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Xu, Wei; Jiang, Yajie; Mu, Chaoxu; Blaabjerg, Frede

Published in:
IEEE Transactions on Power Electronics

DOI (link to publication from Publisher):
10.1109/TPEL.2018.2822769

Publication date:
2019

Document Version
Accepted author manuscript, peer reviewed version

Link to publication from Aalborg University

Citation for published version (APA):

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Improved Nonlinear Flux Observer Based Second-Order SOIFO for PMSM Sensorless Control

Wei Xu, Senior Member, IEEE, Yajie Jiang, Chaoxu Mu, Member, IEEE, and Frede Blaabjerg, Fellow, IEEE

Abstract—The conventional rotor flux estimation method has issues of dc offset and harmonics, which are caused by initial rotor flux, detection errors, etc. To eliminate these defects, one improved nonlinear flux observer is proposed for sensorless control of permanent magnet synchronous machine (PMSM). Firstly, the rotor position estimation method based on PMSM rotor flux observation is studied. Meanwhile, the limitations of the traditional rotor flux estimators, i.e., the saturation of pure integrator, phase shift and amplitude attenuation of low-pass filter are analyzed. Then, two novel flux observers, second-order generalized integral flux observer (SOIFO) and second-order SOIFO are designed for the rotor flux estimation of PMSM. Based on second-order generalized integrator (SOGI) structure, the SOIFO can limit the dc component to a certain value. Furthermore, the second-order SOIFO is developed from the SOGI, which is characterized with effective dc and harmonics attenuation capability. With the second-order SOIFO, even without magnitude and phase compensation, the dc offset and harmonics of estimated rotor flux could be well eliminated. Therefore, the speed and rotor position can be estimated accurately. All the performances of the four methods are analyzed by transfer functions and Bode diagrams. Lastly, the new sensorless control strategy is validated by comprehensive experimental results.

Index Terms—Rotor flux estimation, permanent magnet synchronous machine (PMSM), second-order generalized integral flux observer (SOIFO), sensorless control.

NOMENCLATURE

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
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<tbody>
<tr>
<td>$u_s$</td>
<td>Stator voltage vector ($\alpha, \beta$ axis, respectively)</td>
</tr>
<tr>
<td>$R_s$</td>
<td>Stator resistance</td>
</tr>
<tr>
<td>$i_s$</td>
<td>Stator current vector ($\alpha, \beta$ axis, respectively)</td>
</tr>
<tr>
<td>$p$</td>
<td>Differential operator</td>
</tr>
<tr>
<td>$\psi_s$</td>
<td>Stator flux vector ($\alpha, \beta$ axis, respectively)</td>
</tr>
<tr>
<td>$L$</td>
<td>Stator inductance</td>
</tr>
<tr>
<td>$\psi_f$</td>
<td>Permanent flux linkage</td>
</tr>
<tr>
<td>$\theta_e$</td>
<td>Rotor electrical position</td>
</tr>
<tr>
<td>$\omega_e$</td>
<td>Rotor electrical angular velocity</td>
</tr>
<tr>
<td>$\dot{\theta}_e$</td>
<td>Estimated rotor position</td>
</tr>
<tr>
<td>$\dot{\omega}_e$</td>
<td>Estimated speed</td>
</tr>
<tr>
<td>$e_s$</td>
<td>Back-EMFs ($\alpha, \beta$ axis, respectively)</td>
</tr>
<tr>
<td>$R_{s0}$</td>
<td>Fixed nominal resistance</td>
</tr>
<tr>
<td>$L_0$</td>
<td>Fixed nominal inductance</td>
</tr>
<tr>
<td>$\Delta R_s$</td>
<td>Resistance variation</td>
</tr>
<tr>
<td>$\Delta L$</td>
<td>Inductance variation</td>
</tr>
<tr>
<td>$e_{r0}$</td>
<td>Initial back-EMF vector ($\alpha, \beta$ axis, respectively)</td>
</tr>
<tr>
<td>$A_0$</td>
<td>DC components of back-EMF ($\alpha, \beta$ axis, respectively)</td>
</tr>
<tr>
<td>$A_1$</td>
<td>Amplitude of fundamental waves ($\alpha, \beta$ axis, respectively)</td>
</tr>
<tr>
<td>$A_h$</td>
<td>Amplitude of harmonics ($\alpha, \beta$ axis, respectively)</td>
</tr>
<tr>
<td>$\varphi_1$</td>
<td>Initial angel of the fundamental wave of back-EMF</td>
</tr>
</tbody>
</table>

ABBREVIATIONS

<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Description</th>
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</thead>
<tbody>
<tr>
<td>PMSM</td>
<td>Permanent magnet synchronous machine</td>
</tr>
<tr>
<td>FOC</td>
<td>Field-orientation control</td>
</tr>
<tr>
<td>EMF</td>
<td>Electromotive force</td>
</tr>
<tr>
<td>LPF</td>
<td>Low-pass filter</td>
</tr>
</tbody>
</table>

\[ \Psi_{r,s}(t) = \left[ \psi_{r,s}(t) \quad \psi_{r,\beta}(t) \right]^T \]

Estimated rotor flux by pure integrator (\( \alpha, \beta \) axis, respectively)

\[ \omega_c \]

Cut-off frequency of low-pass filter

\[ \Psi_{r,LPF}(t) = \left[ \psi_{r,s,LPF}(t) \quad \psi_{r,\beta,LPF}(t) \right]^T \]

Estimated rotor fluxes by LPF (\( \alpha, \beta \) axis, respectively)

\[ \omega \]

Center frequency of SOGI

\[ k \]

Coefficient of SOGI

\[ \Psi_{r,SOIFO}(s) = \left[ \psi_{r,s,SOIFO}(s) \quad \psi_{r,\beta,SOIFO}(s) \right]^T \]

Observed rotor fluxes by second-order SOIFO (\( \alpha, \beta \) axis, respectively)

\[ \psi_{r,s,SOIFO}(s) = \left[ \psi_{r,s,SOIFO}(s) \quad \psi_{r,\beta,SOIFO}(s) \right]^T \]

Observed rotor fluxes by second-order SOIFO (\( \alpha, \beta \) axis, respectively)

\[ t_s \]

Setting time

\[ T_s \]

Sample period

I. INTRODUCTION

Permanent magnet synchronous machine (PMSM) drive system with the field-orientation control (FOC) has been widely used in industrial applications, for its high performance, low torque ripple, fast current response, etc. In general, the mechanical sensor can provide the necessary rotor position information for the implementation of FOC, which leads to higher cost and lower reliability of PMSM drive. Therefore, the rotor position estimation involving sensorless control technique for PMSM drives is in the ascendant [1]-[4].

The sensorless control methods include high-frequency signal injection, back-electromotive force (EMF) estimator and rotor flux observer. For low-speed operations, the high-frequency signal injection methods, including sinusoidal-wave and square-wave, have been widely used in PMSM systems [5]-[7].

For medium- and high-speed regions, the universal sensorless control scheme is the back-EMF estimator, where the rotor position is extracted from the fundamental wave of back-EMF [8]-[11]. The back-EMF extraction-based sensorless control algorithms can be implemented by various methods, for example, extended Kalman filter [1], [8], sliding mode observer [9], Luenberger observer [10], artificial intelligence-based estimator [11], etc. These methods have been investigated in a large amount of literature. However, the back-EMF cannot be precisely extracted with noises at low speed [9]. Therefore, the back-EMF estimation scheme is not qualified at low-speed.

Apart from the back-EMF, the PMSM rotor position information can also be acquired from the rotor flux via direct trigonometric relations [12]. Benefiting from that the magnitude of rotor flux remains constant in the transient state, the rotor flux estimation method has the potential to become a general sensorless control scheme for the PMSM drive systems. However, the traditional rotor flux observation methods, i.e. the pure integrator and low-pass filter (LPF), cannot provide the accurate rotor flux.

On basis of the PMSM voltage and flux models, the pure integrator is a universal part in building a traditional rotor flux observer [13], [14]. However, the dc offset and harmonics of the estimated flux are generated by the parameter mismatches, unknown integral initial value, voltage and current detection errors and converter disturbances. Hence, the accuracy of estimated speed and rotor position are very low. Especially, the observed flux would be deviated from the real flux because of the saturation effect, which is mainly caused by the unknown integral initial value [15]. Therefore, some works focus on improving the pure integrator-based rotor flux observer. In [14] and [15], one initial flux condition estimator based gradient algorithm is developed to estimate the unknown integral initial value, which is used for the rotor flux compensation. In [16] and [17], the stability is introduced to rotor flux observer with new coefficients designs. A disturbance observer is proposed in [18], in which the rotor flux is obtained from the stator flux integrator. In [19], the parameter identification is used in rotor flux observation. Although these approaches can improve the estimation accuracy of rotor flux, the effectiveness and practicability of these methods cannot be guaranteed yet.

The LPF is another one of flux estimation methods which can effectively reduce the dc offset and harmonics, but it also leads to the amplitude attenuation and phase shift [14]. Meanwhile, the LPF method requires the rotor phase compensation, which means that the machine speed information is needed.

By intensively analyzing the existing work, the rotor flux observation based sensorless control strategy has not be adequately investigated. Thanks to the advanced filter capability and low computational cost, the second-order generalized integrator (SOGI) has been commonly used for filter, quadrature signal generation and phase extraction [20], [21]. By analyzing its transfer functions, the SOGI can be considered as a filter combined with an integrator. Moreover, the amplitude attenuation and phase shift can be avoided. Thereby, in [22], the SOGI is used for stator flux estimation in an induction machine drive system.

Till date, the SOGI has not been used for flux observation in PMSM drive system. Therefore, on basis of SOGI, a novel second-order generalized integral flux observer (SOGIFO) is designed in this paper, to remove the aforementioned drawbacks of both the pure integrator and LPF methods. However, the theoretical analysis shows that the eliminating ability of the SOIFO is not excellent. To solve this problem, an improved second-order SOIFO structure is proposed. With a fourth-order transform function, the second-order SOIFO has strong attenuation ability against the dc offset and harmonics. So the above application limitations of rotor flux observation in parameter mismatches, external
disturbances and unknown integral initial value are removed, while the rotor flux, rotor position and speed are estimated accurately. The proposed sensorless control scheme can be used in many industrial applications, such as the electric vehicle, servo system, and so on. Its advantages contain wide speed region, simple implementation, abilities to eliminate the dc offset and harmonics. Thereby, the extra parameter identification and disturbance attenuation structures are not needed. Specially, the proposed method can be used at 1% - 100% rated speed region, while most back-EMF estimation-based sensorless control strategy could not work effectively below 5% rated speed region. Furthermore, the detailed discretization implementation of the second-order SOGI is also proposed in this paper.

The structure of this paper is organized as follows. In Section II, the traditional PMSM rotor flux observers and their performance are studied. The SOIFO and its second-order version for the PMSM rotor flux observation are proposed and intensively investigated in Section III. In Section IV, comprehensive experimental results are presented. Finally, conclusions are drawn in Section V.

II. CONVENTIONAL PMSM ROTOR FLUX OBSERVERS

As stated in [8], the voltage and flux equations of PMSM in the stationary coordinate can be described as

\[ \mathbf{u}_s = \mathbf{R}_s \cdot \mathbf{i}_s + \mathbf{p} \mathbf{\Psi}_s \]

\[ \mathbf{\Psi}_s = \mathbf{L} \cdot \mathbf{i}_s + \mathbf{\psi}_f \]

where \( \mathbf{u}_s \) is the stator voltage vector, \( \mathbf{i}_s \) is the stator current vector, \( \mathbf{\Psi}_s \) is the stator flux vector, \( \mathbf{u}_s = [u_a \ u_b]^T \), \( \mathbf{i}_s = [i_a \ i_b]^T \), \( \mathbf{\psi}_f = [\psi_{fa} \ \psi_{fb}]^T \), \( \mathbf{L} \) is the stator inductance, \( \mathbf{R}_s \) is the stator resistance, \( \mathbf{p} \) is the differential operator, \( \theta_e \) is the rotor electrical position, \( \mathbf{\psi}_f \) is the flux linkage, \( u_a \) and \( u_b \) are the \( \alpha \)- and \( \beta \)-axis stator voltages, \( i_a \) and \( i_b \) are the stator currents, \( \psi_{fa} \) and \( \psi_{fb} \) are the stator flux. The rotor flux of PMSM can be given as

\[ \mathbf{\Psi}_r = \mathbf{\Psi}_s - \mathbf{L} \cdot \mathbf{i}_s = \mathbf{\psi}_f \begin{bmatrix} \cos(\theta_e) \\ \sin(\theta_e) \end{bmatrix} \]

where \( \mathbf{\Psi}_r \) is the rotor flux vector, \( \mathbf{\Psi}_r = [\psi_{ra} \ \psi_{rb}]^T \). It can be seen that the rotor position and speed can be extracted from the rotor flux, which are given as

\[ \theta_e = \tan^{-1}(\psi_{rb}/\psi_{ra}) \]

\[ p\dot{\theta}_e = \omega_e \]

where \( \omega_e \) is the rotor electrical angular velocity. The PMSM sensorless control based on the rotor flux observation is shown in Fig. 1, where \( \dot{\mathbf{\Psi}}_r \) is the observed rotor flux vector, \( \dot{\mathbf{\Psi}}_r = [\dot{\psi}_{ra} \ \dot{\psi}_{rb}]^T \), \( \dot{\theta}_e \) is the estimated rotor position, and \( \dot{\omega}_e \) is the estimated speed.

As shown in (1) and (4), rotor flux is the integration of back-EMF, which can be given as

\[ \mathbf{\Psi}_r = \int \left( \mathbf{u}_s - \mathbf{R}_s \cdot \mathbf{i}_s - \mathbf{L} \cdot \mathbf{\dot{p}} \right) dt = \int \mathbf{e}_r dt \]

where \( \mathbf{e}_r \) is the back-EMF vector, \( \mathbf{e}_r = [e_{ra} \ e_{rb}]^T \). The integral initial value of the estimated rotor flux is defined as \( \mathbf{\Psi}_r(0) = \mathbf{\Psi}_r(0) - L \cdot \mathbf{i}_s(0) \). Meanwhile, considering the disturbances of a practical PMSM drive system, i.e., parameter mismatches, unknown integral initial value, detection errors and converter nonlinearities, (6) is modified as

\[ \mathbf{\Psi}_r = \int \left( \mathbf{u}_s - \left( R_{r0} + \Delta R \right) \mathbf{i}_s - \left( L_{r0} + \Delta L \right) \cdot \mathbf{\dot{p}} \right) dt + \mathbf{e}_r + \mathbf{\chi} \]

where \( R_{r0} \) and \( L_{r0} \) are the fixed nominal parameters, \( \Delta R \) and \( \Delta L \) are the parameter variations, \( \mathbf{e}_r \) is the initial back-EMF vector, \( \mathbf{e}_r = [e_{r0} \ e_{r0}]^T \), \( e_{r0} dt = \mathbf{\Psi}_r(0) \), and \( \mathbf{\chi} \) represents the other disturbances. Then, the back-EMF can be written as

\[ \mathbf{e}_r = \mathbf{A}_r + \mathbf{A}_s \sin(\alpha t + \phi_1) + \sum \mathbf{A}_h \sin(\alpha h t + \phi_h) \]

where \( \mathbf{A}_r \) is the dc component, \( \mathbf{A}_s = [\mathbf{A}_{sa} \ \mathbf{A}_{sb}]^T \), \( \mathbf{A}_s \sin(\alpha t + \phi_1) \) is the fundamental components, \( \mathbf{A}_h \) is the amplitude of fundamental wave, \( \mathbf{A}_h = [\mathbf{A}_{ha} \ \mathbf{A}_{hb}]^T \), \( \sum \mathbf{A}_h \sin(\alpha h t + \phi_h) \) is the sum of harmonics, \( \mathbf{A}_h \) is the amplitude of harmonics, \( \mathbf{A}_h = [\mathbf{A}_{ha} \ \mathbf{A}_{hb}]^T \), \( \phi_1 \) and \( \phi_h \) are the initial angel of fundamental wave and harmonics of the back-EMF, \( \alpha_h \) and \( \alpha_h \) are the corresponding angular frequencies, and \( h \) is the order of harmonics. Take Laplace transformation of (8), it can be written as

\[ \mathbf{E}_r(s) = \frac{\mathbf{A}_r}{s} + \frac{\mathbf{A}_s \sin(\alpha t + \phi_1)}{s^2 + \alpha_1^2} + \sum \frac{\mathbf{A}_h \sin(\alpha h t + \phi_h)}{s^2 + \alpha_h^2} \]

where \( \mathbf{E}_r(s) \) is the Laplace-transform of \( \mathbf{e}_r \), \( s \) the Laplace operator. Taking pure integrator 1/s as flux observer, the observed rotor flux \( \mathbf{E}_r(s)/1/s \) can be given as

\[ \mathbf{\Psi}_{r,s}(t) = \mathbf{A}_r t + \frac{\mathbf{A}_s}{\alpha_1} \sin(\alpha t + \phi_1 - 0.5\pi) + \frac{\mathbf{A}_s \cos(\phi_1)}{\alpha_1} \]

\[ + \sum \frac{\mathbf{A}_h \sin(\alpha h t + \phi_h - 0.5\pi)}{\alpha_h} + \sum \frac{\mathbf{A}_h \cos(\phi_h)}{\alpha_h} \]

where \( \mathbf{\Psi}_{r,s}(t) \) is the estimated rotor flux vector by the pure integrator, \( \mathbf{\Psi}_{r,s}(t) = [\psi_{rsa}(t) \ \psi_{rsb}(t)]^T \). In (10), except the fundamental part \( \mathbf{A}_r \sin(\alpha t + \phi_1 - 0.5\pi)/\alpha_1 \), the dc component and harmonics are also contained in the estimated rotor flux. Moreover, the dc component consists of two parts.
where \( A_i \cos(\varphi_i) + \sum A_{n} \cos(\varphi_{n}) \) is constant, and \( A_f \) linearly increases with time. Especially, the time-increasing term \( A_f \), resulted from the integral initial value, would lead to the saturation and serious distortion of flux. In this case, the sensorless control cannot be achieved ultimately. In addition, the harmonics of estimated rotor flux also affect the accuracy of position observation, especially when the PMSM is operated at low speed.

Another rotor flux estimator is LPF: \( 1/(s + \omega_c) \), where \( \omega_c \) is the cut-off frequency. Taking the inverse Laplace transform of \( E_r(s) \cdot 1/(s + \omega_c) \), the estimated rotor flux can be described as

\[
\Psi_{r_{LPF}}(t) = \frac{A_i}{\omega_c} \cdot e^{-\omega_c t} + \frac{A_i}{\omega_c^2} \cdot \sin(\omega_c t + \varphi_i - 0.5\pi + \theta_i) + A_i \cos(\varphi_i + \theta_i) e^{-\omega_c t} + \sum A_{n} \cos(\varphi_{n} + \theta_{n}) \sqrt{\omega_n^2 + \omega_c^2}
\]

where \( \omega_c \) is the center frequency. It is well known that \( D(s) \) is used for filtering, and \( Q(s) \) is used for the integration. In this way, a novel PMSM rotor flux observation method, SOIFO, can be constructed as

\[
\Psi_{r_{SOIFO}}(s) = \frac{1}{\omega_c} \cdot Q(s) \cdot E_r(s) = \frac{k_\omega}{s + k_\omega \cdot \omega_c} \cdot E_r(s) = \frac{1}{s} E_r(s)
\]

where \( \Psi_{r_{SOIFO}}(s) \) is the observed rotor flux vector by SOIFO. \( \Psi_{r_{SOIFO}}(s) = \left[ \Psi_{r_{\alpha_{SOIFO}}} \quad \Psi_{r_{\beta_{SOIFO}}} \right]^T \). The structure of SOIFO is shown in Fig. 3. In order to guarantee \( \omega = \omega_c \), a frequency locked loop (FLL) is also proposed with SOIFO. In the figure, \( \hat{E}_r(s) \) is the filtered output, \( e_k \) is an intermediate variable, and \( \Gamma \) is a coefficient.

### III. Flux Observer Based Second-Order SOIFO

To eliminate the limitations of the traditional methods, a SOIFO on basis of the SOGI structure is firstly proposed for the PMSM rotor flux observation in this paper. With the SOIFO, the dc offset and harmonics in the observed rotor flux are reduced. Furthermore, an improved second-order SOIFO with a four-order transform function is designed to enhance the ability to eliminate the dc offset and harmonics.

#### A. SOIFO

As stated in [23], the SOGI has been used for the phase and amplitude extractions of grid voltage. Its structure is shown in Fig. 2.

In Fig. 2, \( v \) is the input signal, \( \nu \) and \( q\nu \) are two outputs. The transfer functions are given as

\[
D(s) = \frac{v(s)}{v(s)} = \frac{k_\omega s}{s^2 + k_\omega s + \omega_c^2} \tag{12}
\]

\[
Q(s) = \frac{q\nu(s)}{v(s)} = \frac{k_\omega^2}{s^2 + k_\omega s + \omega_c^2} \tag{13}
\]

where \( \omega_c \) is the center frequency. It is well known that \( D(s) \) is used for filtering, and \( Q(s) \) is used for the integration. As shown in (14), \( Q(s) \) can be considered as an integrator. In this way, a novel PMSM rotor flux observation method, SOIFO, can be constructed as

\[
\Psi_{r_{SOIFO}}(s) = \frac{1}{\omega_c} \cdot Q(s) \cdot E_r(s) = \frac{k_\omega}{s + k_\omega \cdot \omega_c} \cdot E_r(s) = \frac{1}{s} E_r(s)
\]

where \( \Psi_{r_{SOIFO}}(s) \) is the observed rotor flux vector by SOIFO. \( \Psi_{r_{SOIFO}}(s) = \left[ \Psi_{r_{\alpha_{SOIFO}}} \quad \Psi_{r_{\beta_{SOIFO}}} \right]^T \). The detailed calculation is provided in Appendix C. As shown in (16), with the reduced dc offset of estimated rotor flux, the saturation-effect is removed without both phase shift and amplitude attenuation. Meanwhile, the magnitude of harmonics is inversely proportional to \( \sqrt{(1-h^2)^2 + k^2 h^2 + 1} \), which can be simplified as \( \sqrt{(1-h^2)^2} \), \( h \gg 1 \). It means that the amplitudes of harmonics are reduced. The dc component of rotor flux is reduced to \( A_k / \omega_1 \). However, because \( \Psi_{r_{SOIFO}}(s) \) is the low-pass--filtered version of the input, the dc component of rotor flux by SOIFO cannot be removed.
completely, which would result in rotor position estimation error.

B. Second-Order SOIFO

Recently, a novel second-order SOGI with enhanced ability to reject harmonic and dc offset is introduced in [21]. The transfer functions of second-order SOGI are described as

\[ D_s(s) = \frac{v(s)}{v(s)} = \frac{K_1 K_2 \omega^3 s^3}{s^4 + K_1 \omega s^3 + (2 + K_2) \omega^2 s^2 + K_2 \omega^3 s + \omega^4} \]  

(17)

\[ Q_s(s) = \frac{q(s)}{v(s)} = \frac{K_1 K_2 \omega^3 s^3}{s^4 + K_1 \omega s^3 + (2 + K_2) \omega^2 s^2 + K_2 \omega^3 s + \omega^4} \]  

(18)

where \( K_1 \) and \( K_2 \) are two coefficients. Similarly with SOGI, \( D_s(s) \) is used for filtering, and \( Q_s(s) \) is used for the integration. In the steady state, it is defined \( s = j \omega \), then \( q' \) can be regarded as the integration of \( v \). The relationship between the input and \( q'(s) \) of second-order SOGI can be equally accessed as

\[ Q_s(s) = \frac{K_1 K_2 \omega^3 s^3}{s^4 - jK_1 \omega s^3 + K_2 K_1 \omega^2 s^2 - 2\omega^3 + jK_2 \omega^3 s + \omega^4} = \omega \frac{1}{s} \]  

(19)

Taking the back-EMF as the input of second-order SOGI, \( q' \) can be considered as the integration of back-EMF. By further modification, the second-order SOIFO can be obtained as

\[ \Psi_{r,SOSIFO}(s) = \frac{1}{\omega} \frac{K_1 K_2 \omega^3 s^3}{s^4 + K_1 \omega s^3 + (2 + K_2) \omega^2 s^2 + K_2 \omega^3 s + \omega^4} \]  

where \( \Psi_{r,SOSIFO}(s) \) is the observed rotor flux vector by second-order SOIFO, \( \Psi_{r,SOSIFO}(s) = [\Psi_{r,SOSIFO}(s) \quad \Psi_{r,SOSIFO}(s)]^T \). The structure of second-order SOIFO is given in Fig. 4, where a FLL is applied for the frequency adaption. And the steady-state time domain of (20) is written as

\[ \Psi_{r,SOSIFO}(t) = \frac{A_1}{\omega_1} \sin(\omega_1 t + \phi_1 - 0.5\pi) + \]  

Fig. 4. The diagram of second-order SOIFO/SOGI.

\[ \sum_{h=1}^{\infty} \left[ \frac{1}{(h^2 K_1)^2} \right] \left[ \frac{1}{(h^2 K_1)^2} \right] \left[ \frac{1}{(h^2 K_1)^2} \right] = \left( \frac{h^2}{K_1} \right)^2 + \left( \frac{h^2}{K_1} \right)^2 \left( \frac{h^2}{K_1} \right)^2 > h \gg 1 \right. \]  

(21)

where \( \gamma_{h2} = \tan^{-1} \left( \frac{K_2 \omega^3 - K_1 \omega^3}{\omega^4 - h^2 (2 + K_1 K_2) \omega^2 + \omega^4} \right) \). The detailed calculation process has been proposed in Appendix D. (21) shows that the dc component of the rotor flux observation is eliminated. Meanwhile, the magnitude of high-order harmonics is inversely proportional to

\[ \sqrt{\frac{1}{(h^2 K_1)^2} \left( \frac{h^2}{K_1} \right)^2 + \left( \frac{h^2}{K_1} \right)^2 \left( \frac{h^2}{K_1} \right)^2} \]  

which can be simplified as

\[ \sqrt{\frac{1}{(h^2 K_1)^2} \left( \frac{h^2}{K_1} \right)^2 + \left( \frac{h^2}{K_1} \right)^2 \left( \frac{h^2}{K_1} \right)^2} = \frac{h^2}{K_1 K_2} \gg h \gg 1 \]  

It means that the second-order SOIFO can produce the lower gain magnitude at higher frequencies than that of SOIFO. The transfer functions and observed rotor flux of the four methods are summarized in Appendices A and B, respectively. The comparison of the four methods is further illustrated by the Bode diagram in Fig. 5, where the center frequency is \( \omega = \omega_1 = 100\pi \text{ rad/s} \), and the cut-off frequency of LPF is \( \omega_c = 100\text{rad/s} \). Meanwhile, the coefficients, \( k \), \( K_1 \), and \( K_2 \) are described as 1.414, 1.56 and 3.11 respectively, where the setting time \( t_s \) of SOIFO and second-order SOIFO are both kept as 0.018s.

Fig. 5. The Bode diagram of pure integrator, LPF, SOIFO and second-order SOIFO.
From the above analysis and Bode diagram, the proposed flux observation methods can be concluded as follows:

1. Due to the pole of pure integrator at the origin, the dc-gain is infinite. The ideally dc inputs will result in a ramp output, which leads to some saturation. It has poor ability to attenuate the harmonics.

2. The LPF provides the attenuation of dc component, and its magnitude frequency response decays at a rate of -20 dB/decade at high frequency. However, the phase shift and the amplitude attenuation of fundamental wave are generated by it.

3. The SOIFO can obtain better magnitude attenuation of harmonics, reaching ~40 dB/decade beyond center frequencies. However, its attenuation of lower frequency is 0 dB/decade, which means that it is sensitive to the dc offset.

4. The ideal dc component and harmonics attenuation capability are demonstrated by the second-order SOIFO, whose magnitude frequency response decays at rates of ~20 and ~60 dB/decade at lower and higher frequencies, respectively. With the negative magnitude response in the low-frequency and dc regions, the second-order SOIFO can eliminate dc component very well.

\[ D_1(z) = \frac{K_1 K_3 \omega^2 f_z^2}{f_z^2 + K_2 \omega f_z^2 + (2 + K_1 K_3) \omega^2 f_z^2 + K_1 \omega^3 f_z + \omega^4} \]

\[ Q_1(z) = \frac{K_1 K_3 \omega f_z}{f_z^2 + K_2 \omega f_z^2 + (2 + K_1 K_3) \omega^2 f_z^2 + K_1 \omega^3 f_z + \omega^4} \]

It defines \[ u = 2K_1 K_3 \omega^2 T_z^2 \], \[ v = 8K_1 \omega T_z \], \[ w = 4(2 + K_1 K_3) \omega^2 T_z^2 \], \[ x = 2K_1 \omega T_z^3 \], and \[ y = \omega T_z^4 \].

Substituting \[ b_0 = \frac{u}{16 + v + w + x + y} \], \[ a_1 = \frac{96 - 2w + 6y}{16 + v + w + x + y} \], \[ a_2 = \frac{-64 - 2v + 2x + 4y}{16 + v + w + x + y} \], \[ a_3 = \frac{-64 - 2v + 2x + 4y}{16 + v + w + x + y} \], and \[ a_4 = \frac{16 - v + w - x + y}{16 + v + w + x + y} \] into (22) and (23), and the final discretized forms of \[ Q_1(s) \] and \[ D_1(s) \] are given as

\[ Q_1(z) = \frac{b_0 \omega T_z (1 + 2z^{-1} - 2z^{-2} + z^{-4})}{1 + a_1 z^{-1} + a_2 z^{-2} + a_3 z^{-3} + a_4 z^{-4}} \]

\[ D_1(z) = \frac{2b_0 (1 - 2z^{-2} + z^{-4})}{1 + a_1 z^{-1} + a_2 z^{-2} + a_3 z^{-3} + a_4 z^{-4}} \]

The trapezoidal discrete implementation structure of the second-order SOIFO or SOGI is shown in Fig. 7.
switches to sensorless operation. The PMSM experimental platform is shown in Fig. 9. Main parameters of the PMSM are listed in Table II.

![Block diagram of the experimental bench](image1)

**Fig. 8.** The block diagram of the experimental bench.

![PMSM experimental platform](image2)

**Fig. 9.** The PMSM experimental platform.

### Table II

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \Phi_0 )</td>
<td>0.35 Wb</td>
</tr>
<tr>
<td>( n_N )</td>
<td>2000 rpm</td>
</tr>
<tr>
<td>( n_p )</td>
<td>3</td>
</tr>
<tr>
<td>( L )</td>
<td>5.0 mH</td>
</tr>
<tr>
<td>( R_s )</td>
<td>0.8 ( \Omega )</td>
</tr>
<tr>
<td>( T_N )</td>
<td>14 N ( \cdot ) m</td>
</tr>
<tr>
<td>( J )</td>
<td>3.78 ( \times 10^4 ) kg ( \cdot ) m²</td>
</tr>
</tbody>
</table>

To achieve the same setting time of SOIFO and second-order SOIFO, the experiential coefficients \( k \), \( K_1 \) and \( K_2 \) are taken as 1.414, 1.56 and 3.11 respectively.

#### A. Steady-State Performance

The estimated rotor flux trajectories with the four methods at a high speed (2000 rpm, 0 Nm), are shown in Figs. 10 and 11, respectively. As can be seen, the estimated rotor flux of pure integrator is deviated from the actual flux circle. Its deviation quantity increases with time, which results in some saturation in the rotor flux. The amplitude attenuation is generated by LPF. Meanwhile, the estimated rotor flux of SOIFO and second-order SOIFO are consistent with the actual rotor flux. The performance can be further illustrated by the overhead rotor flux cycles in Fig. 11, in which the black curves stand for the actual flux cycles. Compared with the flux deviation of pure integrator and the amplitude attenuation of LPF, the proposed methods can provide precise flux observation without drawbacks. Then, the estimated rotor positions are shown in Figs. 12. It should be noted that the phase shift is generated by LPF, which is shown in the rotor position estimation error, 0.4 rad. Specially, the rotor position estimation errors by SOIFO, second-order SOIFO are 0.15 rad and 0.13 rad, respectively. As shown in the picture, the flux and position estimation by second-order SOIFO are more accurate than those of SOIFO.

![Rotor flux circles](image3)

**Fig. 11.** The rotor flux circles. (a) Pure integrator, (b) LPF, (c) SOIFO, and (d) Second-order SOIFO.

Figs. 13 and 14 show the estimated rotor flux trajectories at a medium speed (400 rpm, 0 Nm). The results are similar with those of high speed, i.e. vertically eccentric rotor flux of pure
integrator, amplitude attenuation of LPF. As the distortion appears in the estimated rotor flux by the SOIFO method, the best flux estimation result is produced by the second-order SOIFO. The estimated rotor positions and their errors are given in Fig. 15. In these figures, the position estimation errors of pure integrator, LPF, SOIFO and second-order SOIFO, are 6.0 rad, 1.2 rad, 0.035 rad and 0.03 rad, respectively. It should be noted that the phase shift and amplitude attenuation of LPF are more severe at relatively low speed.

Except the performance analysis at mid- and high-speed regions, the accurate rotor flux and the position estimation of second-order SOIFO at the 1.0% rated speed (20 rpm) region are...
B. Dynamic Performance

To further verify the proposed observer, the estimations of speed variation between 400 rpm and 2000 rpm of LPF, SOIFO and second-order SOIFO are presented in Fig. 18. It should be noted that the speed can be well estimated by the three methods. However, the rotor position estimation errors vary widely, which are 1.2 rad of LPF, 0.15 rad of SOIFO, and 0.13 rad of second-order SOIFO. Specially, the rotor position estimation error of second-order SOIFO is caused by the sampling period delay, which could occupy a large proportion in a motor rotating period at high speed. Therefore, the rotor position estimation error caused by the delay is relatively large, and vice versa.

Fig. 19 shows the dynamic performance of the three methods during a step load disturbance from 0 to 10 Nm at 2000 rpm. In the figures, the rotor position estimation errors of LPF, SOIFO and second-order SOIFO are 1.3 rad, 0.25 rad and 0.20 rad, respectively. As can be seen, the oscillation of estimation error for rotor position is appeared with LPF. The phenomenon is eliminated by SOIFO and second-order SOIFO. As shown in Figs. 18 and 19, the SOIFO and second-order SOIFO can provide accurate rotor position during dynamic process.

As mentioned above, since \( s = j\omega \) is contained in the denominator of SOIFO and second-order SOIFO, it cannot operate at zero speed or be used for start-up. Thereby, the open-loop startup is adopted in this paper. The experimental result is shown in Fig. 20. The PMSM starts up in open loop and switches to sensorless operation at 20 rpm. Then the speed increases to 200 rpm, in which one step load is added, and then reduced. It can be seen that the estimated speed tracks the actual speed well at steady state and even dynamic process.

---

**Fig. 15.** The rotor position estimation at 400 rpm. (a) Pure integrator, (b) LPF, (c) SOIFO, and (d) Second-order SOIFO.

**Fig. 16.** The rotor flux trajectory at 20 rpm by second-order SOIFO.

**Fig. 17.** The position estimation at 20 rpm by second-order SOIFO.

**Fig. 18.** Experimental comparison of speed and position estimation errors during continuous speed variation without load. (a) LPF, (b) SOIFO, and (c) Second-order SOIFO.

**Fig. 19.** Experimental comparison of speed and position estimation error with load step disturbance at 2000 rpm. (a) LPF, (b) SOIFO, and (c) Second-order SOIFO.
C. DC Disturbance Suppression

To further validate the excellent dc elimination performance of second-order SOIFO, the results under dc disturbances, i.e. $\Delta u_\alpha = 5 \text{ V}$ and $\Delta i_\beta = 1.5 \text{ A}$, are given in Figs. 21 and 22, respectively.

In Fig. 21, under the $\alpha$-axis voltage disturbance, the imbalance $0.05 \text{ Wb}$ appears between the estimated $\alpha$- and $\beta$-axis rotor flux linkages by LPF. Meanwhile, the position estimation error fluctuates with $0.65 \text{ rad}$ under the dc offset.

By SOIFO, the flux estimation imbalance is reduced to $0.03 \text{ Wb}$ and the position estimation error oscillation range is $0.25 \text{ rad}$. Fortunately, the rotor flux estimation imbalance and position estimation error do not appear by the help of second-order SOIFO.

The similar situation also happens at the $\beta$-axis current disturbance, as shown in Fig. 22. The rotor flux estimation imbalances between the estimated $\alpha$- and $\beta$-axis rotor flux of LPF, SOIFO and second-order SOIFO are $0.03 \text{ Wb}$, $0.017 \text{ Wb}$ and $0.0 \text{ Wb}$, respectively.

This article has been accepted for publication in a future issue of this journal, but has not been fully edited. Content may change prior to final publication. Citation information: DOI 10.1109/TPEL.2018.2822769, IEEE Transactions on Power Electronics
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Fig. 22. Estimated rotor flux, rotor position and errors at 1000 rpm under current disturbance (ΔI = 1.5A). (a) LPF, (b) SOIFO, (c) second-order SOIFO.

and 0.005 Wb, respectively. The oscillations of position estimation error of LPF and SOIFO are 0.6 rad and 0.22 rad, respectively. Fortunately, such kind of phenomena do not appear in the second-order SOIFO.

Fig. 23 shows the comparisons of observed rotor flux and position based on LPF, SOIFO and second-order SOIFO with mismatch resistance $R_s$ and 10 Nm load. The resistance is changed from $R_s=0.4R_{so}$ to $0.4R_{so}$. Because of relatively large current, the resistance mismatch leads to some variations of estimated rotor flux. Generally, the variation will cause the rotor position estimation errors. Fortunately, with strong filtering ability of the proposed flux observer, the estimated rotor position does not fluctuate.

As shown in Figs. 21, 22 and 23, it can be further concluded that the second-order SOIFO can suppress the position estimation error ripple effectively, and offer strong ability to suppress the dc and harmonics in various operating states.

V. CONCLUSION

Two flux observers, SOIFO and second-order SOIFO are proposed in this paper for estimating the rotor flux of PMSM. Compared with the traditional rotor flux observer, the technique can reject the negative effects caused by dc offset and harmonics. Then, the accurate rotor position and speed for sensorless control are calculated by the observed rotor flux. Thereby, the application problems, i.e., parameter mismatches, voltage or current detection errors and unknown integral initial value in the conventional rotor flux observers, are eliminated. The proposed scheme is also easy to implement, and is suitable in low speed region. Its steady-state, dynamic and anti-disturbance performances are verified by comprehensive experiments.

APPENDIX

A. Summarization of the Transfer Function

<table>
<thead>
<tr>
<th>Observers</th>
<th>Transfer function</th>
</tr>
</thead>
<tbody>
<tr>
<td>Pure integrator</td>
<td>$1/s$</td>
</tr>
<tr>
<td>LPF</td>
<td>$\frac{1}{s + \omega}$</td>
</tr>
<tr>
<td>SOIFO</td>
<td>$\frac{k\omega}{s^2 + k\omega s + \omega^2}$</td>
</tr>
<tr>
<td>Second-order SOIFO</td>
<td>$\frac{K_s K_p \omega^2 s}{s^3 + K_s K_p \omega s^2 + (2 + K_s K_p) \omega^2 s + K_s \omega^3 + \omega^4}$</td>
</tr>
</tbody>
</table>

B. Summarization of the Observed Rotor Flux

C. Steady-State Rotor Flux Estimation of SOIFO

1) Analysis of dc Component
The poles are located in the left half plane, the fundamental component of \( \Psi \) with (20), it is written as
\[
\Psi = \sum_{k=1}^{n} \frac{A_k}{s_k} \sin(o_1 t + \phi_1 - 0.5\pi)
\]
where \( \gamma_1 = \tan^{-1} \left( \frac{o_2 - o_1}{k o_1} \right) \). Ultimately, the steady-state rotor flux linkage of SOIFO is described as
\[
\Psi_{r,SOIFO} = \frac{A_k}{\omega_1} + \frac{A_h}{\omega_h} \sin(o_1 t + \phi_1 - 0.5\pi)
\]
and
\[
= \sum_{k=1}^{n} \frac{A_k}{s_k} \sin(o_1 t + \phi_1 - 0.5\pi + \gamma_k)
\]

### TABLE IV
SUMMARIZATION OF THE OBSERVED ROTOR FLUX

<table>
<thead>
<tr>
<th>Observers</th>
<th>dc component</th>
<th>Fundamental component</th>
<th>High-order component</th>
</tr>
</thead>
<tbody>
<tr>
<td>Pure integrator</td>
<td>( A_I + A_i \cos(o_I) ) + ( \sum A_i \cos(o_i) )</td>
<td>( \sum A_i \sin(o_i t + \phi_i - 0.5\pi) )</td>
<td>( \sum A_i \sin(o_i t + \phi_i - 0.5\pi + \theta_i) )</td>
</tr>
<tr>
<td>LPF</td>
<td>( \sum A_i \cos(o_i + \theta_i) )</td>
<td>( \sum A_i \sqrt{o_i^2 + \omega^2} \sin(o_i t + \phi_i - 0.5\pi + 0.5\pi) )</td>
<td>( \sum A_i \sqrt{o_i^2 + \omega^2} \sin(o_i t + \phi_i - 0.5\pi + \gamma_i) )</td>
</tr>
<tr>
<td>SOIFO</td>
<td>( A_i \omega_i / \omega_h )</td>
<td>( \sum A_i \sqrt{o_i^2 + \omega^2} \sin(o_i t + \phi_i - 0.5\pi) )</td>
<td>( \sum A_i \sqrt{o_i^2 + \omega^2} \sin(o_i t + \phi_i - 0.5\pi + \gamma_i) )</td>
</tr>
<tr>
<td>Second-order SOIFO</td>
<td>0</td>
<td>( \sum A_i \sin(o_i t + \phi_i - 0.5\pi) )</td>
<td>( \sum A_i \sin(o_i t + \phi_i - 0.5\pi + \gamma_i) )</td>
</tr>
</tbody>
</table>

Combining (9) and (15), it obtains
\[
s \Psi_{r,SOIFO}(s) = s \cdot k \omega / s^2 + k \omega s + \omega^2 \cdot A_0 / s
\]
And its pole points can be calculated as
\[
\rho_{SOIFO,2} = \frac{k \omega \pm \omega \sqrt{k^2 - 4}}{2}
\]
When \( k = \sqrt{2} \), the poles are located in the left half plane of the complex frequency domain. Therefore, the system is stable. According to the final value theorem of Laplace transform, it can be obtained as
\[
\lim s \Psi_{r,SOIFO}(s) = \lim \Psi_{r,SOIFO}(s)
\]
where
\[
\gamma_1 = \tan^{-1} \left( \frac{o_2 - o_1}{k o_1} \right)
\]

D. Steady-State Rotor Flux Estimation of Second-order SOIFO

1) Analysis of dc Component
Combining (9) and (20), it obtains
\[
s \Psi_{r,SOIFO}(s) = s / s^2 + K_1 \omega^2 s + (2 + K_2) \omega^2 s + K_3 \omega^2 s + S
\]
It can be concluded that the poles are located in the left half plane of the complex frequency domain. Thus, the final value theorem of Laplace transform can be used as
\[
\lim s \Psi_{r,SOIFO}(s) = \lim \Psi_{r,SOIFO}(s)
\]
where
\[
\gamma_1 = \tan^{-1} \left( \frac{o_2 - o_1}{k o_1} \right)
\]

2) Analysis of Fundamental Component
Combining \( s = j \omega \) with (15), it is written as
\[
k \omega / s^2 + k \omega s + \omega^2 = 1 / j \omega = 1 / \omega \angle - 0.5\pi
\]
Considering the fundamental component of EMF \( A_i \sin(o_1 t + \phi_1) \) and \( o_1 = \omega_1 \), the fundamental component of estimated rotor flux of SOIFO is given as
\[
A_i \omega_1 / \omega_1 \sin(o_1 t + \phi_1 - 0.5\pi)
\]

3) Analysis of Harmonics:
Combining \( s = j \omega \) with (15), it is written as
\[
k \omega / s^2 + k \omega s + \omega^2 = 1 / j \omega = 1 / \omega \angle - 0.5\pi
\]
where \( \theta_1 = \tan^{-1} \left( \frac{k o_1 o_2}{o_2 - o_1} \right) \). Considering the harmonics of back-EMF \( \sum A_i \sin(o_i t + \phi_i) \), the high-order component of estimated rotor flux linkage of SOIFO is given as
\[
\sum \frac{1}{\sqrt{(1-h^2)^2 + h^2 k^2}} A_i k / \omega_1 \angle o_i t + \phi_i + \theta_i
\]
\[
= \sum \frac{1}{\sqrt{(1-h^2)^2 + h^2 k^2}} A_i \omega_1 / \omega_1 \sin(o_1 t + \phi_1 - 0.5\pi + \gamma_1)
\]
\[ s^4 + K_2 \omega_0^2 s^3 + (2 + K_2 \omega_0^2) \omega_0^2 s^2 + K_2 \omega_0^2 s + \omega_0^4 = j K_1 K_2 \omega_0^3 \]
\[ h \omega_0^4 - j h K_2 \omega_0^3 - h^2 (2 + K_2 \omega_0^2) \omega_0^3 + j h K_2 \omega_0^3 + \omega_0^5 \]
\[ = h K_2 - h K_2 + j [1 - h^2 (2 + K_2 \omega_0^2) \omega_0^3 + (K_2 \omega_0^2 - h \omega_0^3)] \frac{h K_2}{1 - h^2 (2 + K_2 \omega_0^2) \omega_0^3 + (h K_2 - h \omega_0^3)^2} \]
\[
\frac{h K_2}{\omega_0} \left[ \frac{1}{1 - h^2 (2 + K_2 \omega_0^2) \omega_0^3 + (h K_2 - h \omega_0^3)^2} \right]
\]
\[ = \frac{1}{\sqrt{1 - h^2 (2 + K_2 \omega_0^2) \omega_0^3 + (h K_2 - h \omega_0^3)^2}} \]
\[ \sum \sin(\omega_0 t + \phi_0) \]

where \( \theta = \tan \left( \frac{\omega_0^4 - h (2 + K_2 \omega_0^2) \omega_0^3 + (K_2 \omega_0^2 - h \omega_0^3)}{K_2 \omega_0^2 \omega_0^4 - K_2 \omega_0^2 + \omega_0^4} \right) \). The steady-state rotor flux of second-order SOIFO is described as

\[ \Psi_{T, \text{soifo}}(t) = \frac{A_1}{\omega_0} \sin(\omega_0 t + \phi_0 - 0.5 \pi) + \]
\[ \sum \frac{1}{\sqrt{1 - h^2 (2 + K_2 \omega_0^2) \omega_0^3 + (h K_2 - h \omega_0^3)^2}} \frac{A_1}{\omega_0} \sin(\omega_0 t + \phi_0 - 0.5 \pi + \gamma_{s2}) \]

REFERENCES


Wei Xu (M’09–SM’13) received the B.E. and M.E. degrees from Tianjin University, Tianjin, China, in 2002 and 2005, respectively, and the Ph.D. degree from the Chinese Academy of Sciences, Beijing, China, in 2008, all in electrical engineering. From 2008 to 2012, he held several academic positions with Australian and Japanese universities and companies. Since 2013, he has been a Full Professor with the State Key Laboratory of Advanced Electromagnetic Engineering and Technology, School of Electrical and Electronic Engineering, Huazhong University of Science and Technology, Wuhan, China. His research interests mainly include the electromagnetic design and control algorithms of linear/rotary machines, including induction, permanent magnets, switched reluctance, and other emerging novel structure machines.

Yajie Jiang received the B.S. degree from the School of Electrical Engineering, Zhengzhou University, Zhengzhou, China, in 2015. He is currently working toward the M.S. degree in the School of Electrical and Electronic Engineering, Huazhong University of Science and Technology, Wuhan. His research interests include advanced control methods for power electronics.

Chaoxu Mu (M’15) received the Ph.D. degree in control science and engineering from Southeast University, Nanjing, China, in 2012. She was a Visiting Ph.D. Student with the Royal Melbourne Institute of Technology University, Melbourne, Australia, from October 2010 to November 2011, and was a Postdoctoral Fellow with the Department of Electrical, Computer and Biomedical Engineering, University of Rhode Island, Kingston, RI, USA, from December 2014 to August 2016. She is currently an Associate Professor in the School of Electrical and Information Engineering, Tianjin University, Tianjin, China. Her current research interests include nonlinear system control and optimization, and adaptive and learning systems.

Frede Blaabjerg (S’86–M’88–SM’97–F’03) was with ABB-Scandia, Randers, Denmark, from 1987 to 1988. From 1988 to 1992, he was a Ph.D. Student with Aalborg University, Aalborg, Denmark. He became an Assistant Professor in 1992, an Associate Professor in 1996, and a Full Professor of power electronics and drives in 1998. From 2017 he became a Villum Investigator. His current research interests include power electronics and its applications such as in wind turbines, PV systems, reliability, harmonics and adjustable speed drives. He has published more than 450 journal papers in the fields of power electronics and its applications. He is the co-author of two monographs and editor of 6 books in power electronics and its applications. He has received 24 IEEE Prize Paper Awards, the IEEE PELS Distinguished Service Award in 2009, the EPE-PEMC Council Award in 2010, the IEEE William E. Newell Power Electronics Award 2014 and the Villum Kann Rasmussen Research Award 2014. He was the Editor-in-Chief of the IEEE TRANSACTIONS ON POWER ELECTRONICS from 2006 to 2012. He has been Distinguished Lecturer for the IEEE Power Electronics Society from 2005 to 2007 and for the IEEE Industry Applications Society from 2010 to 2011 as well as 2017 to 2018. He is nominated in 2014, 2015, 2016 and 2017 by Thomson Reuters to be between the most 250 cited researchers in Engineering in the world. In 2017 he became Honoris Causa at University Politehnica Timisoara (UPT), Romania.