High Speed Brushless DC Motor Position Sensor less Control Based on Non Ideal Back EMF

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ABSTRACT

This paper provides atypical idea of position sensor less control of high-speed brushless DC motors using non-ideal back electromotive force (EMF) in a magnetically suspended control moment gyro for space to improve motor reliability. The traditional line-to-line voltage has commutation angle error detected by zero-crossing points detection method. A two-stage commutation error compensation method is suggested to achieve the highly reliable and accurate commutation in the operating speed region of the proposed sensor less process of controlling the motor. The transformative line voltages is used to compensate the commutation angle error with the help of hysteresis comparators and the appropriate low pass filter design in the low and high-speed region. Highly precise commutations are achieved in the high-speed region motor steady state loss reduction. The experimental results reveal that the proposed method can attain an effective compensation in the entire operating speed region.

KEY WORDS: Back electromotive force (EMF) detection, Brushless DC (BLDC) motor, Position sensor less control.

1. INTRODUCTION

In space stations and satellites as a key actuator for orientation control, a magnetically suspended control moment gyro (MSCMG) has the aspect of high precision, large moment, and long durability due to the negligible friction and intensified damping of the high-speed rotor part. The high-speed motor of an MSCMG is normally operating at a high speed constantly to supply the required angular momentum for the rotor system at high speed. For space application, high reliability and low power loss are demanded in the high-speed motor. Brushless DC (BLDC) motors with ironless and slot less stator are introduced (Fang, 2014).

A BLDC motor is usually operated by a three-phase full-bridge inverter employing a two-phase conduction method. Normally, the commutation logic of the transistors in the inverter is provided by the position (hall) sensors. It has the advantage of a simple control algorithm and mature application. However, it has some constraints in the harsh environment especially in space application. The lead wires of the position sensors which are unavoidable will reduce the system reliability. Moreover, the steep placement accuracy of Hall sensors is difficult to achieve, which will increase power loss of the motor in the high-speed steady state region (Fang, 2014). For the past few years, position sensor less control methods have been a popular research topic for the increased reliability. Also, the error compensation of the position detection is a research area in sensor less control techniques nowadays.

The error in the estimation scheme of the position of the will surely decrease the efficiency of the motor notably in the high-speed region. Thus, for the sake of achieving higher reliability and lower steady-state loss of power, a high-accuracy commutation scheme for sensor less high-speed BLDC motor control of an MSCMG is been proposed.

The phase back electromotive force (EMF) detection using zero-crossing points (ZCPs) detection method is widely used in controlling without the position sensor. The ZCPs detection method of phase voltages is attained using the motor neutral voltage. The assumption of a virtual neutral point is adopted in (Cao, 2009) for the motor in which the attainment of the required neutral point voltage is not so easy. The ideal commutation periods in which are the ZCPs of the line-to-line back EMF delayed by 30 electrical degrees to the ZCPs of the phase back EMF voltages. The research takes up coordinate transformation method for compensation of errors in the rotor position detection by controlling the angular coordinates in real time. The method is suitable for the wide speed range. The method using the integration of the un conducting phase back EMF voltage is presented in (Shao, 2006). However, the error accumulation problem in the low-speed region should be taken into account. The sensor less control method which is based on the phased-lock loop and the third-harmonic back EMF is proposed. In this method, the motor neutral point voltage is required. A software based error compensation method is proposed, which may bring in complicated algorithm and need high sampling frequency for a high-speed motor. The research proposes a robust and much accurate sensor less method based on a speed-independent position function. This method also needs highly accurate sampling frequency in the high-speed application. Hysteresis comparator is adopted for ZCPs detection in the three-phase voltage line-to-line based on ideal trapezoidal back EMF. The commutation error is decreased in the whole region of operation of the sensor less control method. However, this method is not effective in the low-speed region as the voltage drop effect on the stator resistance on the commutation detection error is obvious. Also, the motor neutral voltage is needed in the proposed method. A special value for the voltage source of the hysteresis comparator should be obtained.
This paper proposes an improved method of position sensor less control for the high-speed Brushless DC motor based on non-ideal back EMF. Based on the position detection error analysis of the traditional line voltage detection method, a two-stage commutation error compensation method which uses a virtual neutral point without the need of a complicated software algorithm is suggested. The transformative line voltages and hysteresis comparators are adopted to achieve high precision commutation in the medium- and high-speed region. Meanwhile, LPFs and the transformative line voltages are adopted to achieve high reliable commutation in the low-speed region. The parameter design of the proposed method is presented in detail. The effect of the non-ideal back EMF voltage is eliminated by the characteristic parameters measured in the offline mode. The proposed method can directly generate commutation signals replacing Hall signals and achieve high reliable and high accurate commutation without phase shift measures in the whole operating speed region of the back EMF-based sensor less control process. Experimental results show the validity of the proposed method.

**Mathematical Model of BLDC Motors:** The inverter topology of the high-speed motor of an MSCMG is shown in Fig.1. The motor input voltage is regulated by a buck converter containing two energy storing elements, a coil and a capacitor. The output voltage of the buck converter satisfies \( u_d = Du_{\text{in}} \), where \( D \) is the duty ratio the switch \( T_d \). The voltage.

![Figure 1](image.png)

Figure.1. Topology of the buck-converter

Equations of three phases can be represented as

\[
\begin{align*}
    u_A &= R_p i_A + Ld i_A/dt + e_A + u_N \\
    u_B &= R_p i_B + Ld i_B/dt + e_B + u_N \\
    u_C &= R_p i_C + Ld i_C/dt + e_C + u_N
\end{align*}
\]  

(1)

Where \( R_p = R + r \) is satisfied, \( R \) is the stator resistance, \( r \) is the switch on-state resistance, \( L \) is the phase inductance, \( u_A, u_B, \) and \( u_C \) are the three-phase voltages, \( i_A, i_B, \) and \( i_C \) denote the three-phase currents, \( e_A, e_B, \) and \( e_C \) are the back EMF voltages of three phases, and \( u_N \) is the neutral voltage.

In this study, the resistances of the three-phase windings are considered to be constant and equal. Also, the phase inductance of the ironless stator motor is negligible. The ideal commutation instants are the intersections of every two-phase back EMFs. When and only when the motor is commutating with the ideal commutation instants, the maximum electromagnetic torque with the minimum dc-link current, and the lowest commutation torque ripple will be achieved.

**Traditional Line Back EMFS Sensing Scheme:**

**Traditional Line Back EMFs Sensing:** The line-to-line back EMF voltages are contained in the corresponding line-to-line voltages of three phases. Considering that \( x, y, z \) represent the upper switch conducting phase, the lower switch conducting phase, and the floating phase, the three line-to-line voltages can be expressed as;

\[
\begin{align*}
    u_{xy} &= 2R_p i_x + e_{xy} \\
    u_{yz} &= -R_p i_x + e_{yz} \\
    u_{zx} &= -R_p i_x + e_{zx}
\end{align*}
\]  

(2)

![Figure 2](image.png)

Figure.2. Operating Voltage Waveforms

From (1) by neglecting the low-phase inductance value. \( u_{xy} > 0, u_{yz} < 0, \) and \( u_{zx} < 0 \) can be deduced since \( e_{xy} > 0, e_{yz} \leq 0, \) and \( e_{zx} \leq 0 \) satisfy. Fig.3, shows the three-phase back EMF voltages and line-to-line back EMF voltages, respectively. Since the value of stator resistance is small, the ZCPs of line-to-line voltages can be used to determine the commutation instants.

Generally, the LPF is adopted to deal with the commutation ripple and the measurement noise of the terminal voltages. Since the buck converter can provide smooth and controllable motor input voltage and the six switches in
three-phase full bridge are controlled without PWM, the high-frequency ripples in the terminal voltages caused by the traditional PWM control scheme.

**Figure 3. Nonideal back EMF and Hall Signals of the motor**

Are decreased. Thus, the cut-off frequency of the LPF can be chosen as a higher value to avoid large phase delay at the high speed for the high-speed motor of an MSCMG.

The commutation ripple should be taken into account in the design of LPF. Taking the commutation from state 3 to state 4 shown in Fig. 3, for example, the switch T1 is turned OFF after the commutation. The current of phase A will not decrease to zero immediately due to the existence of phase inductance. It will freewheel through the diode D4. Substituting $u_A = 0$, $u_B = u_d$, $u_C = 0$, $e_A = e_B = E_c$, and $e_C = -E$ into (1), $u_N = (u_d - 2E_c + E)/3$ is achieved, where $E$ denotes the flat magnitude of phase back EMF voltage and $E_c$ denotes the magnitude of the crossing point of two phase back EMF voltages. Equation (1) can be written as

$$R_p i_A + L di_A/dt = -u_d + E + E_c = u_{c1}$$

$$R_p i_B + L di_B/dt = 2u_d - E_c - E = u_{b1}$$

$$R_p i_C + L di_C/dt = -u_d - 2E_c - 2E = u_{c1}. \quad (3)$$

Then, the three-phase currents can be calculated as

$$i_A = \frac{u_{a1}}{R_p} + \left( I - \frac{u_{a1}}{R_p} \right) e^{-\frac{R_p}{L} t}$$

$$i_B = \frac{u_{b1}}{R_p} - \frac{u_{c1}}{R_p} - \frac{E_c}{R_p} e^{-\frac{R_p}{L} t}$$

$$i_C = \frac{u_{c1}}{R_p} - \left( \frac{u_{c1}}{R_p} + I \right) e^{-\frac{R_p}{L} t} \quad (4)$$

By solving the differential equations in (3) with the initial value $i_A(t = 0) = I$, $i_B(t = 0) = 0$, and $i_C(t = 0) = -I$. The time intervals of the current of phase A decreasing from $I$ to zero and the current of phase B increasing from zero to $I$ can be calculated as

$$t_a = \frac{3IL}{3IR_p + u_d + E + E_c}$$

$$t_b = \frac{3IL}{2u_d - E_c - E}. \quad (5)$$

### Table 1. System ratings

<table>
<thead>
<tr>
<th>Specifications</th>
<th>Quantity</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of pole pairs</td>
<td>6</td>
</tr>
<tr>
<td>Moment of inertia $J$</td>
<td>0.00955 kg · m²</td>
</tr>
<tr>
<td>Torque constant $k_t$</td>
<td>0.01 N · m/A</td>
</tr>
<tr>
<td>Back EMF constant $k_e$</td>
<td>0.0015 V/r/min</td>
</tr>
<tr>
<td>Terminal resistance $R$</td>
<td>0.15 Ω</td>
</tr>
<tr>
<td>Armature inductance $L$</td>
<td>3.4 µH</td>
</tr>
<tr>
<td>Converter input voltage $u_{in}$</td>
<td>35 V</td>
</tr>
<tr>
<td>Capacitor of the buck converter $C_0$</td>
<td>5 µF</td>
</tr>
<tr>
<td>Inductance of buck converter $L_f$</td>
<td>1 mH</td>
</tr>
</tbody>
</table>

**Figure 4. Different operating currents**
When the motor is operating at a low speed, \( u_d \geq 3IR_p + 2E + 2E_c \) may be achieved. \( t_a \geq t_b \) and \( t_a \leq IL/(2IR_p + E + E_c) \) can be deduced. When the motor is operating at a medium or high speed, \( u_d < 3IR_p + 2E + 2E_c \) is achieved. \( t_a < t_b \) and \( t_b \leq IL/(2IR_p + E + E_c) \) can be calculated since \( u_d \geq 2IR_p + E + E_c \) satisfies.

According to the motor ratings shown in Table I, the time intervals of commutation ripple with different operating currents are depicted in Fig. 4. It can be learnt that the commutation interval decreases with the increasing motor speed. Thus, for the high-speed motor of the MSCMG with low inductance, the commutation interval is less than 10 μs. Appropriate LPF should be chosen to insure high reliability of the ZCPs detection of line-to-line voltages.

**Position Detection Error:**

**Voltage Drop on the Stator Resistance:** From (2), the ZCPs of the line-to-line voltage \( u_{CA} \) shown in Fig. 4 satisfy \( u_{CA} = R_p i + e_{CA} = 0 \), where \( i \) is the dc-link current. Then, \( e_{CA} = -R_p i \) is achieved in the detected commutation instant. Since the ideal commutation instant satisfies \( e_{CA} = 0 \), the detected ZCP of the line-to-line voltage is delayed to the ideal commutation instant.

Using the piecewise linearization of the nonideal phase back EMF voltage, the delayed angle \( \theta_r \) can be calculated as

\[
\theta_r = \begin{cases} 
\frac{\pi}{12} R_p i - R_p i \leq E_3 - E_1 \\
\frac{\pi}{12} R_p i - (E_3 - E_1) - \frac{\pi}{12} R_p i > E_3 - E_1
\end{cases}
\]

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\[
\theta_r = \begin{cases} 
\frac{\pi}{12} R_p i - R_p i \leq E_3 - E_1 \\
\frac{\pi}{12} R_p i - (E_3 - E_1) - \frac{\pi}{12} R_p i > E_3 - E_1
\end{cases}
\]

**Phase Delay of the LPF:** The LPF circuit shown in Fig. 6 is adopted in the detection of the three-terminal voltages. The transfer function of the LPF and the delayed angle caused by the LPF are given as \( U_{af} \).

\[
\frac{U_{af}}{R_i} = \frac{f_t}{f_c} + \frac{R_c}{f_c^2} \]

\[
\theta_f = -\arctan \frac{2\pi R_1 C_1 f_t}{R_c + R_i} = -\arctan 2\pi R_1 f_t
\]

Where \( E_0, E_1, E_2 \) are the magnitudes of back EMF voltage of phase A at the electrical angle of \( \pi/6, -\pi/12, \) and \( \pi/12 \) shown. Fig.5, shows the Gate Pulses caused by voltage drop on stator resistance with the different operating currents.

The influence of phase delay caused by voltage drop on stator resistance is obvious at low speed. In the medium- and high-speed region, the phase delay is small. Also, it should be noted that the delayed angle increases with the operating current.

**Abnormal Conduction of the Freewheeling Diode in the Unexcited Phase:** The research in (Fang, 2014) mentioned that the nonideal commutation instant will cause the abnormal conduction
Of the freewheeling diode in the unexcited phase at high speed. It should be noted that the conduction of the freewheeling diode will affect the terminal voltages, which may introduce commutation instant detection error as well.

We still take the commutation instant shown in Fig. 5, for example. When the detection commutation instant is delayed to the ideal one, the three-phase voltages satisfy \( u_A = e_A + u_N \), \( u_B = 0 \), and \( u_C = u_d \). Substituting the three-phase voltages into (1), the unconducting phase voltage and the current of phase \( C \) can be calculated as

\[
u_A = u_d + 2e_A - e_B - e_C
\]

\[i_{C1} = \frac{u_d - (e_C - e_B)}{2R_p} > 0.
\]

The abnormal conduction of the freewheeling diode in the unexcited phase will happen when \( u_A > u_d \) satisfies. The voltages of the two commutation phases can be expressed as

\[
\begin{align*}
  u_{AN} &= R_p i_A + e_A \\
  u_{CN} &= R_p i_C + e_C.
\end{align*}
\]

Since \( i_A = 0 \) is satisfied when the conduction of the freewheeling diode does not happen and \( i_A < 0 \) is satisfied when the conduction of the freewheeling diode happens. Also, substituting \( u_A = u_d \), \( u_B = 0 \) and \( u_C = u_d \) into (1), the current of phase \( C \) can be calculated as

\[
i_{C2} = \frac{u_d - (e_C - e_B) + (e_A - e_C)}{3R_p}.
\]

Compared (11) with (13), \( i_{C1} - i_{C2} < 0 \) is achieved from (10) and \( u_A > u_d \). Thus, the voltage \( u_{CN} \) will increase when the conduction of the freewheeling diode happens. The line-to-line voltage \( u_{AC} \) is smaller than the normal state, which will cause the ZCP of the line-to-line voltage with abnormal conduction of the freewheeling diode in the unexcited phase delayed to the ZCP of the line-to-line voltage in the normal state.

**System Delay:** The system delay caused by hardware circuit and software algorithm may cause the position detection error. The response time of the amplifier and comparator and the computing time of the sensor less control algorithm should be taken into account. The delayed angle will increase with the motor speed since the delayed time of hardware circuit and software algorithm is constant. Since the software delay is very small, it can be negligible. The measured hardware delay of the system is shown in Table 2. The detected commutation angle error caused by system delay is expressed as

\[
\theta_s = -2\pi T_d.
\]

Where \( T_d \) is the sum of the hardware system delayed time.

The commutation angle error caused by traditional ZCPs detection of line voltages during the whole operating speed region of the high-speed motor of an MSCMG. The sensor less control based on back EMF detection is working in the speed region of a low speed such as 1000 r/min to the rated speed for the high-speed motor of an MSCMG. It can be learnt that the commutation angle error decreases with the increasing motor speed in the low-speed region and increases in the medium- and high-speed region. Also, the delayed angle caused by the small value of the cut-off frequency of the LPF in the high-speed region is large, which inevitably increases the motor steady-state power loss. Thus, a proper compensation method of the delayed commutation angle should be proposed to achieve high-reliable commutation in the operating speed region of the back EMF-based sensor less control process and high-accurate commutation in the high-speed region for the high-speed motor of an MSCMG.

**Design of The Proposed Two-Stage Error Compensation Method:** A commutation instant detection method based on the compensated ZCPs of line voltages is adopted. As shown in Fig. 8, the ZCPs of the line voltages are advanced to the ZCPs of the filtered line voltage \( u_{ABS}, u_{BCSF1}, \) and \( u_{CASF1} \) satisfying \( u_{ABS} = u_{ASF} - u_{BSF1}, u_{BCSF1} = u_{BSF} - u_{CSF1}, u_{CASF1} = u_{CSF} - u_{ASF} \) are advanced to the ZCPs of the filtered line voltages \( u_{ABS}, u_{BCSF}, \) and \( u_{CASF} \) satisfying \( u_{ABS} = u_{ASF} - u_{BSF}, u_{BCSF} = u_{BSF} - u_{CSF}, u_{CASF} = u_{CSF} - u_{ASF} \). The amplified voltages satisfying \( u_{ASF} = mu_{ASF}, u_{BSF1} = mu_{BSF}, \) and \( u_{CASF1} = mu_{CSF} \) are used to generate the compensated commutation instants, where \( m > 1 \) is achieved. The compensation method based on the compensated line voltages is effective when the compensation angle is less than 30 electrical degrees. In order to achieve accurate compensation effect in the rated speed, the cut-off frequency of the LPF should not be chosen too small.

<table>
<thead>
<tr>
<th>Hardware system delay</th>
<th>Specifications</th>
<th>Quantity</th>
</tr>
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<tbody>
<tr>
<td>The detection circuit</td>
<td>The comparator LM139</td>
<td>1.3 ( \mu s )</td>
</tr>
<tr>
<td></td>
<td>The amplifier TL084</td>
<td>4 ( \mu s )</td>
</tr>
<tr>
<td>The drive circuit</td>
<td>The three-phase bridge driver IR2130</td>
<td>3 ( \mu s )</td>
</tr>
<tr>
<td></td>
<td>The power MOSFET IRF540</td>
<td>0.1 ( \mu s )</td>
</tr>
</tbody>
</table>
A two-stage compensation method is proposed to achieve high accurate commutation in the high speed region and high reliable position sensor less commutation scheme in the back EMF-based sensor less control process. In the medium and high-speed region, the transformative line voltages and hysteresis comparators are adopted to achieve high accurate commutation angle compensation effect. Then, the commutation signals $S_{1H}$, $S_{2H}$, and $S_{3H}$ are generated to replace the Hall position signals. Meanwhile, LPFs and transformative line Voltages are adopted to achieve high reliable commutation in the low-speed region of the proposed sensor less control process, where the commutation signals $S_{1L}$, $S_{2L}$, and $S_{3L}$ are generated. The circuit of the proposed system is shown in Fig. 9.

Two-Stage Switching Point of the Proposed Method: In the motor operating speed region, the delayed angle of the detected commutation instant can be expressed as $\theta_d(f_t) = \theta_1(f_t) + \theta_2(f_t) + \theta_3(f_t)$. From (7), (9), and (14), the equation of $\theta'_d(f_t) = 0$ is adopted to calculate the switching point of the proposed two-stage error compensation scheme, where $f_t$ is the motor terminal voltage frequency in the two-stage switching point speed of the proposed error compensation method. The equation can be simplified as

$$\theta'_d(f_t) = \begin{cases} \frac{\pi}{12} R_p \frac{i}{f_t^2} - 2\pi(R_f + T_s), & 2R_p i \leq k_c f_t \\ \frac{\pi}{12} \frac{R_p i}{(k_3 k_c - k_c) f_t^2} - 2\pi(R_f + T_s), & 2R_p i > k_c f_t \end{cases}$$

(15)

By using the first-order Taylor series expansion, which satisfies $\arctan(2\pi R/ff) \approx 2\pi R/ff$, where $k_c = (k_2 - k_1) k_c$, $k_1 = E_v/E_c < 1$, $k_2 = E_v/E_c > 1$, $k_3 = E_v/E_c > k_2$, $k_c = E_v/f$ are achieved. The two-stage switching point of the proposed method can be calculated as
Where \( I_m \) is the maximum motor operating current in the speed region of the proposed sensorless control process.

### Table 3. System parameters

<table>
<thead>
<tr>
<th>Specifications</th>
<th>Quantity</th>
</tr>
</thead>
<tbody>
<tr>
<td>( k_1 )</td>
<td>0.7109</td>
</tr>
<tr>
<td>( k_2 )</td>
<td>1.0937</td>
</tr>
<tr>
<td>( k_3 )</td>
<td>1.1233</td>
</tr>
<tr>
<td>( k_4 )</td>
<td>1.151</td>
</tr>
<tr>
<td>( k_c )</td>
<td>0.01124</td>
</tr>
</tbody>
</table>

Using the piecewise linearization method, the characteristic parameters of the phase back EMF voltage of the discussed high-speed motor shown in Table III can be measured in the offline mode at a reference speed.

### Error Compensation Scheme in the Medium- and High-Speed Region:

The voltage of the virtual neutral point \( u_S \) is given as

\[
    u_S = q(u_{AF} + u_{BF} + u_{CF}) \sqrt{3}
\]

Where \( q \) is the divider resistance ratio of the LPF satisfying \( q = R_4/(R_1 + R_4) = R_9/R_2 + R_3 = R_6/R_3 + R_5 \). From (1), the voltage difference between the virtual neutral point and the actual neutral point can be calculated as

\[
    u_{SN} = u_S - qu_N = q(e_A + e_B + e_C) \sqrt{3}.
\]

From Fig. 8 which shows the compensation scheme, the filtered back EMF voltages of the two commutation phases, \([\pi/12, \pi/6]\) are given as

\[
    \begin{align*}
    e_{AF}(\omega t) &= \frac{12E_1}{\pi}\omega t + E_1 \\
    e_{CF}(\omega t) &= E_2 - \frac{12(E_0 - E_2)}{\pi}\omega t \\
    e_{AF}(\omega t) &= \frac{12(E_c - E_1)}{\pi}\omega t + E_1 \\
    e_{CF}(\omega t) &= E_2 - \frac{12(E_2 - E_c)}{\pi}\omega t \\
    e_{AF}(\omega t) &= \frac{12(E_2 - E_c)}{\pi}(\omega t - \frac{\pi}{12}) + E_c \\
    e_{CF}(\omega t) &= E_c - \frac{12(E_c - E_1)}{\pi}(\omega t - \frac{\pi}{12})
    \end{align*}
\]

\[
    \begin{align*}
    V_{\text{ASF1n1}} - \frac{R_{10} + R_{11}}{R_{11}}u_{\text{BSF1H}} &= \frac{R_{10}}{R_{11}}V_{dc} \\
    V_{\text{ASF1n2}} - \frac{R_{10} + R_{11}}{R_{11}}u_{\text{BSF1H}} &= -\frac{R_{10}}{R_{11}}V_{dc}
    \end{align*}
\]

The hysteresis comparator is adopted to insure high reliable senseless detection in the medium- and high-speed region. The hysteresis comparator will introduce phase delay which should be compensated in the proposed sensorless method. Taking the phase voltages of A and B, for example, the voltage equations of the hysteresis comparator are given as VASF1 and VASF2 satisfying VASF1 > VASF2 are the voltages of phase A when the output voltage of the hysteresis comparator switches from \( -V_{dc} \) to \( V_{dc} \) and from \( V_{dc} \) to \( -V_{dc} \), respectively.

From (22), the hysteresis comparator has the amplifying effect expressed as \( u_{\text{ASF1}} = (1 + n)u_{\text{BSF1H}} \). The sum amplifying ratio \( m_h \) is expressed as

\[
    m_h = m_a(1 + n)
\]
where \( m_s = 1 + R_{ds}/R_{19} = 1 + R_{22}/R_{23} = 1 + R_{22}/R_{23} \) is achieved. Since \( u_{CSF} = qe_{AF} - u_{SN} \) and \( u_{CSF} = qe_{CF} + qR_{ti} - u_{SN} \) are achieved, the equation of
\[
\begin{align*}
uc_{CSF} &= u_{CSF} - (1 + n)u_{AFH} \\
&= qe_{CF} + qR_{ti} - qm_{4}c_{AF} + (m_{th} - 1)u_{SN} \quad (25)
\end{align*}
\]
Can be deduced. The proper parameters \( m_{th} \) and \( u_{th} \) should be chosen to achieve high accurate error compensation. When the delayed angle caused by LPF and system delay satisfies \( \theta_2 + \theta_3 \geq -\pi/12, 1 + 3(R_{ti} + u_{th})/(E_c + E) < m_\theta < 1 + 3(k_2 - k_1 + R_{ti} + u_{th})\}[(2k_1 - k_2 + k_3)e_c] \) is achieved. Using the transformative line voltage, the advanced angle is given as
\[
\theta_a = \frac{\pi}{12} \left[ \frac{(m_\theta - 1)(1 + k_1)}{m_\theta - 1 + (m_\theta + 2)k_2 - (2m_\theta + 1)k_1} - \frac{3R_{ti} + 3m_{4}k_1}{m_\theta - 1 + (m_\theta + 2)k_2 - (2m_\theta + 1)k_1} \right] \quad (26)
\]
By substituting \( u_{th} = u_{th}/q, \theta = \pi/12 - \theta_s \) (18), (20), and (25) into \( u_{CSF}(\theta) = -u_{th}, k_4 = E/E_c \geq 1 \) is achieved. When the delayed angle caused by LPF and system delay satisfies \( \theta_2 + \theta_3 \leq -\pi/12, m_\theta > 1 + 3(k_2 - k_1 + R_{ti} + u_{th})\}[(2k_1 - k_2 + k_3)e_c] \) is achieved. From (19), the advanced angle is given as
\[
\theta_a = \frac{\pi}{12} \left[ \frac{(2m_\theta + 1)k_1 - (m_\theta + 2)k_2 + (m_\theta - 1)k_3}{(2m_\theta + 1)k_1 + (m_\theta + 2)k_2 - (2m_\theta + 1)k_1} - \frac{3R_{ti} + 3m_{4}k_1}{(2m_\theta + 1)k_1 + (m_\theta + 2)k_2 - (2m_\theta + 1)k_1} \right] \quad (27)
\]
Thus, when \( m_\theta > 1 + 3(R_{ti} + u_{th})/(E_c + E) \) is achieved, the delayed angle caused by the voltage drop on stator resistance can be compensated completely. The compensation angle is expressed as
\[
\theta_a = \theta_2 + \theta_3. \quad (28)
\]
Using the detected angle error and the advanced angle at the rated speed to calculate the parameter of the compensation scheme, the amplifying ratio \( m_\theta \) and hysteresis threshold \( u_{th} \) satisfy.
\[
\begin{align*}
\theta_2(f_m) + \theta_3(f_m) &= -\theta_a(f_m) \quad (29) \\
\theta_2(f_r) + \theta_3(f_r) &= -\theta_a(f_r) \quad (30)
\end{align*}
\]
Where \( f_m \) is the motor terminal voltage frequency in the midpoint of the medium- and high-speed region, satisfying \( f_m = (f_i - f_3)/2, f_i \) is the motor terminal voltage frequency, satisfying \( f_i = 3(f_h - f_3)/4, f_h \) is the motor terminal voltage frequency in the rated speed. Using the first-order Taylor series expansion which satisfies \( \arctan2\pi R/f_h \approx 2\pi R/f_h \), (29) and (30) can be simplified, and solved as
\[
\begin{align*}
m_\theta &= 1 + \frac{2(f_2 + f_3(R_i + T_s)(k_2 - k_1))}{(1 + k_4)f_r - 24(f_2^2 + f_3^2)(R_i + T_s)k_{h3}} \\
u_{th} &= \frac{24(1 + k_4)(R_i + T_s)k_{h3}f_m f_2^2 g}{(1 + k_4)f_r - 24(f_2^2 + f_3^2)(R_i + T_s)k_{h3}} - qR_p I_s \quad (31)
\end{align*}
\]
From (9), (14), and (26) when \( \theta_2(f_r) + \theta_3(f_r) \geq -\pi/12 \) satisfies, where \( k_{h3} = 1 - 2k_1 + k_3 \) satisfies and \( I_s \) is the motor current in the steady state. Also, from (27), when \( \theta_2(f_m) + \theta_3(f_m) < -\pi/12 \) satisfies, the equations can be solved as
\[
\begin{align*}
m_\theta &= 1 + \frac{2(f_2 + f_3(R_i + T_s)(k_3 - k_1)) + 3(f_3 - k_1)f_r}{k_{h3} f_3 - 24(f_2^2 + f_3^2)(R_i + T_s)k_{h3}} \\
u_{th} &= \frac{8k_{h3}(R_i + T_s)k_{h3}f_m f_2^2 g}{k_{h3} f_3 - 24(f_2^2 + f_3^2)(R_i + T_s)k_{h3}} - qR_p I_s \quad (32)
\end{align*}
\]
Where \( k_{h2} = 4k_1 - 2k_2 + k_3 + k_4, k_{h3} = 2k_1 - k_2 + k_3, \) and \( k_{h4} = 3k_1k_2 + 10k_1k_3 + k_2k_4 + 3k_3k_4 \) are achieved. When \( \theta_2(f_m) + \theta_3(f_m) \geq -\pi/12 \) and \( \theta_2(f_r) + \theta_3(f_r) < -\pi/12 \) satisfy, the equations can be solved as
\[
\begin{align*}
m_\theta &= 1 + \frac{2(R_i + T_s)(k_3 - k_1)f_r^2 + (k_3 - k_1)f_2^2}{k_{h3} f_3 - 24(k_3f_2^2 + k_3 f_2 k_{h3})(R_i + T_s)} \\
u_{th} &= \frac{576(R_i + T_s)(k_3 - k_1)f_r^2 f_2^2 k_{h3} q}{k_{h3} f_3 - 24(k_3f_2^2 + k_3 f_2 k_{h3})(R_i + T_s)} \quad (33)
\end{align*}
\]
Where \( k_{h2} = 2k_1 + 2k_2 - k_3 - k_1 k_3 - k_2 k_4 s, k_{h3} = 2k_1 + k_2 k_3 + k_3 k_4, k_{h4} = 3k_1k_2 + 10k_1k_3 + k_2k_4 + 3k_3k_4 \) satisfies. Similarly, the compensation parameters in the medium- and high-speed region can be calculated using the piecewise linearization method to measure the non-ideal phase back EMF voltage in the offline mode. The compensation effect with different operating currents is shown in Fig.11. \( R_f = 0.00003 \) is selected to insure acceptable phase delay in the rated speed. The switching point speed is calculated
as 4000 r/min, where \( f_z = 400 \) Hz is achieved. The amplifying ratio is calculated as \( m_h = 2.28 \). The hysteresis threshold is calculated as \( u_{th1} = 3.3 \), \( u_{th2} = 3.1 \), and \( u_{th3} = 2.9 \), respectively, when the motor operating current is 1, 2, and 3 A. From Fig. 11, the compensation effect in the low-speed region is imperfect.

**Fig. 11. Gate Drive Pulse**

**Compensation Effect in the Whole Back EMF-Based Sensor less Control Process:** The cut-off frequency of the LPF in the medium and high speed region is chosen by considering both the commutation interval of the motor and the delayed angle caused by the filter. If the cut-off frequency is selected too high, the commutation ripple of the motor voltage may not be filtered. Moreover, if the cut-off frequency is too small, the delayed angle caused by the LPF will increase in the high-speed region, which may decrease the compensation effect of the proposed method. According to the operating speed region of the high-speed motor of the MSCMG, \( R_f = 0.00003 \) and \( R_{dC4} = 0.00007 \) are selected. The system delay is calculated as 10 \( \mu \)s. Normally, the operating current of the high-speed motor is smaller than 3 A. \( 1_m = 3 \) A and \( 1_r = 2 \) A is achieved. It should be noted that the back EMF-based sensor less control is ineffective when the motor speed is zero or very low. When the motor speed is higher than 800 r/min, the ZCPs detection of the relevant voltages can be achieved reliably for the high-speed motor of the MSCMG. In practice, the speed region of the proposed sensor less control process is chosen to be from 1000 r/min to 10 000 r/min. The delayed and compensated angles adopting the proposed compensation scheme with the operating current of 1, 2 & 3 A.

**Figure.12. Motor Outputs**

It should be noted that the abnormal conduction of the freewheeling diode in the unexcited phase can be eliminated when the phase delay caused by voltage drop on stator resistance, hysteresis comparator, LPF and system delay is compensated. Thus, the phase delay caused by the abnormal conduction of the freewheeling diode in the unexcited phase is not taken into account in the design of the proposed compensation scheme.

In the proposed error compensation method, rotor speed information is required to determine the commutation compensation scheme in different speed regions. The rotor speed can be derived from the compensated sensor less detection signals which can replace the Hall position signals without speed sensors in the speed region of the proposed back EMF based sensor less control process.

**Start-Up Technique:** The design of the start-up process in the sensor less drive system is necessary. Normally, the high-speed motor in an MSCMG is operating at a constant speed to supply constant angular momentum for the rotor system. The rotor is designed to have large inertial mass. However, since the torque constant \( k_t \) and the rated current of the motor for space application are small, the speeding up process of the high-speed rotor is much longer than the common system.

The high-speed motor of the MSCMG is driven from the still by the traditional starting method “align and go”. The open loop subsection approximation linearized \( V/f \) starting method is adopted to insure a reliable start-up process. The whole start-up process contains three steps. The first step is aligning the rotor to a known position by exciting the certain two phases for a pre-set time. Then the open-loop accelerating scheme is applied to start the rotor from standstill. When the stable commutation instant detection is available, the motor will switch.

**Figure.13. Simulated voltage and current waves at the speed of 1000 r/min**
To sensor less control mode by the proposed ZCPs detection of line-to-line voltages. In practice, the start-up scheme can rotate the high-speed motor of the MSCMG from standstill to 1000 r/min effectively and reliably.

3. SIMULATED AND EXPERIMENTAL RESULTS

Sufficient experiments have been performed on the setup shown in Fig.9. The ratings of the BLDC motor under study are listed in Tables.1 and 3. A BLDC motor system with a buck converter is established to run the simulation in MATLAB/Simulink. In the proposed error compensation method, the calculated speed of the two-stage switching point is 4000 r/min. The parameters of the proposed compensation method are chosen as $m_l = 1.82$ in the low-speed region and $m_l = 2.28$, $u_{thl} = 3.1$ in the medium- and high-speed region.

Fig.10, show the simulated result of the line voltages and motor current waves with the traditional ZCPs detection of line-to-line voltage method and the proposed compensation method at the speed of 5000 r/min. The operating current of the motor is 1.5 A. $H_l$ and $S_l$ denote the signals of the reference Hall position sensors and traditional

![Figure 14. Simulated voltage and current waves at the speed of 5000 r/min](image1)

![Figure 15. Experimental voltage and current waves at the speed of 10000 r/min](image2)

The phenomenon of the abnormal conduction of the freewheeling diode in the unexcited phase is obvious. It will generate negative electromagnetic torque, which may increase the motor power consumption. The abnormal conduction of the freewheeling diode is eliminated using the proposed sensor less control method. Also, the motor steady-state power loss is decreased by 16% and 20% at the speed of 5000 and 10000 r/min, respectively. The simulated and experimental results show that the proposed sensor less control method can achieve effective commutation angle compensation in the whole operating speed region using the ZCPs detection method. Moreover, the compensated commutation angle error is restrained within 3° under the operating current of 1.5 A in the medium- and high-speed region of 5000 r/min to 10 000 r/min in order to achieve low-steady-state power loss.

4. CONCLUSION

In this paper, a two-stage position sensor less control method based on ZCPs detection of the transformative line-to-line voltages is proposed for the high-speed BLDC motor with low inductance and non-ideal back EMF. The compensated commutation signals are generated using virtual neutral point and the non-ideal back EMF voltage without phase shift measures. The proposed sensor less control method can achieve reliable commutation in the operating speed region of the back EMF based sensor less control process. Also, high-accurate commutation is achieved to decrease the motor steady-state power loss for the high-speed motor of the MSCMG in the high-speed region. Simulation and experimental results show the validity and effectiveness of the proposed method.

REFERENCES


