{e}_\beta$. Compared to the proposed HSM observer scheme with [14], no low-pass filtering is required.

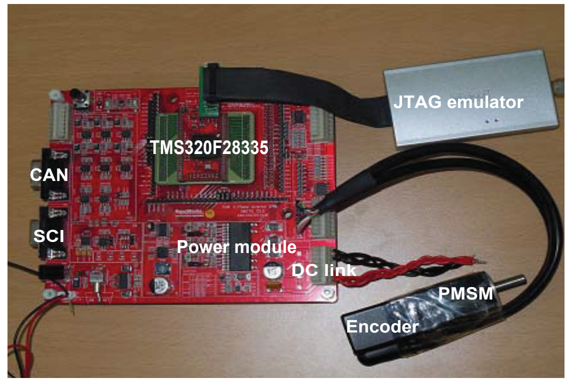


Fig. 1. Experimental setup.

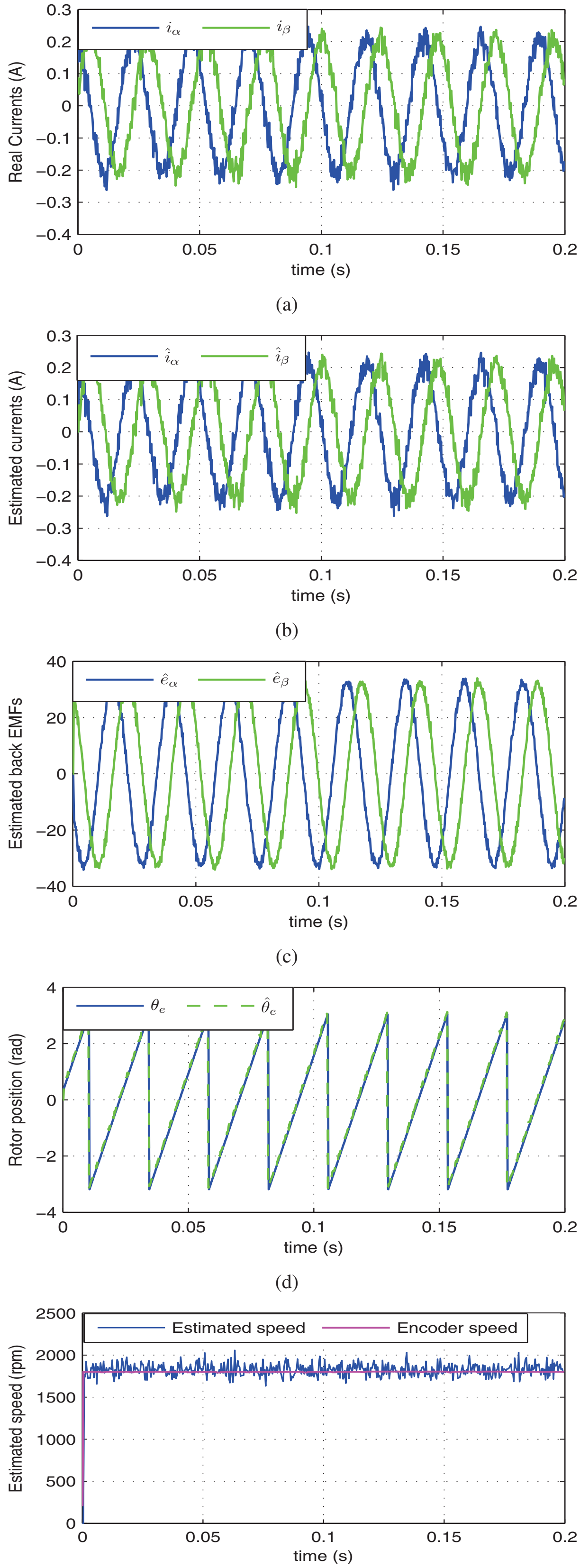
IV. EXPERIMENTAL RESULTS

The motor used in the experimental setup is a TBL-i model TS4632N2050E510 3-phase PMSM. The PMSM is powered by a Fairchild FSB50325S smart power module which includes 6 fast-recovery MOSFET (FRFET) inverters and 3 half-bridge high voltage integrated circuits (HVICs) for FRFET gate driving. Since, it employs FRFET as a power switch, it has much better robustness and larger safe operation areas (SOA) than that using an IGBT-based power module or one-chip solution. The experimental setup is shown in Fig. 1. The switching frequency of the PWM inverter is 15 kHz. SMC 75 Evaluation Module with TMS320F28335 DSP controller is used. It contains Texas Instruments 32-bit floating point DSP as well as analog interfaces and JTAG emulator port. The board has analog-to-digital converter (A/D) with 16 channels. All the control variables are monitored using graph window of Code Composer Studio (CCS v3.3) after being converted to analog signals through the digital-to-analog (D/A) converter. Real motor speed (ω_e) is measured using a high-resolution incremental encoder with 2000 pulses/rotation and the estimated speed ($\hat{\omega}_e$) is obtained with the proposed HSM scheme. The specifications and parameters are provided in Table I.

TABLE I
SPECIFICATIONS OF PMSM

Rating		26 [W]
Speed	ω	4000 [r/min]
Stator resistance	R	2.0 [Ω]
Stator inductance	L	0.51 [mH]
Back EMF constant	K_E	0.156 [Vs/rad]
Number of poles	P	8

The stator currents of the PMSM are measured from the current sensors and they are sent to TMS320F28335 via A/D converters. In the same way, stator voltages are calculated using dc-bus voltage sensors and duty cycles of the inverter when the switching functions are known. Three-phase currents and voltages are transformed to two-phase stationary ($\alpha - \beta$) reference frame. They are again transformed to rotating ($d{-}q$) reference frame for the control design.



(a)

(b)

(c)

(d)

(e)

Fig. 2. Speed estimation using the proposed HSM observer. (a) i_α and i_β. (b) $\hat{i}_\alpha$ and $\hat{i}_\beta$. (c) $\hat{e}_\alpha$ and $\hat{e}_\beta$. (d) θ_e and $\hat{\theta}_e$. (e) Estimated speed, $\hat{\omega}_e$.

Two experiments have been performed to validate the proposed HSM scheme. For the first experiment, a constant reference speed of 1800 rpm is chosen. While in the second one, the motor speed is varied from 1000 rpm to 2000 rpm. In the first experiment, the motor operates in speed control mode with a constant speed reference of 0.45 p.u.(1 p.u. = 4000 revolutions/min). The reference current in $d-$ axis is set at a constant value, i.e. $i_d^{ref} = 0$. The reference current in q axis i_q^{ref} is generated by a speed PI controller. Fig. 2(a) shows the real currents (i_α, i_β) and Fig. 2(b) shows the estimated $(\hat{i}_\alpha,\ \hat{i}_\beta)$ currents. Real and estimated currents are very similar in both magnitude and phase with the proposed HSM observer. Fig. 2(c) depicts the estimated back EMFs obtained using (24). Despite the noisy currents, the back EMFs are relatively smooth. Fig. 2(d) shows the real and estimated rotor positions. The estimated rotor position is robust with respect to noise measurements and exactly matches with the real rotor position without any phase delay. So, the estimated rotor position may be used instead of the measured one in the vector control of PMSM drive. Fig. 2(e) shows the estimated speed for a speed reference of 0.45 p.u using the proposed HSM method.

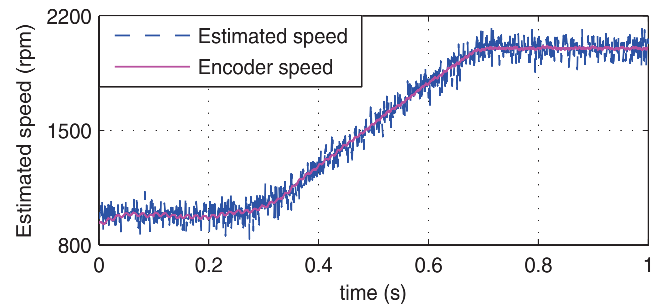


Fig. 3. HSM estimated speed.

In the second experiment, to test the robustness of the proposed method, the motor speed is varied from 0.25-0.5 p.u. The estimated rotor speed obtained through estimated back EMFs is depicted in Fig. 3. The estimated speed tracks the reference speed accurately. However due to presence of noise measurements in currents, the estimated speed also contains noise. The HSM speed estimate was obtained with (28) without low-pass filtering.

From the implementation, one can conclude that:

- The HSM method requires the proper selection of sliding mode gains K_1, K_2, K_3 and K_4. The sliding mode gains should satisfy the conditions given by Eqs. (19)−(20) for the desired speed range. If the motor operates in wide speed range, the sliding mode gains must be appropriately selected.
- For all selections of the sliding mode gains, the method may not yield good estimation of back EMFs in real-time. Since the quality of the speed estimate highly depends on the estimated back EMFs, it deteriorates when more noisy back EMFs (obtained due to high gains) are used in the calculation. Further, the calculation of

speed also involves the motor constant K_E that can be uncertain.
- Its worth to point out that, the proposed method is computationally complex. However, if properly tuned, it has more advantages. The experiments conducted in this paper validates the advantages of this method.

V. CONCLUSION

This paper has presented a sensorless speed estimation method for the PMSM drive. The HSM method is based on a modified version of super–twisting algorithm. The observer dynamics consist of robust terms to reconstruct the unknown back EMFs from the sliding mode. The speed is then analytically computed from back EMFs. The HSM speed estimator obtained has a good steady-state accuracy and acceptable dynamic behavior. The proposed method has good robustness to noise and no delay in estimation. Further, lowpass filtering is not required with the proposed method. The theoretical development has been validated by experimental results.

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