Elimination of Mutual Flux Effect on Rotor Position Estimation of Switched Reluctance Motor Drives

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Abstract—An approach to eliminate mutual flux effect on rotor position estimation of switched reluctance motor drives at rotating shaft conditions without a prior knowledge of mutual flux is proposed in this paper. Neglecting the magnetic saturation, the operation of conventional self-inductance estimation using phase current slope difference method can be classified into three modes: Mode I, II, and III. At positive-current-slope and negative-current-slope sampling point of one phase, the sign of current slope of the other phase changes in Mode I and II, but does not change in Mode III. Theoretically, based on characteristics of a 2.3 kW, 6000 rpm, three-phase 12/8 SRM, mutual flux introduces a maximum $\pm 7\%$ self-inductance estimation error in Mode I and II, while, in Mode III, mutual flux effect does not exist. Therefore, in order to ensure that self-inductance estimation is working in Mode III exclusively, two methods are proposed: variable-hysteresis-band current control for the incoming phase and variable-sampling self-inductance estimation for the outgoing phase. Compared with the conventional method which neglects mutual flux effect, the proposed position estimation method demonstrates an improvement in position estimation accuracy by 2°. The simulations and experiments with the studied motor validate the effectiveness of the proposed method.

Index Terms—Mutual flux, phase current slope difference, rotor position estimation, switched reluctance motor (SRM) drives.

I. INTRODUCTION

SWITCHED reluctance motor (SRM) is a cost effective solution for the automotive industry due to absence of rotor windings and permanent magnet on the rotor [1]–[9]. In general, the encoder or resolver is installed to obtain the rotor position and speed for the torque or speed control of SRM. This increases the cost and volume of the motor drive, and reduces the reliability. Position sensorless control of SRM is widely studied in the literature [10]–[29]. Magnetic characteristics of the SRM including the flux, self-inductance, and back electromagnetic force (EMF) are rotor position dependent, and therefore, these parameters can be estimated to obtain the rotor position.

Pulse injection method [16]–[20] is one of the rotor position sensorless methods for low speed operation. High-frequency signal is injected to the inactive phase to obtain inductance, which is later converted to rotor position. In [16], the working sector of the SRM is selected by comparing amplitude of current response of inactive phases with two predefined thresholds. This method is simple but it is not promising for instantaneous torque control, where real-time rotor position is required. In [17], voltage pulse is injected to determine the speed range of SRM and then the corresponding dynamic model is defined for rotor position estimation. In [18], a rotor position estimation scheme based on phase inductance vector is proposed. The pulse is injected to get the inductance of inactive phases and full cycle inductance is obtained. This method has advantages of no need of a priori knowledge on magnetic characteristics of the machine. Some pulse injection methods with an additional circuit [20] are also reported and this may increase the cost and implementation complexity. In summary, voltage injection methods often suffer from either additional power losses or low speed constraint. Computation intensive methods such as observer based estimation, and neural network are also presented in [21]–[26]. Although they show robustness or model independence, they are complicated and not promising solutions for practical applications.

Passive rotor position estimation based on measurement of terminal voltage and phase current of active phases are gaining interest due to absence of additional hardware or power losses. One of these methods is to estimate the flux linkage of SRM. A simple approach to obtain the flux linkage is to use the integration of the terminal voltage subtracted by the voltage across the ohmic resistance. This method shows poor accuracy at low speed when back EMF is small. Also, the accuracy is deteriorated by variation of the ohmic resistance and accumulation error due to integration. Several methods are proposed to improve the accuracy of the estimation. In [27], the flux linkage variation instead of flux linkages is used to estimate the rotor position and error analysis of the rotor position estimation is provided. In [28], flux linkage is estimated based on the amplitude of the first-order switching harmonics of the phase voltage and current. The influence of resistance variation and low back EMF for rotor position estimation is suppressed, and therefore, both accuracy and applicable speed range are improved. Self-inductance-based rotor position estimation [18], [29], [30] is an alternative approach of passive position sensorless techniques. By neglecting variation of the speed, back EMF and ohmic resistance in a switching period, the self-inductance is estimated by measuring the phase current slope difference [18]. In [30], incremental inductance is estimated by using phase current slope difference and therefore the application of this method is extended to magnetic saturated region. Compared to flux linkage methods, the influence of the variation of resistance is eliminated and it is capable of operating at low speeds. However, as the speed increases, the overlapping region of the active phases becomes significant and mutual flux...
cannot be neglected anymore. The accuracy of both inductance-based and flux linkage-based rotor position estimation methods are decreased at higher speed due to mutual flux between active phases. In addition, torque sharing function (TSF) [31], [32] is widely used in instantaneous torque control to reduce commutation torque ripples. When TSF is applied, overlapping areas of incoming and outgoing phases are significant even at low speed. Therefore, mutual flux has to be considered to achieve accurate estimation of rotor position over a wide speed range.

Although the influence of the mutual flux on modeling of the SRM is investigated in [33]–[35], publications about the effect of the mutual flux on rotor position estimation are still limited. In [36], mutual flux of SRM with even and odd number of phases is studied and the effect of the mutual flux on position estimation is verified by both simulation and experiments. By calibrating the flux linkage with the measured mutual flux by experiments, the accuracy of position estimation is increased by 3%. In [37], mutual flux linkage obtained by finite element analysis (FEA) is applied to compensate the observed flux linkage. Accuracy of the rotor position estimation is proved to be increased. The mutual flux needs to be either measured or simulated in advance in the two methods proposed in [36] and [37]. However, this process is time consuming. Due to the measurement noise or manufacturing imperfections, mutual flux may not be precise. Meanwhile, flux-linkage based rotor position estimation method is still not desirable at low speed.

In this paper, two methods to eliminate mutual flux effect on rotor position estimation are presented, which do not require a priori knowledge of mutual flux linkage profiles of SRM. In order to investigate the mutual flux effect on rotor position estimation, a dynamic model of SRM incorporating mutual flux is obtained. Then, the self-inductance estimation error due to mutual flux is theoretically derived by using phase current slope difference method. Three operational modes are defined during self-inductance estimation. In Modes I and II, at the positive-current-slope and negative-current-slope sampling point of the outgoing phase, the sign of the current slope of the other phase changes. In Mode III, the sign of the current slope of the other phase does not change. Based on magnetic characteristics of the studied SRM, in Modes I and II, the mutual flux introduces a maximum ±7% self-inductance estimation error. However, in Mode III, the impact of mutual flux on estimation of self-inductances does not exist. Therefore, variable-hysteresis band current controller and variable-sampling self-inductance estimation methods are proposed, so that the self-inductance estimation operates in Mode III exclusively and, hence, a priori knowledge of mutual flux becomes unnecessary. By applying variable-hysteresis-band current controller method in the incoming-phase self-inductance estimation, the switching state of the outgoing phase is kept unchanged and incoming-phase self-inductance estimation is forced to work in Mode III. The hysteresis-band of the outgoing phase is adjusted accordingly. When applied in the outgoing phase self-inductance estimation, variable-hysteresis-band current controller introduces undesirable higher ripples. For this reason, variable-sampling self-inductance estimation is proposed for the outgoing-phase self-inductance estimation. During the outgoing-phase self-inductance estimation, the negative-current-slope sampling point is adjusted to ensure that the current slope of incoming-phase has the same sign at the positive and negative current-slope sampling point of the outgoing phase. Finally, phase self-inductance estimation in the active region for each phase is introduced. Simulation and experimental results are provided to verify the performance of the proposed rotor position estimation scheme at rotating shaft conditions.

II. DYNAMIC MODEL OF SRM CONSIDERING MUTUAL FLUX

A. Circuit Modeling of SRM With Mutual Flux

In a three-phase SRM, no more than two phases are conducted simultaneously. During commutation, incoming and outgoing phases are denoted as $k$th and $(k−1)$th phases, respectively. Phase voltage equations are derived as (1)

$$
\begin{align*}
\lambda_k &= R_l i_k + \frac{\partial \lambda_k}{\partial t} \\
\lambda_{k-1} &= R_l i_{k-1} + \frac{\partial \lambda_{k-1}}{\partial t}
\end{align*}
$$

where $v_k$, $i_k$, and $\lambda_k$ are the phase voltage, current, and flux linkage of $k$th phase, respectively; $v_{k-1}$, $i_{k-1}$, and $\lambda_{k-1}$ are the phase voltage, current, and flux linkage of $(k−1)$th phase, respectively.

As the speed of SRM increases, overlapping areas of the two phases are increased significantly due to higher back EMF. This causes the effect of mutual flux to increase and it cannot be neglected anymore. When mutual flux is considered, flux linkage for incoming and outgoing phases can be expressed as (2)

$$
\begin{align*}
\lambda_k &= \lambda_{k,k} + \lambda_{k−1,k} \\
\lambda_{k−1} &= \lambda_{k−1,k} + \lambda_{k−1,k−1}
\end{align*}
$$

where $\lambda_{k,k}$ and $\lambda_{k−1,k}$ are the self-flux linkages of $k$th and $(k−1)$th phase; $\lambda_{k−1,k}$ and $\lambda_{k−1,k−1}$ are mutual flux linkages.

Neglecting the magnetic saturation, the flux linkage is a linear function of the inductance and, hence, (2) can be reorganized as (3)

$$
\begin{bmatrix}
\lambda_k \\
\lambda_{k−1}
\end{bmatrix} =
\begin{bmatrix}
L_{k,k} & M_{k−1,k} \\
M_{k−1,k} & L_{k−1,k−1}
\end{bmatrix}
\begin{bmatrix}
i_k \\
i_{k−1}
\end{bmatrix}
$$

where $L_{k,k}$ and $L_{k−1,k−1}$ are the self-inductances of the $k$th and $(k−1)$th phase; $M_{k,k−1}$ and $M_{k−1,k}$ are the mutual inductances.

The mutual inductance between two conducted phases meets (4)

$$
M_{k,k−1} = M_{k−1,k}.
$$
Substituting (3) for (1) and (2), the phase voltage equations are derived as (5)

\[
\begin{bmatrix}
    v_k \\
    v_{k-1}
\end{bmatrix}
= \begin{bmatrix}
    R & M_{k,k-1} & L_{k,k-1} \\
    M_{k-1,k} & R & M_{k-1,k-1}
\end{bmatrix}
\begin{bmatrix}
    i_k \\
    i_{k-1}
\end{bmatrix}
+
\begin{bmatrix}
    \frac{\partial L_{k,k}}{\partial \theta} & \frac{\partial M_{k,k-1}}{\partial \theta} & \frac{\partial L_{k,k-1}}{\partial \theta} \\
    \frac{\partial M_{k-1,k}}{\partial \theta} & \frac{\partial M_{k-1,k-1}}{\partial \theta} & \frac{\partial L_{k-1,k-1}}{\partial \theta}
\end{bmatrix}
\begin{bmatrix}
    i_k \\
    i_{k-1}
\end{bmatrix}
\frac{di_k}{dt} + \frac{di_{k-1}}{dt}
\]

where \( \theta \) and \( \omega_m \) are rotor position and angular speed of SRM.

Electromagnetic torque of the \( k \)th phase is represented as (6) neglecting magnetic saturation

\[
T_k(\theta, i) = \frac{1}{2} \frac{\partial L(\theta, i_k)}{\partial \theta} i_k^2
\]

where \( T_k \) is the torque produced by the \( k \)th phase, and \( i_k \) is the \( k \)th phase current.

For an \( n \)-phase SRM, total electromagnetic torque \( T \) is represented as (7)

\[
T = \sum_{k=1}^{n} T_k.
\]

### B. Analysis of Mutual Flux of SRM

The FEA of the studied SRM is conducted in JMAG software and the nonlinear inductance profile and torque profile of studied SRM is shown in Fig. 1(a) and (b), respectively. In three-phase SRM, two phases are excited during commutation. The magnetic flux density distribution of 12/8 SRM during one phase and two-phase excitation are shown in Fig. 2(a) and (b), respectively. Compared with one-phase excitation mode, the two-phase excitation works at short-flux path and the flux linkage of an individual phase includes both self and mutual flux linkage.

Due to alternate polarities of windings of a three-phase motor, mutual flux is always additive and symmetric among individual phases. For the same current on adjacent phases, the mutual inductance profiles obtained from FEA are shown in Fig. 3(a). \( M_{A,B} \) is the mutual inductance between phases A and B. The maximum value of mutual inductance is around 2% of the self-inductance at the same current level. The spatial relationship between self-inductance and mutual inductance of 12/8 SRM is also shown in Fig. 3. The mutual inductance profile \( M_{A,B} \) is shifted by around 7.5° compared with the self-inductance of phase A, \( L_A \).
III. ERROR ANALYSIS OF SELF-INDUCTANCE ESTIMATION DUE TO MUTUAL FLUX

A. Self-Inductance Estimation Without Considering the Mutual Flux

Hysteresis controller is applied for phase current control, as shown in Fig. 4. Upper and lower current references of the kth phase are denoted as $i_{k,\text{up}}$ and $i_{k,\text{low}}$, respectively. So, the hysteresis band is represented as (8)

$$\Delta i_k = i_{k,\text{up}} - i_{k,\text{low}}.$$  \hfill (8)

When the switches $T_1$ and $T_2$ are turned ON as shown as Fig. 5(a), dc-link voltage is applied and the phase current slope is positive. When the switches $T_1$ and $T_2$ are turned OFF, shown as Fig. 5(b), the phase current slope is negative.

The voltage equations neglecting magnetic saturation are then derived as (9) and (10) when switches are ON and OFF, respectively

$$U_{dc} = R i_k + L_{k,k} \frac{d i_k(t_{k,\text{on}})}{dt} + \frac{\partial L_{k,k}}{\partial \theta} i_k \omega_m$$ \hfill (9)

$$-U_{dc} = R i_k + L_{k,k} \frac{d i_k(t_{k,\text{off}})}{dt} + \frac{\partial L_{k,k}}{\partial \theta} i_k \omega_m$$ \hfill (10)

where $t_{k,\text{on}}$ and $t_{k,\text{off}}$ are time instant when the kth phase switching states are ON and OFF during a switching period, respectively; $di_k(t_{k,\text{on}})/dt$ and $di_k(t_{k,\text{off}})/dt$ are the slope of kth phase current at $t_{k,\text{on}}$ and $t_{k,\text{off}}$, respectively; and $U_{dc}$ is the dc-link voltage.

The switching period is short enough and, therefore, variation of the mechanical speed, inductance, back EMF and resistance are neglected. The self-inductance can be derived as (11) by combining (9) and (10). For a given dc-link voltage, unsaturated self-inductance can be estimated by using the phase current slope difference between ON and OFF states

$$L_{k,k} = \frac{2U_{dc}}{di_k(t_{k,\text{on}})/dt - di_k(t_{k,\text{off}})/dt}$$ \hfill (11)

where $L_{k,k}$ is estimated kth phase self-inductance without considering the mutual flux.

B. Analysis of Self-Inductance Estimation Error Due to Mutual Flux

In the previous section, self-inductance estimation is derived by neglecting the mutual flux between two adjacent phases. When overlapping areas of two phases can be neglected, self-inductance estimation without considering the mutual flux is accurate. As the speed increases, the overlapping region becomes significant and the mutual inductance cannot be neglected. Considering the mutual inductance, the kth phase voltage equation is derived as (12) and (13) when kth phase switches are ON state and OFF state, respectively

$$U_{dc} = R i_k + L_{k,k} \frac{d i_k(t_{k,\text{on}})}{dt} + \frac{\partial L_{k,k}}{\partial \theta} i_k \omega_m + M_{k,k-1} \frac{d i_{k-1}(t_{k-1,\text{on}})}{dt} + \frac{\partial M_{k,k-1}}{\partial \theta} i_{k-1} \omega_m$$ \hfill (12)
where \( \frac{di_k(t_k,\omega_{m})}{dt} \) and \( \frac{dL_k}{dt} \) are the slopes of the kth phase current at \( t_{k,\text{on}} \) and \( t_{k,\text{off}} \), respectively.

Similarly, the variation of the mechanical speed, inductance, back EMF and resistance is neglected. Subtracting (12) by (13), \( k \)th phase self-inductance estimation considering the mutual inductance can be obtained as (14)

\[
L_{k,k,\omega_m} = \frac{2U_{dc} - M_{k,k-1} \left( \frac{di_{k-1}(t_{k,\omega_{m}})}{dt} - \frac{di_{k}(t_{k,\omega_{m}})}{dt} \right)}{\frac{dL_k}{dt} - \frac{dL_{k,k-1}}{dt}}
\]

(14)

where \( L_{k,k,\omega_m} \) is the estimated kth phase self-inductance considering mutual flux.

The error of self-inductance estimation due to mutual flux is derived as (15)

\[
\text{err}_{k} = \frac{L_{k,k,\omega_m} - L_{k,k}}{L_{k,k,\omega_m}} = \frac{-M_{k,k-1} \left( \frac{di_{k-1}(t_{k,\omega_{m}})}{dt} - \frac{di_{k}(t_{k,\omega_{m}})}{dt} \right)}{2U_{dc} - M_{k,k-1} \left( \frac{di_{k-1}(t_{k,\omega_{m}})}{dt} - \frac{di_{k}(t_{k,\omega_{m}})}{dt} \right)}
\]

(15)

where \( \text{err}_{k} \) is kth phase self-inductance estimation error due to the mutual flux from \((k-1)\)th phase.

In order to analyze the self-inductance estimation error due to mutual flux, three modes are defined during kth phase self-inductance estimation as shown in Fig. 6: Mode I, II and III. Since the self-inductance of the incoming phase \((k)\) phase is much lower, \( k \)th phase current slope is much higher than \((k-1)\)th phase. Upper and lower current references of \( k \)th phase are denoted as \( i_{k,\text{up}} \) and \( i_{k,\text{low}} \), and upper and lower current references of \((k-1)\)th phase are denoted as \( i_{k-1,\text{up}} \) and \( i_{k-1,\text{low}} \). The positive-current-slope and negative-current-slope of kth phase is sampled at \( t_{k,\omega_{m}(1)} \) and \( t_{k,\omega_{off}(1)} \) in Mode III, \( t_{k,\omega_{m}(1)} \) and \( t_{k,\omega_{off}(1)} \) in Mode II, and \( t_{k,\omega_{m}(1)} \) and \( t_{k,\omega_{off}(1)} \) in Mode I, respectively. The self-inductance estimation error due to mutual flux in Mode I, II and III is derived below. In Mode I and Mode II, the mutual flux leads to self-inductance estimation error, while in Mode III, the mutual flux effect is negligible.

1) Self-Inductance Estimation Error in Mode I: In Mode I, at positive-current-slope sampling point \( t_{k,\omega_{m}(1)} \) and negative-current-slope sampling point \( t_{k,\omega_{off}(1)} \) of kth phase, \((k-1)\)th phase current slope is positive and negative, respectively. Considering the mutual flux from \((k)\) phase, the \((k-1)\)th phase voltage equation can be derived as (16) and (17) at \( t_{k,\omega_{m}(1)} \) and \( t_{k,\omega_{off}(1)} \)

\[
U_{dc} = R_i k - 1 + L_{k,k-1} \frac{di_{k-1}(t_{k,\omega_{m}(1)})}{dt} + \frac{L_{k,k-1}}{\partial \theta} i_{k-1,\omega_{m}} + M_{k,k-1} \frac{di_k(t_{k,\omega_{m}(1)})}{dt}
\]

(16)

\[
-U_{dc} = R_if k - 1 + L_{k,k-1} \frac{di_{k-1}(t_{k,\omega_{off}(1)})}{dt} + \frac{L_{k,k-1}}{\partial \theta} i_{k-1,\omega_{m}} + M_{k,k-1} \frac{di_k(t_{k,\omega_{off}(1)})}{dt}
\]

(17)

Subtracting (17) from (16), (18) can be derived

\[
L_{k,k-1} \left( \frac{di_{k-1}(t_{k,\omega_{m}(1)})}{dt} - \frac{di_{k-1}(t_{k,\omega_{off}(1)})}{dt} \right) + M_{k,k-1} \left( \frac{di_k(t_{k,\omega_{m}(1)})}{dt} - \frac{di_k(t_{k,\omega_{off}(1)})}{dt} \right) = 2U_{dc}.
\]

(18)

Similarly, subtracting (12) by (13), (19) can be derived for \( k \)th phase

\[
L_{k,k} \left( \frac{di_k(t_{k,\omega_{m}(1)})}{dt} - \frac{di_k(t_{k,\omega_{off}(1)})}{dt} \right) + M_{k,k-1} \left( \frac{di_{k-1}(t_{k,\omega_{m}(1)})}{dt} - \frac{di_{k-1}(t_{k,\omega_{off}(1)})}{dt} \right) = 2U_{dc}.
\]

(19)

Subtracting (18) from (19), (20) can be derived

\[
\frac{di_k(t_{k,\omega_{m}(1)})}{dt} - \frac{di_k(t_{k,\omega_{off}(1)})}{dt} = \frac{L_{k,k-1} - M_{k,k-1}}{L_{k,k}} \left( \frac{di_{k-1}(t_{k,\omega_{m}(1)})}{dt} - \frac{di_{k-1}(t_{k,\omega_{off}(1)})}{dt} \right).
\]

(20)

Substituting (20) for (18), the \((k-1)\)th phase current slope difference considering mutual flux from \( k \)th phase in Mode I is derived as (21)

\[
\frac{di_{k-1}(t_{k,\omega_{m}(1)})}{dt} - \frac{di_{k-1}(t_{k,\omega_{off}(1)})}{dt} = \frac{2U_{dc}}{L_{k,k-1} + M_{k,k-1} \frac{L_{k,k-1}}{L_{k,k} - M_{k,k-1}}}.
\]

(21)

Substituting (21) for (15), the error of \( k \)th phase self-inductance estimation due to mutual flux from \((k-1)\)th phase in Mode I is calculated as (22)

\[
\text{err}_{k(1)} = \frac{-M_{k,k-1}}{L_{k,k-1} + M_{k,k-1} \frac{L_{k,k-1}}{L_{k,k} - M_{k,k-1}}} - M_{k,k-1}
\]

(22)
where \( \text{err}_{\text{k(I)}} \) is \( k \)th phase self-inductance estimation error due to mutual flux from \((k-1)\)th phase in Mode I.

2) 2) Self-Inductance Estimation Error in Mode II: In Mode II, at positive-current-slope sampling point \( t_{k, \text{on}(II)} \) and negative-current-slope sampling point \( t_{k, \text{off}(II)} \) of \( k \)th phase, \((k-1)\)th phase current slope is negative and positive, respectively. Similarly, in Mode II, (23) is derived for the \((k-1)\)th phase

\[
L_{k-1, k-1} \left( \frac{d i_{k-1}(t_{k, \text{on}(II)})}{d t} - \frac{d i_{k-1}(t_{k, \text{off}(II)})}{d t} \right) + M_{k, k-1} \left( \frac{d i_{k}(t_{k, \text{on}(II)})}{d t} - \frac{d i_{k}(t_{k, \text{off}(II)})}{d t} \right) = 2U_{dc}. \tag{23}
\]

By using the same approach as Mode I, the error of \( k \)th phase self-inductance estimation due to mutual flux from \((k-1)\)th phase in Mode II is calculated as (24)

\[
\text{err}_{\text{k(II)}} = \frac{M_{k, k-1}}{L_{k-1, k-1} + M_{k, k-1} \frac{L_{1, k-1} + M_{k, k-1}}{L_{1, k-1} - M_{k, k-1}} + M_{k, k-1}} \tag{24}
\]

where \( \text{err}_{\text{k(II)}} \) is \( k \)th phase self-inductance estimation error due to mutual flux from \((k-1)\)th phase in Mode II.

3) 3) Self-Inductance Estimation Error in Mode III: In Mode III, at positive-current-slope sampling point \( t_{k, \text{on}(III)} \) and negative-current-slope sampling point \( t_{k, \text{off}(III)} \) of \( k \)th phase, \((k-1)\)th phase current slope has the same sign (either both negative or positive). By neglecting the inductance and back EMF variation, (25) is derived. Substituting (25) for (15), error of \( k \)th phase self-inductance estimation due to mutual flux from \((k-1)\)th phase in Mode III can be calculated as (26). The self-inductance estimation error is zero, and therefore, the mutual flux coupling effect on self-inductance estimation is eliminated

\[
\frac{d i_{k-1}(t_{k, \text{on}(III)})}{d t} - \frac{d i_{k-1}(t_{k, \text{off}(III)})}{d t} = 0 \tag{25}
\]

\[
\text{err}_{\text{k(III)}} = 0 \tag{26}
\]

where \( \text{err}_{\text{m(III)}} \) is \( k \)th phase self-inductance estimation error due to mutual flux from \((k-1)\)th phase in Mode III.

The error analysis of the outgoing phase \((k-1)\)th phase self-inductance estimation in Mode I, II, and III can also be derived following the same procedure above. Based on the magnetic characteristics of the studied SRM shown in Fig. 3, the absolute values of self-inductance estimation error of phase A due to mutual flux from phase B and phase C in Mode I and II are shown in Fig. 7. Phase A self-inductance estimation error due to mutual flux from phase B and phase C is rotor position dependent. The mutual flux introduces maximum around 7% and minimum around 1% error to the phase A self-inductance estimation in Mode I and II. In Mode III, the mutual flux introduces 0% self-inductance estimation error. Therefore, the methods to ensure the self-inductance estimation working in Mode III exclusively are proposed and explained in the next section.

Based on the magnetic characteristics of the studied motor, the mutual flux introduces a maximum \pm 7% self-inductance estimation error in Mode I and II, while the mutual flux effect does not exist in Mode III. The proposed technique here is based on excluding Modes I and II and, hence, eliminating the error in self-inductance estimation due to the mutual flux. Two methods are proposed which utilizes the operation of Mode III exclusively for self-inductance estimation: variable-hysteresis-band current control for the incoming phase and variable-sampling method for the outgoing phase. In the variable-hysteresis-band current control method, when estimating the phase self-inductance of the incoming phase, the variation in the switching states of the other phase is avoided and the hysteresis band of the outgoing phase is adjusted. However, when the variable-hysteresis-band current control is applied to the outgoing-phase self-inductance estimation, undesirable higher current ripples might be observed in the incoming phase. For this reason, variable-sampling method for the outgoing-phase self-inductance estimation is proposed to overcome the drawback of variable-hysteresis-band current controller when applied to the outgoing-phase.

A. Variable-Hysteresis-Band Current Control for Incoming Phase

1) Principle of the Proposed Method for Self-Inductance Estimation of \( k \)th Phase (Incoming Phase): Fig. 8 illustrates the principle of the proposed variable-hysteresis-band current control during the \( k \)th phase self-inductance estimation.

During commutation, the \( k \)th phase current slope is much higher than that of \((k-1)\)th because of lower self-inductance of \( k \)th phase. The \((k-1)\)th phase current profiles with constant-hysteresis-band control and proposed variable-hysteresis-band control are denoted as solid and dotted line, respectively. The basic concept of the variable-hysteresis-band current control method is to make sure that the switching state of \((k-1)\)th phase is unchanged during the time intervals \( t_{k, \text{on}(II)} - t_{k, \text{off}(II)} \) and \( t_{k, \text{on}(1)} - t_{k, \text{off}(1)} \). Therefore, the sign of \((k-1)\)th phase current
slope is unchanged in these intervals. When the self-inductance estimation of $k$th phase is completed at $t_{k_{\text{off}}(1)}$ and $t_{k_{\text{off}}(1)}$, switches of $(k-1)$th phase are turned OFF or ON according to the error between $(k-1)$th phase current and its reference. When $(k-1)$th phase current at $t_{k_{\text{off}}(1)}$ or $t_{k_{\text{off}}(1)}$ is lower than its lower reference $i_{k-1\text{low}}$, switches are turned ON. When $(k-1)$th phase current $t_{k_{\text{off}}(1)}$ or $t_{k_{\text{off}}(1)}$ is higher than its upper reference $i_{k-1\text{up}}$, switches are turned OFF. Since the sign of $(k-1)$th phase current remains unchanged during the sampling interval, the $k$th phase self-inductance estimation is working in Mode III and mutual flux effect on self-inductance estimation is eliminated.

Hysteresis band for $k$th phase current control ($\Delta i_k$) stays constant. Its upper and lower reference is denoted as $i_{k_{\text{up}}}$ and $i_{k_{\text{low}}}$, respectively. However, in order to keep the switching state of $(k-1)$th phase unchanged during the $k$th phase self-inductance estimation, the hysteresis-band of $(k-1)$th phase ($\Delta i_{k-1}$) varies with time. As shown in Fig. 8, the hysteresis band of $(k-1)$th phase is represented as (27).

\[
\Delta i_{k-1}(t) = \left( i_{k-1\text{up}} + \Delta i_{k-1}(t) \right) - \left( i_{k-1\text{low}} - \Delta i_{k-1}(t) \right)
\]

(27)

where $\Delta i_{k-1}(t)$ and $\Delta i_{k-1}(t)$ are the adjusted hysteresis band of $(k-1)$th phase current in Mode I and II during $k$th phase self-inductance estimation, respectively.

2) Analysis of Adjusted Hysteresis Band: The proposed method increases the hysteresis band and it introduces more current ripples. Therefore, it is necessary to analyze the adjusted hysteresis band for $(k-1)$th phase. The adjusted hysteresis band of $\Delta i_{k-1}(t)$ and $\Delta i_{k-1}(t)$ varies with time. At time instants $t_I$ and $t_{II}$, the $(k-1)$th phase current reaches its upper and lower reference. Neglecting the mutual flux from $k$th phase, the $(k-1)$th phase voltages during the intervals $t_I - t_{k_{\text{off}}(1)}$ are derived as (28).

\[
U_{dc} = R_{I_k-1} + L_{k-1\text{I}_k-1} \frac{\Delta i_{k-1}(t)}{t_{k_{\text{off}}(1)} - t_I} + \frac{\partial L_{k-1\text{I}_k-1}}{\partial \theta} i_{k-1\text{I}_k-1} \omega_m.
\]

(28)

The adjusted hysteresis band for Mode I are derived as (29) according to (28).

\[
\Delta i_{k-1}(t) = \frac{(U_{dc} - R_{I_k-1} - \frac{\partial L_{k-1\text{I}_k-1}}{\partial \theta} i_{k-1\text{I}_k-1} \omega_m)t_{k_{\text{off}}(1)} - t_I}{L_{k-1\text{I}_k-1}}.
\]

(29)

In order to obtain the range of the adjusted hysteresis band, the range of back EMF should be obtained firstly. In current control mode, the back EMF of SRM cannot exceed the dc-link voltage, (30) has to be satisfied

\[
-U_{dc} \leq \frac{\partial L_{k-1\text{I}_k-1}}{\partial \theta} i_{k-1\text{I}_k-1} \omega_m \leq U_{dc}.
\]

(30)

Considering the maximum possible back-EMF in (30) and the resistive voltage drop, (31) has to be satisfied

\[
U_{dc} - R_{I_k-1} - \frac{\partial L_{k-1\text{I}_k-1}}{\partial \theta} i_{k-1\text{I}_k-1} \omega_m \\
\leq U_{dc} - \frac{\partial L_{k-1\text{I}_k-1}}{\partial \theta} i_{k-1\text{I}_k-1} \omega_m \leq U_{dc} + U_{dc}.
\]

(31)

In addition, as shown in Fig. 8, time intervals have to meet the constraints (32)

\[
t_{k_{\text{off}}(1)} - t_I \leq t_{k_{\text{off}}(1)} - t_{k_{\text{on}}(1)}.
\]

(32)

Based on (29), (31) and (32), the adjusted hysteresis band for Mode I must satisfy (33)

\[
\Delta i_{k-1}(t) = \left( U_{dc} - R_{I_k-1} - \frac{\partial L_{k-1\text{I}_k-1}}{\partial \theta} i_{k-1\text{I}_k-1} \omega_m \right) \frac{(t_{k_{\text{off}}(1)} - t_I)}{L_{k-1\text{I}_k-1}}.
\]

(33)

Therefore, the maximum adjusted hysteresis band of the outgoing phase in Mode I is derived as (34)

\[
\Delta i_{k-1}^{\text{max}}(t) = \frac{2U_{dc}\cdot t_{\text{sample}}}{L_{k-1\text{I}_k-1}}.
\]

(34)

where the required sample time is represented as

\[
t_{\text{sample}} = t_{k_{\text{off}}(1)} - t_{k_{\text{on}}(1)}.
\]

Similarly, the maximum adjusted hysteresis band in Mode II $\Delta i_{k-1}^{\text{max}}(t)$ can be derived, and it has the same expression with (34). With the same sampling time $t_{\text{sample}}$ in Mode I and II, the maximum adjusted hysteresis band in Mode I and Mode II is the same. The maximum adjusted hysteresis band is a function of self-inductance. During commutation, the $(k-1)$th phase (outgoing phase) $L_{k-1\text{I}_k-1}$ is close to aligned inductance $L_a$. Therefore, the maximum adjusted hysteresis band is approximated as (35). The aligned inductance is relatively large in SRM, and, therefore, only a slight increase in current hysteresis band and current ripple can be expected

\[
\Delta i_{k-1}^{\text{max}} = \frac{2U_{dc}\cdot t_{\text{sample}}}{L_a}.
\]

(35)

3) Drawback of The Proposed Method For Self-Inductance Estimation of $(k-1)$th Phase (Outgoing Phase): Fig. 9 illustrates the principle of $k$th phase variable-hysteresis-band current control during $(k-1)$th phase self-inductance estimation. During commutation, the $k$th phase self-inductance $L_{k\text{I}_k}$ is close to un-aligned inductance $L_u$ and the maximum adjusted hysteresis band is represented as (36)

\[
\Delta i_{k-1}^{\text{max}} = \frac{2U_{dc}\cdot t_{\text{sample}}}{L_u}.
\]

(36)
Since the unaligned inductance is much smaller than the aligned inductance $L_a$, the adjusted hysteresis band during $(k–1)$th phase self-inductance estimation is much higher than that during $k$th phase self-inductance estimation. As shown in Fig. 9, the hysteresis band of the $k$th phase is modified as (37) by applying variable-hysteresis-band current control

$$\Delta i_k = (i_{k, \text{up}} + \Delta i_{k, (I)}) - (i_{k, \text{low}} - \Delta i_{k, (II)}).$$

Therefore, the variable-hysteresis-band current control for outgoing phase self-inductance estimation has the drawback of high current ripples and torque ripples. In order to overcome the drawback of this method, a variable-sampling self-inductance estimation scheme will be presented for outgoing-phase self-inductance estimation in the next section.

### B. Proposed Variable-Sampling Scheme For Self-Inductance Estimation of $(k–1)$th Phase (Outgoing Phase)

Illustration of the variable-sampling self-inductance estimation scheme for $(k–1)$th phase is shown in Fig. 10. The positive phase current slope of $(k–1)$th phase is sampled at time instants $t_{k–1(1), \text{up}}$ and $t_{k–1(1), \text{off}}$, which are fixed. Since the phase current slope of $(k–1)$th phase is much lower than $k$th phase, the sign of $k$th phase current slope is changed several times during $(k–1)$th phase self-inductance estimation. Therefore, in Mode I, the $(k–1)$th phase negative-phase-current-slope sampling point $t_{k–1(1), \text{off}}$ can be adjusted to ensure $k$th phase current slope at $t_{k–1(1), \text{off}}$ and $t_{k–1(1), \text{on}}$ have the same signs. The same scheme is applied to Mode II. With the proposed method, the outgoing-phase self-inductance estimation is always operating in Mode III and, therefore, mutual flux from $k$th phase is eliminated. Since the phase current slope of $k$th phase (incoming phase) is much higher than $(k–1)$th phase, the sign of $(k–1)$th phase current slope is changed only once or not changed. Therefore, the variable-sampling scheme can not be applied to incoming phase self-inductance estimation.

Therefore, when incoming-phase self-inductance is estimated, the variable-hysteresis-band current controller is applied and the hystereis band of the outgoing phase is adjusted. However, when outgoing-phase self-inductance is estimated, the variable-sampling scheme is applied.

### V. Rotor Position Estimation at Rotating Shaft Conditions

Once the self-inductance of a phase is estimated, rotor position can be obtained based on the inductance-rotor position characteristics. Phase self-inductance is estimated only in the active region by using the phase current slope difference at rotating shaft condition. Therefore, each phase takes up one third of rotor period and three-phase inductance estimation will cover the total rotor period. The classification of the phase self-inductance estimation region is shown in Fig. 11(a). Self-inductance estimation is classified into phase A, B, and C self-inductance estimation. Phase self-inductance estimation regions are selected to avoid the inductance estimation near unaligned rotor position. This is because, near an unaligned position, the change of phase self-inductance with rotor position is relatively low and, therefore,
slight error in inductance estimation might lead to a much higher error in rotor position estimation.

The rotor position is estimated based on the corresponding phase self-inductance estimation region. When the estimated inductance of a phase reaches the maximum value, the self-inductance estimation is transferred to the next region. For example, when phase A self-inductance estimation region is selected, estimated phase A self-inductance is converted to the rotor position at each switching period by using rotor position-inductance characteristics. Once the estimated phase A inductance reaches $L_{\text{max}}$, phase inductance estimation is changed from phase A self-inductance estimation region to phase B self-inductance estimation region. The phase B self-inductance instead of phase A self-inductance is estimated and rotor position is updated based on estimated phase B self-inductance.

Linear TSF [30], [31] is used for instantaneous torque control of SRM. The torque reference of $k$th phase (incoming phase) and $(k-1)$th phase (outgoing phase) defined by linear TSF are expressed as (38)

$$T_k = T_{\text{ref}} \frac{\theta - \theta_{\text{on}}}{\theta_{\text{ov}}}$$

$$T_{k-1} = T_{\text{ref}} - T_{\text{ref}} \frac{\theta - \theta_{\text{off}}}{\theta_{\text{ov}}}$$

where $T_k$ and $T_{k-1}$ is the torque reference for the incoming phase and outgoing phase, respectively; $T_{\text{ref}}$ is the total torque reference; $\theta_{\text{on}}, \theta_{\text{off}}$, and $\theta_{\text{ov}}$ are the turn-on angle, turn-off angle and overlapping angle, respectively.

The waveform of linear TSF is shown in Fig. 11(b). As discussed previously, the variable-hysteresis-band current control is applied during the incoming phase self-inductance estimation, while variable-sampling phase inductance estimation is applied during the outgoing phase self-inductance estimation. Each phase self-inductance estimation region includes both incoming phase and outgoing phase self-inductance estimation, and therefore, two solutions exist in each self-inductance estimation region.

The flow charts of the proposed variable-hysteresis-band current control and variable-sampling self-inductance estimation methods are shown in Fig. 12 (a) and (b), respectively. In Fig. 12 (a), in order to ensure that the sign of the outgoing-phase current slope is unchanged during the incoming-phase self-inductance estimation, the outgoing phase current controller is disabled and enabled after the incoming-phase self-inductance estimation is finished.

The flowchart of the proposed rotor position estimation algorithm at both standstill and rotating shaft conditions is shown in Fig. 13.

VI. SIMULATION VERIFICATION

The proposed method to eliminate the mutual flux effect on rotor position estimation is compared to the rotor position estimation method without variable hysteresis band and sampling methods by simulations. The 2.3 kW, 6000 rpm, three-phase 12/8 SRM is simulated by MATLAB/Simulink using torque as well as inductance profiles given in Fig. 1. The SRM is driven by asymmetric power electronic converter with 300-V dc-link voltage. Linear TSF is used to generate the current reference for each phase. Turn-on angle $\theta_{\text{on}}$, turn-off angle $\theta_{\text{off}}$, and overlapping angle $\theta_{\text{ov}}$ of linear TSF are set to 5°, 20°, and 2.5°, respectively. From this point forward, all angles in this paper are denoted as mechanical angles. The sampling time $t_{\text{sample}}$ is set to 5 $\mu$s and the maximum adjusted hysteresis band is approximately 0.27 A according to (35). Hysteresis control is used to control the phase current and current hysteresis band is set to be 0.5 A. The inductance estimation error and rotor position estimation error is denoted as (39) and (40), respectively

$$\text{err}_L = \frac{L_{\text{real}} - L_e}{L_{\text{real}}}$$

$$\text{err}_\theta = \frac{\theta_{\text{real}} - \theta_e}{\theta_{\text{real}}}$$

(39)
where \( L_{\text{real}} \) and \( L_e \) are real inductance and estimated inductance, respectively; and \( \theta_{\text{real}} \) and \( \theta_e \) are real rotor position and estimated rotor position, respectively.

The inductance and, hence, rotor position is estimated at each switching period. Due to the hysteresis controller, switching period varies during the conduction period of a phase. Rotor position estimation error is the difference between estimated rotor position and the rotor position from the position sensor when sampled at the switching frequency. Real-time rotor position, which can be measured from the position sensor with a constant and much faster sampling, will be updated faster than estimated rotor position. Therefore, the real-time rotor position error (the difference between estimated rotor position at switching frequency and the measured rotor position at a constant and higher sampling frequency) is updated faster than rotor position error (the difference between estimated and measured rotor positions at switching frequency). At this point, rotor position estimation error differs from real-time rotor position estimation error. As the speed increases, the estimated inductance or rotor position is updated slower due to the larger switching period. Therefore, the real-time rotor position estimation error (the difference between estimated rotor position at switching frequency and the measured rotor position at a constant and higher sampling frequency) is also updated faster than rotor position error (the difference between estimated and measured rotor positions at switching frequency). By using the proposed variable-hysteresis-band current controller and variable-sampling self-inductance estimation to eliminate the mutual flux effect, the maximum inductance estimation error is decreased to \( \pm 0.6\% \). As a result, the maximum rotor position estimation error and real-time rotor position estimation error are decreased to \( \pm 0.1^\circ \) and \( \pm 0.5^\circ \). A zoom in plot of time intervals \( a \) and \( b \) of Fig. 14 is shown in Fig. 15 and Fig. 16, respectively.

As shown in Fig. 15(a), the phase C (outgoing phase) rotor position estimation without variable hysteresis band and sampling is working both in Mode I and Mode II. Due to mutual flux effect of phase A on phase C (incoming phase), Mode I lead to about \( \pm 7\% \) self-inductance estimation error and \( \pm 1.5^\circ \) position estimation error, while Mode II lead to approximately \( \pm 7\% \) inductance estimation error and \( \pm 1.5^\circ \) position estimation error. By applying the proposed variable-sampling outgoing-phase

Fig. 14. Simulation results of rotor position estimation (\( T_{\text{ref}} = 0.375 \text{ Nm} \) and \( \text{Speed} = 1200 \text{ rpm} \)). (a) The rotor position estimation method without variable hysteresis band and sampling. (b) The proposed rotor position estimation.

A. Simulation Results at 1200 rpm (\( T_{\text{ref}} = 0.375 \text{ Nm} \))

The torque reference is set to be 0.375 Nm and operational speed of SRM is 1200 rpm. Fig. 14 shows simulation results of the proposed rotor position estimation with and without variable hysteresis band and sampling methods applied at 1200 rpm.

The maximum self-inductance estimation error of the rotor position estimation method without variable hysteresis band and sampling is \( \pm 7\% \), which matches theoretical analysis given in
rotor position estimation method, the phase C position estimator is working exclusively in Mode III shown in Fig. 15(b), and therefore, the phase A mutual flux effect on phase C is eliminated.

Similarly, as shown in Fig. 16(a), phase A (incoming phase) rotor position estimation without variable hysteresis band and sampling is working both in Mode I, Mode II, and Mode III. Due to mutual flux effect of phase C on phase A, Mode I lead to approximately +1.5% inductance estimation error and +0.05° position estimation error, while Mode II lead to approximately −1.5% and −0.05° position estimation error. By applying the proposed variable-hysteresis-band current controller for incoming-phase self-inductance estimation, the phase A position estimation is working exclusively in Mode III as shown in Fig. 16(b), and therefore, the phase C mutual flux effect on phase A is eliminated. Compared with the phase A (incoming phase) mutual flux effect on phase C (outgoing phase) as shown in Fig. 15, the mutual flux effect of phase C on phase A is negligible. However, we still see minor improvement by using variable-hysteresis band current controller. The elimination of mutual flux effect on rotor position estimation is mainly contributed by variable-sampling outgoing-phase rotor position estimation method.

B. Simulation Results at 4500 rpm ($T_{ref} = 0.375 \text{ Nm}$)

The torque reference is set to be 0.375 Nm and operational speed of SRM is 4500 rpm. Fig. 17 shows simulation results of the proposed rotor position estimation with and without variable hysteresis band and sampling at 4500 rpm. The inductance estimation error of the rotor estimation method without variable hysteresis band and sampling is −7%, which matches theoretical analysis given in Fig. 7. Due to inductance estimation error, the rotor estimation error and the real-time rotor position estimation error are −1.5° and ±2°, respectively. Compared with simulation results at 1200 rpm, the real-time rotor position estimation error is increased due to larger switching period. Also, at 4500 rpm, the phase self-inductance estimation only works at Mode I and Mode III, which lead to only negative inductance estimation error. By using the proposed variable-hysteresis-band current controller and variable-sampling inductance estimation to eliminate the mutual flux effect, the maximum inductance estimation error is decreased to +2.5%. As a result, the rotor position estimation error and real-time rotor position estimation error are decreased to +0.5° and +2°. Both self-inductance estimation error and rotor position estimation error is nonnegative, and thus, Mode I is avoided by using the proposed rotor position estimation method.

C. Simulation Results at 6000 rpm ($T_{ref} = 0.2 \text{ Nm}$)

The torque reference is set to be 0.2 Nm and operational speed of SRM is 6000 rpm. Fig. 18 shows simulation results of the proposed rotor position estimation with and without variable hysteresis band and sampling at 6000 rpm. The self-inductance estimation error of the rotor estimation method without variable hysteresis band and sampling is −7%, leading to −1.5° rotor position estimation error. The real-time rotor estimation error is −1.5° and +2.5°. The negative real-time rotor position estimation error is mainly contributed by mutual flux and positive real-time error is mostly due to larger switching period. Similarly, by using the proposed rotor position estimation method, negative real-time rotor position estimation error is eliminated due to elimination of mutual flux effect.

VII. EXPERIMENTAL RESULTS

The proposed variable-hysteresis-band current control and variable-sampling rotor position estimation methods are
compared to the rotor position estimation method without variable hysteresis band and sampling experimentally on a 2.3 kW, 6000 rpm, three-phase 12/8 SRM. The experimental setup is shown in Fig. 19. FPGA EP3 C25Q240 is used for digital implementation of the proposed rotor position estimation method. Current hysteresis band is set to be 0.5 A and dc-link voltage is set to 300 V. The sampling time $t_{\text{sample}}$ is also set to 5 $\mu$s. The self-inductance characteristics are stored as look up tables in FPGA. Rotor position is estimated from this look-up table using the estimated phase self-inductance.

A. Experimental Results at 4500 rpm ($T_{\text{ref}} = 0.375 \text{ Nm}$)

The torque reference is set to 0.375 Nm in this experiment. Fig. 20 shows experimental results of the proposed rotor position estimation with and without variable hysteresis band and sampling at 4500 rpm. From the experimental results, it can be noticed that real-time rotor position estimation error has positive bias. This is because the selected digital-to-analog conversion chip is unipolar. Therefore, 5.625° offset is added to rotor position error in the next a couple of figures. The real-time rotor position estimation error without variable hysteresis band and sampling is $+5^\circ$ and $-3.3^\circ$. By using the proposed method, the real-time rotor position estimation error of the proposed method is decreased to $+2.8^\circ$ and $-1.7^\circ$. In the experiment, phase current sensing contains both the noise and quantization error, leading to slightly higher phase current slope sensing error. Therefore, compared with simulation results, the real-time rotor position estimation error is increased. However, the proposed method still shows an increase of approximately 2° in rotor position estimation accuracy.

B. Experimental Results at 6000 rpm ($T_{\text{ref}} = 0.2 \text{ Nm}$)

The torque reference is set to 0.2 Nm in this experiment. Fig. 21 shows experimental results of the proposed rotor position estimation with and without variable hysteresis band and sampling at 6000 rpm. Similarly, 5.625° offset is added to rotor position error in the next a couple of figures. The real-time rotor position estimation error for the method without variable hysteresis band and sampling is $+5^\circ$ and $-2.8^\circ$. By using the proposed method, the proposed method shows only positive rotor position estimation error up to $+5^\circ$, and therefore, mutual flux effect on rotor position estimation in Mode I is eliminated.

VIII. CONCLUSION

In this paper, an approach to eliminate the mutual flux effect on rotor position estimation of SRM drives at light load conditions is presented. This method has the advantage of no need of a priori knowledge of mutual flux and external voltage injection at rotating shaft condition. According to theoretical analysis of the studied motor, with phase current slope difference method, $\pm 1\%$ to $\pm 7\%$ self-inductance estimation error is introduced by the mutual flux between two conducting phases during commutation. However, the mutual flux effect on self-inductance estimation is eliminated by using the proposed variable-hysteresis-band current control for the incoming-phase self-inductance estimation and variable-sampling method for the outgoing-phase self-inductance estimation. Based on the accurate estimated self-inductance, the accuracy of rotor position estimation is improved. The effectiveness of the proposed method is verified by both simulation and experimental results with a 2.3 kW, 6000 rpm, three-phase 12/8 SRM. The results show that the proposed method improves around 2° rotor position estimation accuracy compared with the method without variable hysteresis band and sampling.
Fig. 21. Experimental result of rotor position estimation at 6000 rpm (T_{ref} = 0.2 Nm). (a) The rotor position estimation without variable hysteresis band and sampling. (b) The proposed rotor position estimation.

REFERENCES


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