INITIAL ROTOR POSITION ESTIMATION FOR LOW SALIENCY INTERIOR
PERMANENT-MAGNET SYNCHRONOUS MOTOR DRIVES

by

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ABSTRACT

This work presents an initial rotor position estimation method for low saliency interior permanent-magnet synchronous motor (IPMSM) drives. The method injects signals into the stationary α - β frame of the motor and substitutes the injected signals into the stationary α - β frame model of the motor to solve for the rotor position. In particular, the method injects specific signals into the motor to eliminate the rotor-position-independent terms while keeping the rotor-position and motor-saliency dependent terms in the motor model. As a result, the rotor-position and motor-saliency dependent terms are invulnerable to errors or noise in the rotor-position-independent terms and therefore when used to solve for the rotor position, lead to accurate results in rotor position estimation. Experimental results show that the position estimation error is less than 5° electrical for a low saliency IPMSM whose d-axis and q-axis inductance are 1.65mH and 1.70mH respectively.

INTRODUCTION

Interior Permanent Magnet Motors and their Control

In recent two decades, interior permanent magnet synchronous motors (IPMSMs) are gaining popularity for various industrial applications. They have various advantages over traditional motors, such as higher torque density, higher efficiency, lower torque ripple, lower noise and lower maintenance requirement [1]-[23].

Generally, an IPMSM is controlled with the field orientation control method, or so-called vector control method [1]-[14], [17].

In the rotating d-q frame, the torque is given as:

$$T = k_T [\varphi_f i_q + (L_d - L_q) i_d i_q] \quad (1)$$

Where:

k_T is a constant dependent on the construction of the motor;

i_d and i_q are the stator current in the d-axis and the q-axis respectively;

L_d and L_q are the stator inductance in the d-axis and the q-axis respectively,

φ_f is the permanent magnet flux linkage.

Equation (1) shows that i_d and i_q can be used for controlling T.

Figure 1 shows the close-loop speed control algorithm of IPMSMs. In the block diagram, the reference rotor speed ω_{ref} is compared with the actual rotor speed ω obtained by the position sensor attached to the shaft of the rotor. Then the speed error is fed into a PI controller to get the reference q-axis current, which is also the torque

reference as introduced before. The reference d-axis current is set to be zero to simplify the control algorithm. Then the current in d-axis and q-axis are compared with the actual currents transformed from the three phase ABC frame into the d-q frame via two transformations, Clarke transformation (the ABC frame to the stationary α - β frame) and Park transformation (the stationary α - β frame to the rotating d-q frame). The resultant current errors in d-axis and q-axis are then fed into two PI controllers, respectively, to get reference voltage in d-q frame. Finally, the voltages in d-axis and q-axis are transformed into the stationary α - β frame by applying inverse Park transformation and applied to control the three phase inverter with space vector PWM control algorithm.

Figure 1: Closed-loop speed control algorithm of IPMSMs.

In the field orientation method, one can see that the knowledge of the instantaneous rotor position is required for implementing the control algorithm. The

instantaneous rotor position is not only required at the running state but also at the start-up. Conventionally, the initial rotor position information is obtained from a mechanical sensor attached to the shaft of the rotor. The mechanical sensor adds additional cost, size and weight to the motor drive, and compromises reliability. All these limitations make sensorless control technique extremely desirable.

Previous Sensorless Control Work

If we do not do sensorless control at the start-up, usually we inject three-phase balanced current into the motor and expect the motor would follow the rotating magnet field created by the current. However, the current may sometimes create positive torque, and sometimes create negative torque, and the average torque would be still very small even if we inject high amplitude current. As a result, the motor may not start up if the load torque is high at standstill. In order to meet the requirement of high-performance start-up, sensorless control at initial state is required.

One simple approach to obtain the initial rotor position without a mechanical sensor is to apply a known dc current to the stator side for certain amount of time, and the stator side would generate a resultant constant flux linkage which would attract the magnet of the rotor to approach the direction of the flux linkage [1]. As a result, the rotor is aligned to a known position. However, with this dc current excitation method, the direction of the rotation of the rotor is unpredictable. This limit makes the method not suitable for many applications in which the rotor is not allowed to rotate back and forth.

More advanced sensorless techniques are called signal injection methods [1]-[20], [22], [23]. These methods can obtain the initial rotor position without causing the rotor to rotate. Over the last two decades, numerous sensorless techniques based on high frequency signal injection at standstill have been developed for IPMSMs. These methods can be generally classified into two categories based on whether the methods are implemented in the stationary α - β frame or the rotating d-q frame.

Rotating d-q Frame Based Methods.

This category of sensorless techniques are based on the mathematic model of the IPM motor in the rotating d-q frame [1]-[14]. The methods estimate the initial rotor position by injecting rotating [1]-[8] or pulsating [7]-[13] signals. The injected signals can be either voltage signals [1]-[12] or current (flux) signals [13].

For the rotating signal injection methods, the injected high frequency signals interact with the spatial saliency of the motor and produce the signal components that contain the rotor position information. Then filters are applied to obtain only the position-error-dependent components from those components. After that, certain control algorithms are utilized to make the position-error-dependent components approach near-zero values. In this case, the position error is considered to be near zero.

For example, in reference [7], a high frequency rotating voltage signal is injected into the d-q frame as:

$$v_{dqsc}^s = V_c e^{j\omega_c t} \quad (2)$$

Where:

ω_c is the carrier frequency;

v_{dqsc}^s is high frequency injected signal;

V_c is the amplitude of the injected voltage.

The injected high frequency voltage signal interacts with the spatial saliency of the motor and produces the current components i_{dqsc}^s that contain rotor position information, as given in equation (3).

$$i_{dqsc}^s \approx I_{c_p1} e^{j(\omega_c t - \frac{\pi}{2})} + I_{c_n1} e^{j(\omega_c t + 2\theta_r + \frac{\pi}{2})} + I_{c_p2} e^{j(2\omega_c t - \theta_r + \varphi_{2p})} + I_{c_n2} e^{j(-2\omega_c t + 3\theta_r + \varphi_{2n})} \quad (3)$$

Where:

I_{c_p1} , I_{c_n1} , I_{c_p2} , I_{c_n2} are complex coefficients related to V_c , ω_c , saliency and many other parameters of the motor;

φ_{2p} , φ_{2n} is related to the rotor flux linkage and phase resistance;

θ_r is the rotor angular position.

A spatial saliency tracking observer, driven by the error, is developed, as given in equation (4), which is obtained after so-called heterodyning process that extracts the negative sequence term.

$$\varepsilon \approx I_{c_n1} \sin[2(\theta_r - \hat{\theta}_r)] \quad (4)$$

Where:

$\hat{\theta}_r$ is the estimated rotor angular position.

The control algorithm makes ε approach a near zero value. In this case, $(\theta_r - \hat{\theta}_r)$ is considered to be almost zero, i.e., $\hat{\theta}_r = \theta_r$. However, ε is also proportional to I_{c_n1} , which is related to the saliency of the motor. If the saliency is very small, I_{c_n1} may have a small value. In this case, $(\theta_r - \hat{\theta}_r)$ might be large even if ε is very small.

In reference [8], similarly, a high frequency rotating voltage signal, as shown in equation (2), is injected into the d-q axis. The interaction between the carrier frequency voltage and the saliency of the motor produces a carrier frequency current, which is sent to a high pass filter to extract the desired rotor position information, as given in equation (5).

$$\varepsilon = \text{HPF}(i_{dqsc}^s) \approx \frac{2V_c}{\omega_c} \frac{(L_d - L_q)}{(L_d + L_q)^2 - (L_d - L_q)^2} \sin[2(\theta_r - \hat{\theta}_r)] \quad (5)$$

Where:

L_d and L_q are inductance on the d-axis and the q-axis, respectively.

Similarly, a control algorithm makes ε approach a near zero value, i.e., $\hat{\theta}_r = \theta_r$. However, that ε approaches a near zero value doesn't necessarily suggest that $\hat{\theta}_r$ is close to θ_r , because ε is also proportional to the difference between L_d and L_q , i.e., $(L_d - L_q)$, which is almost zero if the saliency is very small. In this case, $(\theta_r - \hat{\theta}_r)$ could be large even if ε is very small.

The pulsating signal injection methods inject pulsating signals into the d-axis or q-axis in the rotating d-q frame, and the injected pulsating signals interact with the spatial saliency of the motor and produce signal components which contain the rotor position

information. Then certain control algorithms are developed to make these position-error-dependent components approach near-zero values.

For example, in reference [11], a 550Hz high frequency pulsating voltage signal is injected only into the d-axis as:

$$v_{dsc}^s = V_c \cos(\omega_c t) \quad (6)$$

$$v_{qsc}^s = 0 \quad (7)$$

Then the high frequency current in q-axis is obtained as:

$$i_{qsc}^s = \left(-\frac{1}{2}\right) \frac{v_{dsc}^s}{L_d L_q} (L_d - L_q) \sin[2(\theta_r - \hat{\theta}_r)] \quad (8)$$

The q-axis high frequency current in equation (8) is used as the input signal to the spatial saliency tracking observer. i_{qsc}^s becomes zero when the rotor position estimation error is zero, as stated in reference [11]. However, that i_{qsc}^s becomes zero doesn't necessarily suggest that $\hat{\theta}_r$ is close to θ_r , because i_{qsc}^s is also proportional to the difference between L_d and L_q , i.e., $(L_d - L_q)$, which is almost zero if the saliency is very small. In this case, $(\theta_r - \hat{\theta}_r)$ could be large even if i_{qsc}^s is very small.

In reference [12], a 2kHz high frequency pulsating flux vector signal is injected only into the estimated q-axis as shown in equation (9) and equation (10).

$$\lambda_{dsc}^s = 0 \quad (9)$$

$$\lambda_{qsc}^s = \frac{V_c \sin(\omega_c t)}{\omega_c} \quad (10)$$

Then the high frequency current in d-axis is obtained as:

$$i_{dsc}^s = \frac{V_c \sin(\omega_c t)}{\omega_c} \frac{(L_q - L_d)}{(L_d + L_q)^2 - (L_d - L_q)^2} \sin[2(\theta_r - \hat{\theta}_r)] \quad (11)$$

From equation (11), one can see that i_{dsc}^s is amplitude modulated by the error between the estimated position $\hat{\theta}_r$ and the actual position θ_r . i_{dsc}^s is first demodulated to form a signal that is proportional to the position error, and then the error is fed into an observer that forces the error to be zero. Similarly, $(\theta_r - \hat{\theta}_r)$ could be large even if i_{dsc}^s is very small.

Stationary α - β Frame Based Methods.

The second category of the methods injects high frequency (generally 250Hz to 2kHz sinusoidal) or certain PWM signals into the stationary α - β frame and uses resultant mathematical model in the stationary α - β frame to solve for the rotor position [15]-[23].

The methods inject signals into the stationary α - β frame of the motor and substitute the injected signals into the stationary α - β frame model of the motor to yield motor models that contain rotor-position-dependent terms, which also depend on the saliency of the motor. The rotor-position-dependent terms are then used to solve for the rotor position. However, the rotor-position-dependent terms are also dependent on the saliency of the motor. When the saliency is small, the rotor-position-dependent terms are small compared to other rotor-position-independent terms. As a result, any small errors in the rotor-position-independent terms can cause large errors in the rotor-position-dependent terms and therefore lead to large errors in rotor position estimation.

For example, in reference [15], the following high frequency signals are injected into α -axis:

$$i_{\alpha sc}^s = I_c \cos(\omega_c t) \quad (12)$$

$$i_{\beta sc}^s = 0 \quad (13)$$

Where ω_c is usually selected to be high enough (higher than 650Hz) to ensure that the rotor cannot rotate, and the current amplitude I_c is set to be small enough to avoid magnetic saturation.

The phase difference δ_d between $i_{\alpha sc}^s$ and $v_{\alpha sc}^s$ is a function of the rotor position θ_r , as shown in equation (14).

$$\delta_d = \tan^{-1} \frac{\omega_c(L_d \cos^2 \theta_r + L_q \sin^2 \theta_r)}{r} \quad (14)$$

Where r is the phase resistance.

After exciting the motor in the α -axis and detecting δ_d , as described above, a high frequency current is injected into the β -axis as shown in equation (15) and (16).

$$i_{\alpha sc}^s = 0 \quad (15)$$

$$i_{\beta sc}^s = I_c \cos(\omega_c t) \quad (16)$$

The phase difference δ_q between $i_{\beta sc}^s$ and $v_{\beta sc}^s$ is a function of the rotor position θ_r , as shown in equation (17).

$$\delta_q = \tan^{-1} \frac{\omega_c(L_d \sin^2 \theta_r + L_q \cos^2 \theta_r)}{r} \quad (17)$$

Solving (14) and (17) with respect to θ_r , the estimated value of θ_r can be described as:

$$\hat{\theta}_r = \pm \tan^{-1} \sqrt{\frac{k \tan \delta_d - \tan \delta_q}{k \tan \delta_q - \tan \delta_d}} \quad (18)$$

if $\tan \delta_d \leq \tan \delta_q$

$$\hat{\theta}_r = \frac{\pi}{2} \pm \tan^{-1} \sqrt{\frac{k \tan \delta_d - \tan \delta_q}{k \tan \delta_q - \tan \delta_d}} \quad (19)$$

if $\tan\delta_d > \tan\delta_q$

Where $k = \frac{L_q}{L_d}$.

The polarity identification technique is utilized to obtain a unique solution for θ_r . The sizable difference in the d-axis inductance and the q-axis inductance makes it possible to obtain a solution based on equation (18) and (19). If the saliency of the motor is very small, i.e., L_d is very close to L_q , k is almost 1. As a result, equation (18) becomes:

$$\hat{\theta}_r = \pm \tan^{-1} \sqrt{\frac{\tan\delta_d - \tan\delta_q}{\tan\delta_q - \tan\delta_d}} \quad (20)$$

In addition, when the saliency is small, one can substitute $L_d \approx L_q$ into equation (14) and (17) to obtain equation (21) as:

$$\delta_q \approx \delta_d \quad (21)$$

One can see from equation (20) and (21), solving $\hat{\theta}_r$ would lead to a large error because $\tan\delta_d$ and $\tan\delta_q$ are very close.

The Proposed Method

This project presents an initial rotor position estimation method for low saliency IPMSM drives. Low saliency IPMSM drives are introduced in [24]. The method injects signals into the stationary α - β frame of the motor and substitutes the injected signals into the stationary α - β frame model of the motor to solve for the rotor position. In particular, the method uses specific injected signals to eliminate the rotor-position-independent

terms while keeping the rotor-position and motor-saliency dependent terms in the resultant motor model. As a result, the errors in the rotor-position-independent terms will not cause errors in the rotor-position and motor-saliency dependent terms and therefore will not cause errors in rotor position estimation. Experimental results show the position estimation error is less than 5 °electrical for a small saliency IPMSM whose d-axis and q-axis inductance are 1.70mH and 1.65mH respectively.

PRINCIPLE

Mathematic Model of IPMSMs in the d-q FrameClarke and Inverse Transformation.

Assuming i_a , i_b , i_c are the instantaneous balanced three-phase stator currents:

$$i_a + i_b + i_c = 0 \quad (22)$$

Then it is possible to define the stator current space vector as:

$$i_{sv} = i_a + ai_b + a^2i_c \quad (23)$$

Where: $a = e^{\frac{2\pi}{3}j}$; $a^2 = e^{\frac{4\pi}{3}j}$.

The space vector i_{sv} defined in a-b-c frame in equation (23) can be expressed in α - β frame utilizing two-axis theory: the sum of the real-part in a-b-c frame is equal to the value of the α -axis component i_α and the imaginary part is equal to the value of the β -axis component i_β , as shown in Figure 2.

Figure 2: Stator current space vector in the a-b-c and the α - β frame.

Thus, the stator current space vector can be expressed in the stationary α - β frame as:

$$i_{sv} = i_{\alpha} + j i_{\beta} \quad (24)$$

In symmetric 3-phase IPMSMs,

$$i_a + i_b + i_c = 0 \quad (25)$$

Then, i_{α} and i_{β} are related to the three phase stator current as follows:

$$i_{\alpha} = i_a - \frac{1}{2} i_b - \frac{1}{2} i_c \quad (26)$$

$$i_{\beta} = \frac{\sqrt{3}}{2} (i_b - i_c) \quad (27)$$

By multiplying the transformation constant $\frac{2}{3}$ on the right side of equation (26) and (27), one can get the Clarke transformation as:

$$\begin{bmatrix} i_{\alpha} \\ i_{\beta} \end{bmatrix} = M \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \quad (28)$$

$$M = \frac{2}{3} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\ 0 & 0 & 0 \end{bmatrix} \quad (29)$$

Where M is the Clarke transformation matrix.

The inverse Clarke transformation can be obtained as:

$$\begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} = M^{-1} \begin{bmatrix} i_{\alpha} \\ i_{\beta} \end{bmatrix} \quad (30)$$

Where M^{-1} is the inverse matrix of M.

Park and Inverse Transformation.

The stator current space vector in the a-b-c, the α - β and the d-q frame is shown in Figure 3 where the d-q frame is rotating with the synchronous speed; θ is the angle between α -axis and d-axis.

Figure 3: Stator current space vector in the a-b-c, the α - β and the d-q frame.

The equations corresponding to Park transformation is given as:

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} = N \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} \quad (31)$$

$$N = \begin{bmatrix} \cos(\theta) & \sin(\theta) \\ -\sin(\theta) & \cos(\theta) \end{bmatrix} \quad (32)$$

Where N is the Park transformation matrix.

The inverse Park transformation can be obtained as:

$$\begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} = N^{-1} \begin{bmatrix} i_d \\ i_q \end{bmatrix} \quad (33)$$

Where N^{-1} is the inverse matrix of N .

Mathematic Model of IPMSMs in α - β frame.

The mathematic model of IPMSMs in the rotating d-q frame [1]-[12] is given as:

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \begin{bmatrix} R + pL_d & \omega L_q \\ \omega L_d & R + pL_q \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} 0 \\ \omega \varphi_f \end{bmatrix} \quad (34)$$

Where:

v_d and v_q are the stator voltage in the d-axis and the q-axis respectively;

R is the stator resistance;

ω is the rotor angular speed;

$p=d/dt$.

The rotor position, θ , is defined as the angle between the stationary α - β frame and the rotating d-q frame, as depicted in Figure 4. In this paper, electrical degree is used as the unit of θ .

Figure 4: Rotating d-q reference frame with respect to the stationary α - β reference frame.

Transforming equations (34) into the stationary α - β reference frame yields:

$$\begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix} = R \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \begin{bmatrix} L_0 + L_1 \cos(2\theta) & L_1 \sin(2\theta) \\ L_1 \sin(2\theta) & L_0 - L_1 \cos(2\theta) \end{bmatrix} p \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} \\ + 2\omega L_1 \begin{bmatrix} \sin(2\theta) & \cos(2\theta) \\ \cos(2\theta) & \sin(2\theta) \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \omega \psi_f \begin{bmatrix} \sin \theta \\ \cos \theta \end{bmatrix} \quad (35)$$

Where:

$$L_0 = \frac{L_d + L_q}{2}, L_1 = \frac{L_d - L_q}{2};$$

v_α and v_β are the stator voltage in the α -axis and the β -axis respectively;

i_α and i_β the stator voltage in the α -axis and the β -axis respectively;

θ is the rotor angular position.

Since the motor is at standstill, Substituting $\omega=0$ into (35) yields:

$$\begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix} = R \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + L_0 p \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + L_1 \begin{bmatrix} \cos(2\theta) & \sin(2\theta) \\ \sin(2\theta) & -\cos(2\theta) \end{bmatrix} p \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} \quad (36)$$

The low-saliency characteristics of the IPMSMs make it very challenging to propose an initial sensorless method for the motor. In equation (36), the last term

$L_1 \begin{bmatrix} \cos(2\theta) & \sin(2\theta) \\ \sin(2\theta) & -\cos(2\theta) \end{bmatrix} p \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix}$ on the right hand side contains the rotor position

information θ and can be used to solve for θ , as shown in the following section . However,

L_1 is very small compared to R in the first term and L_0 in the second term on the right hand side of equation (36). Therefore, any small error in R and L_0 would cause large error in rotor position estimation. In order to solve θ , the first and second term on the right hand side of equation (36) need to be eliminated.

Proposed Sensorless Method.

The proposed sensorless method injectes pulse signals to the stationary α - β frame to solve for the rotor position. To obtain the solution, the method uses two steps, or two injections to detect the rotor position.

In the first step, current is injected into phase B and phase C while no current is injected into phase A, i.e.:

$$i_a = 0, i_b = -i_c \quad (37)$$

Applying the Clarke transformation yields:

$$\begin{bmatrix} I_\alpha \\ I_\beta \end{bmatrix} = \frac{2}{3} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} I_a \\ I_b \\ I_c \end{bmatrix} \quad (38)$$

Thus, we can solve i_α as:

$$\begin{aligned} i_\alpha &= \frac{2}{3} \left(I_a - \frac{1}{2} I_b - \frac{1}{2} I_c \right) \\ &= -\frac{1}{3} (I_b + I_c) \\ &= 0 \end{aligned} \quad (39)$$

Substituting equation (39) into the first equation of (36) yields,

$$\begin{aligned} v_\alpha &= R \times 0 + L_0 \times 0 + L_1 \cos(2\theta) \times 0 + L_1 \sin(2\theta) \frac{di_\beta}{dt} \\ &= L_1 \sin(2\theta) \frac{di_\beta}{dt} \end{aligned} \quad (40)$$

We can see that in the first step of high frequency injection, we eliminate the R and L_0 related terms by making i_α zero and finally get a simple relationship between v_α and $\frac{di_\beta}{dt}$.

Transforming v_α into the a-b-c frame yields:

$$\begin{aligned}
 v_\alpha &= \frac{2}{3} \left(v_{an} - \frac{1}{2} v_{bn} - \frac{1}{2} v_{cn} \right) \\
 &= \frac{1}{3} (2v_{an} - v_{bn} - v_{cn}) \\
 &= \frac{1}{3} (2v_{an} - 2v_{cn} + v_{cn} - v_{bn}) \\
 &= \frac{1}{3} (2v_{ac} - v_{bc})
 \end{aligned} \tag{41}$$

Transforming i_β into the a-b-c frame yields:

$$i_\beta = \frac{\sqrt{3}}{3} (i_b - i_c) \tag{42}$$

Substituting equation (37) into (42), one can have:

$$i_\beta = \frac{2\sqrt{3}}{3} i_b \tag{43}$$

Substituting equation (41) and (42) into (40), one can have:

$$(2v_{ac} - v_{bc}) = 2\sqrt{3}L_1 \sin(2\theta) \frac{di_b}{dt} \tag{44}$$

Solving θ from equation (44) yields:

$$\sin(2\theta) = \frac{\sqrt{3}}{6} \left(\frac{2v_{ac} - v_{bc}}{L_1 \frac{di_b}{dt}} \right) \tag{45}$$

In (45), v_{ac} and v_{bc} can either be calculated from the DC link voltage of the motor drive inverter and the gating signals of the switches in the inverter or be sensed with voltage sensors; i_b can be sensed with a current sensor. As a result, $\sin(2\theta)$ can be calculated using (45) and then used to solve for θ . However, using only the $\sin(2\theta)$ value to solve for θ will lead to multiple solutions. This ambiguity issue necessitates the second signal injection, which is described below.

The above injection is done by proper control of the motor drive inverter shown in Figure 5. To ensure $i_a=0$, Q_1 and Q_2 are turned off; to inject i_b and i_c into the motor, Q_3 and Q_6 are turned on while Q_4 and Q_5 are turned off for a period of T , to apply the DC link voltage between phase b and c in this period and therefore build up a current that flows in phase b and phase c. Afterwards, Q_3 and Q_6 are turned off while Q_4 and Q_5 are turned on for a period of T , to apply the DC link voltage between phase c and phase b and therefore bring the current back to zero.

Figure 5: Current flow in the first injection.

In the second step, a high frequency current is injected from phase A to phase B while no current is injected into phase C, i.e.:

$$i_c = 0, i_a = -i_b \quad (46)$$

Applying Clarke Transformation yields:

$$\begin{aligned} \begin{bmatrix} I_\alpha \\ I_\beta \end{bmatrix} &= \frac{2}{3} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & 1 & -1 \end{bmatrix} \begin{bmatrix} I_a \\ I_b \\ I_c \end{bmatrix} \\ &= \frac{2}{3} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} I_a \\ -I_a \\ 0 \end{bmatrix} \\ &= \frac{2}{3} \begin{bmatrix} \frac{3}{2} I_a \\ -\frac{\sqrt{3}}{2} I_a \end{bmatrix} \\ &= \begin{bmatrix} I_a \\ -\frac{\sqrt{3}}{3} I_a \end{bmatrix} \end{aligned}$$

Or:

$$\begin{cases} i_\alpha = i_a \\ i_\beta = -\frac{\sqrt{3}}{3} i_a \end{cases} \quad (47)$$

Substituting (47) into equations (36), one can have:

$$\begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix} = R \begin{bmatrix} i_a \\ -\frac{\sqrt{3}}{3} i_a \end{bmatrix} + L_0 p \begin{bmatrix} i_a \\ -\frac{\sqrt{3}}{3} i_a \end{bmatrix} + L_1 \begin{bmatrix} \cos(2\theta) & \sin(2\theta) \\ \sin(2\theta) & -\cos(2\theta) \end{bmatrix} p \begin{bmatrix} i_a \\ -\frac{\sqrt{3}}{3} i_a \end{bmatrix} \quad (48)$$

Separating (48) into 2 equations, one can have:

$$v_\alpha = R \times i_a + L_0 \frac{di_a}{dt} + L_1 \cos(2\theta) \frac{di_a}{dt} - \frac{\sqrt{3}}{3} L_1 \sin(2\theta) \frac{di_a}{dt} \quad (49)$$

$$v_\beta = -R \times \frac{\sqrt{3}}{3} i_a - \frac{\sqrt{3}}{3} L_0 \frac{di_a}{dt} + L_1 \sin(2\theta) \frac{di_a}{dt} + \frac{\sqrt{3}}{3} L_1 \cos(2\theta) \frac{di_a}{dt} \quad (50)$$

Multiplying $\sqrt{3}$ on both sides of equation (50) yields:

$$\sqrt{3}v_\beta = -R i_a - L_0 \frac{di_a}{dt} + \sqrt{3}L_1 \sin(2\theta) \frac{di_a}{dt} + L_1 \cos(2\theta) \frac{di_a}{dt} \quad (51)$$

Adding equation (49) and (51), one can have:

$$\begin{aligned} v_\alpha + \sqrt{3}v_\beta &= L_1 \cos(2\theta) \frac{di_a}{dt} - \frac{\sqrt{3}}{3} L_1 \sin(2\theta) \frac{di_a}{dt} + \sqrt{3}L_1 \sin(2\theta) \frac{di_a}{dt} + L_1 \cos(2\theta) \frac{di_a}{dt} \\ &= 2L_1 \cos(2\theta) \frac{di_a}{dt} + \frac{2\sqrt{3}}{3} L_1 \sin(2\theta) \frac{di_a}{dt} \\ &= [2 \cos(2\theta) + \frac{2\sqrt{3}}{3} \sin(2\theta)] L_1 \frac{di_a}{dt} \\ &= \frac{4}{\sqrt{3}} [\frac{\sqrt{3}}{2} \cos(2\theta) + \frac{1}{2} \sin(2\theta)] L_1 \frac{di_a}{dt} \\ &= \frac{4}{\sqrt{3}} [\sin(\frac{\pi}{3}) \cos(2\theta) + \cos(\frac{\pi}{3}) \sin(2\theta)] L_1 \frac{di_a}{dt} \\ &= \frac{4}{\sqrt{3}} \sin(2\theta + \frac{\pi}{3}) L_1 \frac{di_a}{dt} \end{aligned} \quad (52)$$

Note that the terms that contains R and L_0 are eliminated in equation (52).

Transforming v_α and v_β into a-b-c frame yields:

$$v_\alpha = \frac{2}{3} \left(v_{an} - \frac{1}{2} v_{bn} - \frac{1}{2} v_{cn} \right) \quad (53)$$

$$v_\beta = \frac{2}{3} \left(\frac{\sqrt{3}}{2} v_{bn} - \frac{\sqrt{3}}{2} v_{cn} \right) \quad (54)$$

Combining equation (53) and (54), one can have:

$$\begin{aligned} v_\alpha + \sqrt{3}v_\beta &= \frac{2}{3} \left(v_{an} - \frac{1}{2} v_{bn} - \frac{1}{2} v_{cn} \right) + \frac{2}{3} \left(\frac{3}{2} v_{bn} - \frac{3}{2} v_{cn} \right) \\ &= \frac{2}{3} \left(v_{an} - \frac{1}{2} v_{bn} - \frac{1}{2} v_{cn} + \frac{3}{2} v_{bn} - \frac{3}{2} v_{cn} \right) \end{aligned}$$

$$\begin{aligned}
&= \frac{2}{3} (v_{an} + v_{bn} - 2v_{cn}) \\
&= \frac{2}{3} (2v_{an} - v_{an} + v_{bn} - 2v_{cn}) \\
&= \frac{2}{3} (2v_{ac} - v_{ab})
\end{aligned} \tag{55}$$

Transforming equation (55) into a-b-c frames and solve for θ yields:

$$\frac{2}{3} (2v_{ac} - v_{ab}) = \frac{4}{\sqrt{3}} \sin\left(2\theta + \frac{\pi}{3}\right) L_1 \frac{di_a}{dt} \tag{56}$$

Rearranging equation (56), one can have:

$$\sin\left(2\theta + \frac{\pi}{3}\right) = \frac{\sqrt{3}}{6} \left(\frac{2v_{ac} - v_{ab}}{L_1 \frac{di_a}{dt}} \right) \tag{57}$$

Similar to the $\sin(2\theta)$ value in (45), the $\sin(2\theta + \pi/3)$ value in (57) can be calculated using the phase voltage and current of the motor. The value can then be combined with the $\sin(2\theta)$ value calculated from (45) to lead to a unique solution for θ .

Figure 6: Current flow in the second injection.

The above current injection is done by the below inverter control shown in Figure

6. Q_5 and Q_6 are turned off to ensure $i_c=0$; Q_1 and Q_4 are turned on while Q_2 and Q_3 are

turned off for a period of T , to apply the DC link voltage between phase a and phase b in this period and therefore build up a current in phase a and phase b, and afterwards, Q_1 and Q_4 are turned off while Q_2 and Q_3 are turned on for a period of T , to apply the DC link voltage between phase b and phase a and therefore bring the current back to zero.

Noted that T is selected to be $200\mu\text{s}$, much smaller than the time constant of the motor (50ms, for example) to ensure the rotor will not move during the signal injection.

HARDWARE SETUP

The block diagram of the hardware setup is depicted in Figure 7.

Figure 7: Block diagram of the hardware setup.

The IPM motor is driven by a voltage source inverter, which uses the insulated gate bipolar transistors (IGBTs) as switching devices. The DC link voltage and phase currents signals are available at one of the connectors of the inverter. The DC link voltage of the inverter is 300V. TTL pulse signals are sent from the dSpace system, transformed into CMOS signal which has 15V supply voltage and then fed into the inverter. The phase currents are available as voltage signals in the inverter and they are amplified through an inverting operational amplifier to gain better resolution. The line-to-line voltage of the motor is measured using an isolated voltage transducer LV25-800 from LEM Company. Then the resultant DC link voltage, phase currents and the line-to-line voltage are sent to dSpace via 16-bit ADC channels on CP1104 board to solve for the rotor position based

on the proposed algorithm. The estimated rotor position is compared with the real rotor position obtained from the encoder attached to the shaft of the rotor.

IPMSM Motor

The IPMSM motor used for the experiment is a 6-pole low salient motor shown in Figure 8. The stator is Y-connected.

Figure 8: The IPMSM motor.

Table 1: Parameters of the IPMSM motor.

<i>Parameters</i>	<i>Value</i>
Polar pairs	3
Voltage	110Vac
Current	5A
Inductance in d-axis	1.65mH
Inductance in q-axis	1.70mH
Nominal Power	2HP
Phase resistance	0.5 Ω
Nominal torque	360lb-in
Nominal speed	600RPM

The parameters of the motor are given in Table 1.

Figure 9: Phase-phase inductance with relation to inductance in d-axis and q-axis.

For an IPM motor, as shown in Figure 9, the inductance in d-axis and q-axis [25] are given by:

$$L_q = \frac{\max(L_{ab})}{2}$$

$$L_d = \frac{\min(L_{ab})}{2}$$

Decoded Real Rotor Position from the Incremental Sensor in Dspace

An embedded incremental encoder with a resolution of 1024 pulse/resolution provides the real rotor position, as shown in Figure 10. The encoder has 8 wires and 5 wires of them are used to get the rotor position. Descriptions of the wires are given in Table 2.

Figure 10: The embedded incremental encoder.

Table 2: Description of the encoder wires used.

<i>Color</i>	<i>Signals</i>
Red	A
Black	B
White	Index
Yellow	GND
Brown	+5VDC

In Table 2, we can see that the encoder is operating on +5VDC voltage with power supplied from yellow and brown wire.

Signal A (red) and signal B(black), which are called quadrature outputs, are 90 degrees out of phase, as shown in Figure 11. As a result, four logic states, i.e., a, b, c and d are defined. The encoder offers 1024 pulses per resolution by means of A, B, and the index signal. Thus, A or B can generate $1024/4=256$ pulses. Also, the rotating direction can be identified with the pulse sequence. In another word, if B is leading A by 90 degrees, it indicates the reverse revolution, or counter-clockwise rotation; if A is leading

B by 90 degrees, it indicates the normal revolution, or clockwise rotation. After one full revolution, the index signal would send out a pulse signal and reset the counter to zero.

Figure 11: A and B signal in quadrature.

The encoder signals are sent to Dspace system via the incremental encoder interface to be decoded in order to obtain the real rotor position information, as shown in Figure 12.

Generally, Dspace system provides the encoder signals together with the corresponding inverted signal to represent the differential input. In another word, the encoder interface, which is “Inc” in Figure 12, consists of signals A, B, and the index signal and the corresponding inverted signal $\bar{A}$, $\bar{B}$ and $\bar{\text{index}}$. Signal A and the inverted signal $\bar{A}$ make two differential inputs. Using the differential inputs, the signal integrity, noise immunity. Thus, system reliability are tremendously improved.

However, as mentioned before, the encoder signals from the motor only provide the signals A, B, and the index. The corresponding pins for $\bar{A}$, $\bar{B}$ and index are left unconnected.

Figure 12: Input encoder signals to Dspace system.

The DS1104 is equipped with a 24-bit counter. Due to the 4-fold subdivision of each encoder line, the software related counter are 4 times slower than the hardware counter. The DS1104 allows the encoder to measure lines up to 2^{22} in the range of -2^{21} to $+2^{21} - 1$. This is sufficient for the encoder embedded in the motor which has only 256 encoder lines for A or B.

In order to decode the encoder information and transform them into electrical degrees in the range of $[0^\circ, 360^\circ)$, a real-time simulink-based model in Matlab/DS1104 is built, as shown in Figure 13.

The DS1104ENC_POS_C2 block is used to read the θ and $\Delta\theta$ value of the second encoder channel, which is connected to the embedded encoder in the motor. In order for this block to work normally, a master setup block (DS1104ENC_SETUP) must be placed

in the model. DS1104ENC_SETUP block is used to set the global parameters for the encoder channel. The dialog setting is mainly to define whether it is a differential signal type or a single-ended TTL signal type. Because no inverted signals are present for the encoder, we choose single-ended TTL signal.

Figure 13: Decoding model for encoder signals in Matlab/DS1104.

The position from DS1104ENC_POS_C2 is given in the encoder lines. In addition, this RTI block allows for initialization of the encoder position at the beginning of the simulation.

Note that the position value is represented in $\frac{1}{4}$ lines. For example, when the rotor rotates one mechanical cycle, or $360^\circ \times 3$ electrical degree, there will be 1024×3 pulses, or $1024 \times 3/4$ lines. If we need to represent the rotor position in electrical degree, the encoder needs to be reset for three times in one mechanical cycle. The switch

block and logical operator are used to reset the value when the rotor rotates every 360 degree.

DS1104ENC_HW_INDEX_C2 block is used to handle the encoder index interrupt. When the rising edge of the index signal from the encoder is detected, the corresponding position counter would be reset by hardware without any additional time delay, which is realized by sending a trigger signal to the DS1104ENC_SET_POS_C2 block.

Figure 14: Decoded encoder signal in Matlab/DS1104.

To test whether the decoded encoder signals work properly, the rotor was manually rotated clockwise first and then counter-clockwise. Figure 14 shows the decoded encoder signal waveform in electrical degree.

PWM Signal from Dspace

The important characteristics of the I/O unit on the DS1104 board are listed as follows:

- 20-bit digital I/O;
- Direction selectable for each channel individually;
- TTL voltage range for both input and output.

A simplified circuit diagram of the I/O input/output is shown in Figure 15. All digital I/O lines, with pull-up 10k Ω resistors connected to +5V VCC. The I/O channels are in the input mode by default. If “1” is sent to the Output Enable pin, the I/O channels are set to be in the output mode and therefore can be used to send out PWM signals to the inverter.

Figure 15: Simplified circuit diagram of the I/O input/output.

The digital I/O connector on CP1104 connector panel is a 37-pin, male Sub-D connector shown in Figure 16. To control the inverter, the digital I/O needs to be set in

the output mode and PWM signals are sent to each digital I/O bit using DS1104BIT_OUT_Cx, allowing for selection of a channel number within the range from 0 to 19. By default, the block is shown as DS1104BIT_OUT_C0. For example, if the channel number parameter of the block is selected as 3, the block would automatically be renamed as DS1104BIT_OUT_C3.

The relation between the digital output and the simulation input of the block is shown in Table 3.

Table 3: Relation between the digital output and the simulation input

Simulink Input		Digital Output(TTL)
Without data typing	With data typing	
>0 (double)	1(boolean)	high
<=0(double)	0(boolean)	low

In this work, by converting the electrical PWM signal sent from the PWM generator block in simulink to boolean signal, we control the signals sent to the digital I/O.

Figure 16: Digital I/O connector on CP1104 connector panel.

The IGBT Inverter

The inverter used in the experiment is SEMIKRON integrated SKiiP 2 type with the model number of 342GD120-314CTV, which is shown in Figure 17.

The SKiiP system integrates semiconductor switches, a heat sink, and a gate driver unit with the protection and monitoring units. It is equipped with closed loop current sensors and a temperature sensor. The signals of the current sensors are used for protection purpose, such as short circuit and over-current protection. Analog voltage signals of the actual AC current value and the DC-link voltage value are available at the DIN41651 connector, near the AC side terminals of the inverter.

Figure 17: The inverter: 342GD120-314CTV.

TTL to CMOS Interface Circuit

The PWM signal sent from the digital I/O is of TTL type. However, the signal needed for the inverter is CMOS type. Therefore, an interface circuit transforming TTL signals to CMOS signals is needed.

Figure 18 shows the interface circuit. There are three IR4427 chips on the board, used to transform six TTL PWM input to CMOS output signals.

Figure 18: TTL to CMOS interface board.

The IR4427 is a low voltage, high speed power IGBT driver. Logic inputs are compatible with CMOS outputs. The primary features needed are listed as follows:

- Fixed supply voltage V_s ranges from 6V to 20V.
- Outputs in phase with inputs.

Figure 19 shows the typical connection for IR442x series.

INA: A channel input signal (TTL);

INB: B channel input signal (TTL);

OUTA: A channel output signal (CMOS);

OUTB: B channel output signal (CMOS).

Figure 19: Typical connection of IR442x series.

For proper operation the device should be used within the recommended conditions, Table 4 shows the DC electrical characteristics values. V_{IH} represents the logic “1” of the input voltage and V_{IL} the logic “0”. We can see the input voltage are typical TTL signals.

Table 4: DC electrical characteristics.

Symbol	Min.	Max.
V_{IH}	2.7V	N/A
V_{IL}	N/A	0.8V

Figure 20 shows the functional block diagram of IR4427. For IR4427, the output signals are in phase with inputs, as shown in Figure 21, for IR4426 and IR4428, the outputs are out of phase with inputs.

Figure 20: Function block diagram of IR4427.

Figure 21: Timing diagram of IR442x series.

Analog Signals Sent to Dspace

Phase Current and DC Link Voltage.

The rotor initial position is calculated based on the line-to-line voltage, phase currents and the DC link voltage.

342GD120-314CTV inverter integrates current sensor per ceramic substrate to measure the AC output current. The measured current at DIN 41651 connector is normalized: $8V \leftrightarrow 300A$. The peak value of the motor phase current can be calculated as 7A based on the parameters given in Table 1. In a short time period like several microseconds, the motor can tolerant up to 10A current without causing any saturation to the magnet. Assuming the phase current is 1A, the measured current value provided by the inverter is calculated as:

$$\frac{0.1 \times 8}{300} \approx 0.02667 \text{ (V)} \quad (58)$$

As introduced before, the 16-bit analog I/O input of the dSpace system provides a resolution of 65536, the input voltage range is from -10V to 10V, and the maximum quantization error can be calculated as:

$$\frac{20}{65536} \approx 0.0003 \text{ (V)} \quad (59)$$

The phase current error due to the quantization error is given as:

$$\frac{0.0003}{0.02667} \approx 1.1\% \text{ (V)} \quad (60)$$

The lower the phase current, the more sensitive the phase current is to the impact of the quantization error. Taking the 0.1A input current for example, the phase current error caused by the quantization error would be as high as 11%.

In order to reduce the current error and thus overcome the small SNR problem, the faint current signals from the inverter are amplified by 10 times and then sent through the analog I/O channels to dSpace.

The amplifier is shown in Figure 22.

Figure 22: Inverting operational amplifier.

The output voltage is:

$$V_{out} = \frac{R_f}{R_{in}} V_{in} \quad (61)$$

R_f and R_{in} are chosen to be 10k Ω and 1k Ω .

By amplifier the measured current signal, the phase current error is reduced to 0.11% when the value is 1A and 1.1% when 0.1A.

According to the proposed method, the actual value of the phase current is not needed; instead, we only need the derivative value of the phase current. Thus, we can set a threshold for the input current signal. In another word, when the input is smaller than

1A, the sampled data will not be valid for calculating $\frac{di}{dt}$. In this case, the phase current error would be less than 0.11%.

342GD120-314CTV inverter also integrates the DC link voltage sensor. A normalized, analog voltage signal of the actual DC-link voltage is also available at the connector as: 9V↔900V.

The DC link voltage in this work uses 300V voltage applied at the DC side of the inverter. Similarly, the measured DC link voltage signal can be calculated as 3V and the DC link voltage error can be calculated as:

$$\frac{0.0003}{3} \approx 0.03\% \quad (62)$$

Equation (62) shows that the DC link voltage sampled at the analog I/O channel has sufficient accuracy for further data processing.

Also, unlike the measured current signal, the measured DC link voltage does not need to be amplified.

Line-to-Line Voltage.

The line-to-line voltage is measured using an voltage transducer LV25-800 shown in Figure 23.

LV25-800 is used for the application of AC variable speed drives and servo motor drives, to measure AC and pulsed voltage. It provides a galvanic isolation between the primary circuit and the secondary circuit. Important electrical data of LV25-800 are:

- The input voltage measured ranges from 0V to 1200V.
- Response time to 1080V input is 25μs.

- Supply voltage is 15V.

Figure 23: Inverting operational amplifier circuit board.

LV25-800 uses the Hall effect to measure the current I_s , the measure I_s value is then transformed into voltage signal by passing through a resistor R_m , as shown in Figure 24.

Figure 24: Circuit connection of LV25-800.

I_s is positive when the positive potential of the input voltage is applied on terminal +HT and three terminals at the right side of Figure 24 represent:

Terminal +: supply voltage +15V;

Terminal M: measurement;

Terminal -: supply voltage -15V.

Figure 25 shows the photo of ADC channels on dSpace DS1104 board.

Figure 25: ADC channels on dSpace DS1104 board.

EXPERIMENTAL RESULTS.

To verify the proposed sensorless control technique, an experiment was carried out on a low saliency IPMSM. Figure 26 shows the block diagram of the proposed method.

Figure 26: Block diagram of the proposed method.

Figure 27 and 28 shows the experimental results at the position of 30 electrical degree. Figure 27 shows the waveforms for the first injection. From top to bottom, shows the waveforms of the DC link voltage, AC line voltage and the phase B current. A 200 μ s pulse signal is injected into phase B and phase C. Because dSpace is a real-time system, the figures do not show the initial time for the pulse signal.

Figure 27: The first injection experimental results at 30 electrical degree. (Top: DC link voltage; Middle: AC line voltage; Bottom: phase B current.)

At the beginning of the pulse injection, the upper switch of phase B and the lower switch of phase C are turned on and the 300V DC link voltage is applied between phase B and phase C. The current would flow from phase B to the neutral point N which is connected to phase A in this case, and then to phase C. Due to the inductive

characteristics of the motor, the phase B current would ramp up linearly to the peak value of 9.1A. As a result, AC line voltage would go positive, with amplitude of 152.7V. The derivative value of phase B current is calculated based on the initial current value zero, peak current value and the pulse period which is 100 μ s. Also, the positive value of AC line voltage, together with the derivative value of phase B current is used to calculate the rotor position.

Then, the lower switch of phase B and the upper switch of phase C is turned on and the voltage between phase B and phase C becomes negative. This period lasts 100 μ s, the same time as the previous period. It is used to release the energy stored in phase B and phase C. As a result, phase B current would go back to zero again. After 2 minutes, the experiment is repeated and after several times, an average value for estimated position can be obtained.

Figure 28, from top to bottom, shows the waveform of the DC link voltage, AC line voltage and the phase A current. The process is similar to that for the first injection. A 200 μ s pulse signal is injected into phase B and phase C. The upper switch of phase A and the lower switch of phase B are turned on and the phase A current ramps up linearly to the peak value of 8.85A. As a result, AC line voltage would go positive with 152.6V amplitude. The positive value of AC line voltage and the derivative value of phase A current in this period are used for rotor position estimation.

Similarly, after 100 μ s, the lower switch of phase A and the upper switch of phase B is turned on and the voltage between phase A and phase B becomes negative.

Figure 28: The second injection experimental results at 30 electrical degree. (Top: DC link voltage; Middle: AC line voltage; Bottom: phase A current.)

Figure 29 and 30 show the experimental results at the position of 0 electrical degree.

Figure 29: The first injection experimental results at 0 electrical degree. (Top: DC link voltage; Middle: AC line voltage; Bottom: phase B current.)

Figure 29 shows the waveforms for the first injection. From top to bottom, shows the waveform of the DC link voltage, AC line voltage and the phase B current.

Similarly, a 200 μ s pulse signal is injected into phase B and phase C. The peak value of the phase B current is 9A and the amplitude of the AC line voltage is 149.3V.

Figure 30 shows the waveforms for the second injection. From top to bottom, shows the waveform of the DC link voltage, AC line voltage and the phase B current. Similarly, a 200 μ s pulse signal is injected into phase B and phase C. The peak value of the phase A current is 9A and the amplitude of the AC line voltage is 152.7V.

Figure 30: The second injection experimental results at 0 electrical degree. (Top: DC link voltage; Middle: AC line voltage; Bottom: phase A current.)

Figure 31: Estimated results of the proposed method.

The voltage and current signals shown above are sampled and sent to the 16-bit ADC channels of the dSpace system to solve for the rotor position at every 10 electrical degrees from 0 °to 180 °.

Then, the estimated rotor position is compared with the real rotor position from the encoder attached to the rotor shaft.

Figure 31 shows the estimated rotor position versus the actual rotor position. The maximum error is 5 electrical degrees.

Figure 32: Estimation error.

Figure 32 shows the estimation error versus actual rotor position. The peak error is 5 electrical degrees and the rms and the variance error are calculated in equation (63) and equation (64):

$$\varepsilon_{\text{rms}} = \sqrt{\frac{\sum_{i=1}^{19} \varepsilon_i^2}{19}} \approx 3.40^\circ \quad (63)$$

Where ε_i is the estimation error at the i th sampled point.

$$\sigma = \sqrt{\frac{\sum_{i=1}^{19} (\varepsilon_i - \bar{\varepsilon})^2}{19}} \approx 3.32^\circ \quad (64)$$

Where $\bar{\varepsilon}$ is the mean value of the estimation error.

The estimation error might mainly for two reasons. Firstly, the mathematical model in the stationary α - β is not 100% accurate. Secondly, the estimation accuracy is restricted by the measurement resolution of the equipments. Theoretically, the error due to the quantization error for the phase currents is less than 0.11%, DC link voltage less than 0.03% and line-to-line voltage less than 0.06%, as introduced in the hardsetup section.

CONCLUSION

This work presents a new signal injection method for low saliency IPMSM. The proposed method is based on the mathematical model in stationary α - β frame and overcomes the challenge of the small signal-to-noise ratio of most current sensorless technique proposed for low saliency IPM motor.

Experimental results show that the proposed method can estimate the rotor position with satisfactory error. The maximum error is 5 electrical degrees.

Usually, the motors are started up with open-loop control assuming any random initial position. After that the sensorless techniques developed for non-zero speed are utilized. However, the motor can first start in the wrong direction or rotate back and forth for a while, or a low starting results if a wrong angle is assumed. In applications where high starting torque is required, the knowledge of the initial rotor position is needed.

The proposed algorithm allows for field orientation control at standstill, and therefore, allows a high starting torque to be produced at standstill.

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