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Sensorless Control of Standalone Brushless Doubly-Fed Induction Generator Feeding Unbalanced Loads in Ship Shaft Power Generation System

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Abstract—The standalone brushless doubly-fed induction generator (BDFIG) has demonstrated excellent energy saving performance in ship shaft power generation applications. As a standalone system, it exhibits unbalanced terminal voltages and poor performance under unbalanced loads. However, the existing control scheme of grid-connected BDFIGs cannot be directly applied to stabilize the amplitude and frequency of terminal voltage when the rotor speed and electrical load vary. This paper presents a new sensorless control scheme for the standalone BDFIG under unbalanced load conditions in the ship shaft power generation system. A second-order generalized integrator based quadrature signal generator is introduced to realize the rotor speed observer for the standalone BDFIG feeding unbalanced loads. The compensation method of negative-sequence power winding voltage is proposed to eliminate the negative-sequence component of the unbalanced power winding voltage. Comprehensive experiments are carried out on a prototype BDFIG with and without the compensation of negative-sequence power winding voltage. The good performance of the proposed sensorless control scheme is verified by the experimental test results.

Index Terms—Brushless doubly-fed induction generator (BDFIG), unbalanced load, sensorless control, ship shaft power generation, standalone operation.

I. INTRODUCTION

THE brushless doubly-fed induction generator (BDFIG) consists of a specially designed rotor and two sets of stator windings with different number of pole pairs, in which the indirect electromagnetic coupling effect between the two stator windings is generated via the rotor [1], [2]. Compared with the conventional wound rotor doubly-fed induction generator (DFIG), the BDFIG, without electric brushes and slip rings, has
the merits of longer work lifetime, higher reliability and lower maintenance cost [3].

The standalone BDFIG has exhibited excellent energy saving performance in variable speed constant frequency ship shaft power generation systems. It can be connected to the main engine of the ship directly or through a gearbox to exploit the redundant power of the main engine for power generation by employing a fractionally rated power converter [4]. The structure of the standalone BDFIG system for ship shaft power generation applications is shown in Fig. 1, where the CW is supplied by a CW side converter with the frequency variable exciting current, and an active front end rectifier with a filter is connected to the PW to regulate the dc bus voltage and to achieve the bidirectional power flow. Generally, the electrical loads of a ship include both three-phase loads and single-phase loads. If the three-phase loads fail, or the loads on the three single-phase power lines are unequal, the total three-phase load of the power generation system would be unbalanced. The unbalanced load would cause unbalanced three-phase PW current, and result in unbalanced three-phase PW terminal voltage.

The vector control schemes of the grid-connected BDFIG wind power generation system under balanced and unbalanced grid conditions have been proposed in [5]-[7] with the active and reactive power as the control objectives. Unlike the grid-connected BDFIG, the standalone BDFIG system should stabilize the amplitude and frequency of the output voltage when the rotor speed and electrical load vary. Hence, the control scheme of the grid-connected BDFIG cannot be directly applied to the standalone BDFIG.

For a standalone BDFIG feeding balanced three-phase loads, a direct voltage control scheme for standalone wind energy conversion systems has been developed in [8]. The transient control of reactive current for the active front end rectifier in the standalone BDFIG power generation system has been proposed to improve the quality of output voltage [9]. Without considering the unbalanced loads, a control scheme for the standalone BDFIG ship shaft power generation system has been designed [10] and analyzed [11], based on the CW current orientation. However, no study has been reported on the control scheme for standalone BDFIGs feeding unbalanced loads.

Many other typical BDFIG control schemes have also been proposed, such as the fault ride-through control [12], generalized vector control [13], phase angle control [14], indirect stator-quantities control [15], and direct torque control [16].

The differential operation principle and dynamic vector model of BDFIG

A. Basic Operation Principle

The BDFIG can be operated in various modes, including the doubly-fed, cascaded, and induction modes [3]. Among them, the doubly-fed mode is the optimal one, in which the rotor speed can be expressed as

\[ \omega_2 = \omega_f \cos \phi \]

(1)

When \( \omega_2 \) is zero, the rotor is rotating at the natural synchronous speed \( \omega_s \). The rotor speed above \( \omega_s \) is called the
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sub-synchronous speed. When the rotor speed varies, in order to reduce when the rotor speed increases under constant load.

The proposed control scheme just requires the accurate rotor speed information, but not necessarily the real rotor position. When ignoring the integration constant, the following relationship can be derived for the rotor position by integrating both sides of (1):

\[ \theta_r = \theta + \theta_1 \quad (6) \]

where \( \theta_r \) is the virtual rotor position, which is not the accurate rotor position by the general definition.

Therefore, from (6), the difference between the actual and estimated virtual rotor positions, \( \Delta \theta_r \), can be derived as

\[ \Delta \theta_r = (p_1 + p_2) (\theta - \hat{\theta}) \quad (7) \]

When the system is working in the quasi-locked state (i.e., \( \omega = \omega_r = \hat{\theta}_r \)), the value of (7) will be sufficiently small, and \( (p_1 + p_2) \Delta \theta_r \) can be approximated as \( \sin(\theta + \hat{\theta}) - (p_1 + p_2) \hat{\theta}_r \). Hence, based on the principle of the stationary-frame phase-locked loop (PLL) [21], the basic RSO as shown in Fig. 2(a) can be employed to obtain the rotor speed [22], in which the virtual rotor position \( \theta_r \) is just an intermediate variable.

In the standalone BDFIG system, since the PW voltage is generated by injecting an exciting current in the CW, the CW current is not zero even under the no-load condition [17]. In addition, once the CW current is not zero, the PW voltage will be generated. So, there can be no case of the zero voltage of the PW side and zero current of the CW side. Consequently, \( \sin \theta_1 \) and \( \cos \theta_1 \) can be obtained from the three-phase PW voltage, and \( \sin \theta_2 \) and \( \cos \theta_2 \) from the three-phase CW current, as shown in Fig. 2(a). Hence, the basic RSO can observe the rotor speed of the standalone BDFIG based on the three-phase PW voltage and the three-phase CW current. The parameter tuning method of the PI controller in Fig. 2(a) can refer to [22].

However, an unbalanced load would result in an unbalanced three-phase PW current, which could produce different voltage drops across the three-phase PW impedances, resulting in unbalanced PW terminal voltages. The negative-sequence PW current would induce abundant harmonic current in the CW due to the indirect coupling between the PW and CW via the rotor. When using the basic RSO, the rotor speed cannot be accurately estimated by the unbalanced PW voltage and distorted CW current.

To solve this problem, the second-order generalized integrator-based quadrature-signal generator (SOGI-QSG) [23] and the positive-sequence calculator [6] are introduced to construct the improved RSO for the standalone BDFIG under unbalanced load conditions, as shown in Fig. 2(b). The SOGI-QSGs 1 and 2 are used to obtain the filtered results (\( u_{10f} \) and \( u_{20f} \)) and 90° phase-shifted versions (\( q_{10f} \) and \( q_{20f} \)) of the \( \alpha \)- and \( \beta \)-components of the PW voltage. The \( \alpha \)- and \( \beta \)-components of the positive-sequence PW voltage, \( u_{10f} \), \( u_{20f} \), are derived from the positive-sequence calculator and sent to the stationary-frame PLL to make the SOGI-QSGs 1 and 2 frequency-adaptive. The SOGI-QSGs 3 and 4 are employed to filter the \( \alpha \)- and \( \beta \)-components of the CW current, where the resonance frequency is derived from (2). Finally, these quantities \( u_{10f} \), \( u_{20f} \), \( i_{2af} \) and \( i_{2bf} \) are input to the basic
This page contains a diagram and text discussing the control scheme for a standalone BDFIG feeding unbalanced loads.

**IV. CONTROL SCHEME DEVELOPMENT**

**A. Overall Sensorless Control Scheme**

According to the instantaneous symmetrical component method, the unbalanced three-phase voltage can be decomposed into balanced positive-, negative- and zero-sequence components. In a ship shaft power generation system, the BDFIG is connected to the loads through the three-phase three-wire system, which has no neutral line, or the path for the zero-sequence current components. Hence, only the negative-sequence component of the output voltage need to be compensated in the proposed control scheme.

The overall sensorless control scheme, as shown in Fig. 3, consists of the positive-sequence PW voltage controller, the negative-sequence PW voltage controller, the improved RSO, and the dual second-order generalized integrator PLL (DSOGL-PLL). The improved RSO is employed to estimate the rotor speed, which would be used in the control of the PW frequency and the calculation of the feed-forward compensation in the current loops of the positive-sequence PW voltage controller and the negative-sequence PW voltage compensator.

The DSOGL-PLL can be employed to estimate the amplitudes, frequencies and phase angles of the positive- and negative-sequence PW voltages, which is with good steady-state and dynamic performance when the three-phase voltages contain unbalanced drops [23].

**B. Positive-SequenCe PW Voltage Controller**

According to (2), in order to generate the positive-sequence PW voltage, the CW current frequency in the positive-sequence PW voltage controller, \( \omega_1^d \), should be specified by

\[
\omega_1^d = \omega_1 (p_1 + p_2) - \omega_0. \tag{8}
\]

The proposed positive-sequence PW voltage controller is based on the CW current vector orientation, i.e. substituting \( \omega_0 \) in (5) with \( \omega_1^d \), and the dynamic model of the BDFIG can be rewritten as

\[
\begin{align*}
U_{dq_{dq}}^+ &= R_1 I_{d_{dq}}^+ + s \Psi_{d_{dq}}^+ + j \omega_0 \Psi_{d_{dq}}^+, \\
\Psi_{d_{dq}}^+ &= L_1 I_{d_{dq}}^+ + L_s I_{d_{dq}}^+, \\
U_{dq_{dq}}^- &= R_1 I_{d_{dq}}^- + s \Psi_{d_{dq}}^- + j (\omega_0 - (p_1 + p_2) \omega_1) \Psi_{d_{dq}}^-, \\
\Psi_{d_{dq}}^- &= L_1 I_{d_{dq}}^- + L_s I_{d_{dq}}^-.
\end{align*}
\tag{9}
\]

Rearranging (8), one obtains \( \omega_0 = -p_1 \omega_1. \) By substituting this equation into the \( 5^\text{th} \) equation of (9) and setting the rotor voltage to zero, the \( 5^\text{th} \) and \( 6^\text{th} \) equations of (9) in the form of \( d \)- and \( q \)-components can be expressed as

\[
\begin{align*}
\psi_{d_{dq}}^+ &= R_1 i_{d_{dq}}^+ + \Psi_{d_{dq}}^+ + (\alpha_p - p_2 \omega_1) \Psi_{d_{dq}}^- \\
\psi_{q_{dq}}^+ &= L_1 i_{d_{dq}}^+ + L_s i_{d_{dq}}^+ + \psi_{d_{dq}}^+ + (\alpha_p - p_2 \omega_1) \Psi_{d_{dq}}^- \\
\psi_{d_{dq}}^- &= L_1 i_{d_{dq}}^- + L_s i_{d_{dq}}^- + \psi_{d_{dq}}^- + (\alpha_p - p_2 \omega_1) \Psi_{d_{dq}}^+ \\
\psi_{q_{dq}}^- &= L_1 i_{d_{dq}}^- + L_s i_{d_{dq}}^- + \psi_{d_{dq}}^- + (\alpha_p - p_2 \omega_1) \Psi_{d_{dq}}^+.
\end{align*}
\tag{10-13}
\]
Substituting (12) and (13) into (10) and (11) yields
\[ i_{rd}^* = \frac{L_2 s^2 + R_2 s + L_1 (\alpha - p_1 \omega) y}{(R_r + L_s) s^2 + L_1 (\alpha - p_1 \omega) y} \left[ L_1 i_{rd}^* + L_2 i_{rd}' \right] \]
\[ R_r (\alpha - p_1 \omega) (L_1 i_{rd}^* + L_2 i_{rd}') \]
\[ \frac{R_r (\alpha - p_1 \omega) L_1 (\alpha - p_1 \omega) y}{(R_r + L_s) s^2 + L_1 (\alpha - p_1 \omega) y} \]
\[ i_{rd}' = -\frac{a_1 - p_1 \alpha}{R_r} \left[ L_1 i_{rd}^* + L_2 i_{rd}' \right] \]
\[ \frac{R_r (\alpha - p_1 \omega) L_1 (\alpha - p_1 \omega) y}{(R_r + L_s) s^2 + L_1 (\alpha - p_1 \omega) y} \left[ L_1 i_{rd}^* + L_2 i_{rd}' \right] \]
\[ \frac{a_1 - p_1 \alpha}{R_r} \left[ L_1 i_{rd}^* + L_2 i_{rd}' \right] \]  
(14)

In order to employ a fractionally rated power converter to drive the standalone BDFIG system, the CW angular frequency \( \omega_2 \) is usually limited to 30\% of the PW frequency, and thus the rotor speed range is limited according to (1). Consequently, the value of \( (a_1 - p_1 \alpha) ^2 \) will be much greater than 1. Since in general the values of \( R_r \) and \( L_r \) are close, \( R_r ^2 \) is much smaller than \( L_1 (\alpha - p_1 \omega) ^2 \), and \( R_r ^2 \) in the denominator of the first term of (14) can be ignored. Therefore, the first term of (14) can be approximated as
\[ \left[ s^2 + (R_r / L_s) s + (\alpha - p_1 \omega) y \right] \left[ L_1 i_{rd}^* + L_2 i_{rd}' \right] \]
Since the zeros and poles of this term are very close such that they can cancel each other, the first term of (14) can be simplified as \( -(L_1 i_{rd}^* + L_2 i_{rd}') / L_r \).

In the steady state, the \( (R_r + L_s) s \) in the denominator of the second term of (14) would be much smaller than \( L_2 (\alpha - p_1 \omega) ^2 \) and can be ignored. Consequently, the second term in (14) can be simplified as
\[ \left[ R_r (L_1 i_{rd}^* + L_2 i_{rd}') \right] \left[ L_1 (\alpha - p_1 \omega) y \right] \]
Therefore, (14) can be simplified as
\[ i_{rd}^* = -L_1 i_{rd}^* + L_2 i_{rd}' \]
\[ \frac{L_1 i_{rd}^* + L_2 i_{rd}'}{L_r} \]  
(16)

Similarly, (15) can be simplified as
\[ i_{dq}^* = -(L_1 i_{dq}^* + L_2 i_{dq}') / L_r \]
(17)

Since \( a_1 - (p_1 + p_2) \omega = \alpha \), splitting the 3rd and 4th equations of (9) into the \( d \)- and \( q \)-components results in
\[ u_{rd}^* = R_r i_{rd}^* + \alpha \psi_{d2}^* \]
\[ u_{dq}^* = R_r i_{dq}^* + \alpha \psi_{d2}^* \]
\[ \psi_{d2}^* = L_1 i_{rd}^* + L_2 i_{rd}' \]
\[ \psi_{dq}^* = L_1 i_{dq}^* + L_2 i_{dq}' \]
(18)

(19)

(20)

(21)

Substituting (16) and (17) in (18)-(21) yields
\[ u_{rd}^* = K_{rd} i_{rd}^* + D_{rd} \]
\[ u_{dq}^* = K_{dq} i_{dq}^* + D_{dq} \]
(22)

(23)

where \( K_{rd}^* = K_{rd} \), \( D_{rd} = \alpha \), \( D_{dq} = \alpha \), and \( K_{dq} = 1 - L_1 (L_r / L_r) \) is the leakage constant of CW. The detailed mathematical expressions for \( \alpha_1 \), \( \alpha_2 \), \( \alpha_3 \), \( \alpha_4 \), and \( \alpha_5 \) are given in Appendix. \( K_{rd}^* \) and \( K_{dq}^* \) represent the direct relationship between the \( d \)- and \( q \)-components of \( I_{rd}^* \) and \( I_{dq}^* \), respectively, and \( D_{rd}^* \) and \( D_{dq}^* \) the cross coupling between the PW and the CW.

In the steady state, the PW flux linkage can be regarded as a constant, and thus, the 1st equation of (9) can be simplified as
\[ \frac{U_{rd}^*}{L_r} = R_r i_{rd}^* + j \omega \psi_{d2}^* \]
(24)

By splitting (24) and the 1st equation of (9) into the \( d \)- and \( q \)-components, the following expressions can be obtained
\[ u_{rd}^* = R_r i_{rd}^* - \alpha \psi_{d2}^* \]
\[ u_{dq}^* = R_r i_{dq}^* + \alpha \psi_{d2}^* \]
\[ \psi_{d2}^* = L_1 i_{rd}^* + L_2 i_{rd}' \]
\[ \psi_{dq}^* = L_1 i_{dq}^* + L_2 i_{dq}' \]
(25)

(26)

(27)

(28)

By combining (16), (17), (25)-(28), if \( i_{rd}^* = 0 \) and \( i_{dq}^* = I_{dq}^* \), \( u_{rd}^* \) and \( u_{dq}^* \) can be expressed as
\[ u_{rd}^* = R_r i_{rd}^* + \frac{a_1 (L_1 - L_2) i_{dq}^*}{L_r} \]
\[ u_{dq}^* = -\frac{a_1 (L_1 - L_2) i_{dq}^*}{L_r} + R_r i_{dq}^* + \frac{a_1 L_2}{L_r} - \frac{a_1 L_1}{L_r} \]
(29)

(30)

The amplitude of the positive-sequence PW voltage can be expressed as
\[ U_{rd}^* = \alpha \psi_{d2}^* \]
\[ U_{dq}^* = \alpha \psi_{d2}^* \]
\[ \psi_{d2}^* = \alpha \psi_{d2}^* \]
\[ \psi_{dq}^* = \alpha \psi_{d2}^* \]
(31)

(32)

For a small variation in the control variable \( \Delta I_{dq}^* \), (31) can be approximated by its Taylor expansion as
\[ U_{rd}^* (I_{dq}^* + \Delta I_{dq}^*) = U_{rd}^* (I_{dq}^*) + \Delta I_{dq}^* \frac{dU_{rd}^*}{dI_{dq}^*} |_{I_{dq}^*} \]
(33)

where subscript \( S \) indicates the steady-state value.

From (32), the relationship between the amplitude increments of the positive-sequence PW voltage and the positive-sequence CW current can be expressed as
\[ \Delta U_{rd}^* = K_{rd}^* \Delta I_{dq}^* \]
(34)

\( I_{dq}^* \) is the amplitude of steady-state positive-sequence CW current.

The amplitude of steady-state positive-sequence PW voltage under \( I_{dq}^* \) is approximately equal to the amplitude reference of
positive-sequence PW voltage, \( U_{1}^{*} \). Therefore, according to (31), \( I_{d}^{*} \) can be derived as

\[
I_{d}^{*} = \beta_{d}\bar{I}_{d} + \frac{[2\beta_{d}R_{d} \bar{I}_{d}^{2} - R_{d}^{2}(\bar{I}_{d})^{2} - (\beta_{d})^{2}(\bar{I}_{d})^{2} + U_{1}^{*}]^{2}}{\beta_{d}^{2}} (35)
\]

where the detailed mathematical expressions for \( \beta_{d} \) and \( \beta_{i} \) are given in Appendix.

According to (22), (23), and (33)-(35), a cascaded controller for the positive-sequence PW voltage can be designed, as shown in Fig. 3. The controller utilizes a CW current vector controller as the inner loop, and a PW voltage amplitude controller and a frequency controller as the outer loops.

C. Compensation of Negative-Sequence PW Voltage

According to (2), in order to compensate the negative-sequence PW voltage, the CW current frequency should be given by

\[
\omega_{c} = \omega_{r}(p_{1} + p_{2}) + \omega_{c} \quad (36)
\]

From (8) and (36), the relationship between \( \omega_{c} \) and \( \omega_{1} \) can be obtained by

\[
\omega_{c} = \omega_{1} + 2\omega_{c} \quad (37)
\]

As it can be seen from (37), the difference between the frequencies of the CW fundamental current and compensation current is twice the PW frequency.

Based on the CW current vector orientation, the angular speed of the \( dq \) rotation reference frame \( \omega_{d} \) in the dynamic vector model should be set as \( \omega_{p} \). Equation (5) can then be rewritten as

\[
\begin{align*}
[ \Phi_{1}^{*} &= R_{d}I_{d} + s\Phi_{1}^{*} + jo_{q} \Phi_{1}^{*} \\
\Phi_{q}^{*} &= L_{r}\Phi_{q}^{*} + L_{d}I_{d} + j\Phi_{q}^{*} \\
\end{align*}
\]

Employing a similar method of deriving (22) and (23), one can obtain the following relationships

\[
\begin{align*}
u_{di} &= K_{d}^{-1}\dot{i}_{d} + D_{d} \\
u_{qi} &= K_{i}^{-1}\dot{i}_{q} + D_{i}
\end{align*}
\]

where

\[
k_{d} = k_{d} + \sigma_{l}I_{ci} \quad D_{d} = c_{d}^{*}i_{ci} + c_{d}i_{ci} + c_{d}^{*}i_{ci} \quad \text{and} \\
k_{i} = k_{i} + \sigma_{l}I_{ci} \quad D_{i} = c_{i}^{*}i_{ci} + c_{i}i_{ci} + c_{i}^{*}i_{ci} \quad \text{The detailed expressions of} \quad c_{d}, \quad c_{d}^{*}, \quad c_{d}, \quad c_{d}^{*}, \quad c_{i}, \quad \text{and} \quad c_{i} \quad \text{are given in Appendix.}
\]

Similar to (33), the relationship between the amplitude increments of the negative-sequence PW voltage and the negative-sequence CW current can be derived as

\[
\Delta U_{1}^{*} = K_{d}^{*}\Delta I_{d} \quad (41)
\]

where

\[
k_{d} = c_{d}^{*}L_{di} \quad I_{ci} \quad \text{and} \quad I_{ci} \quad \text{are performed on a prototype BDFIG, whose detailed parameters are listed in Table I. In this experimental setup, a 37 kW three-phase induction motor driven by a Siemens MM430 inverter is mechanically coupled to the BDFIG and used as the prime mover. The CW is supplied by the control side converter connected with a commercial active front end through dc bus capacitors.}

B. Performance Test of Proposed RSO

In this part, an experiment is carried out to test the performance of the proposed RSO when the rotor speed and load vary. During the experiment, the standalone BDFIG system is controlled without compensation of the negative-sequence PW voltage. Since these experimental results are just used to validate whether the RSO can estimate the rotor speed with accuracy similar to that from an encoder under unbalanced load conditions, the observed results are not fed back to the controller.

The initial load is an unbalanced three-phase resistive load with resistances of 12 \( \Omega \), 12 \( \Omega \), and 6 \( \Omega \) in phases \( a \), \( b \), and \( c \), respectively. The rotor speed of the BDFIG increases from 550
rpm to 900 rpm with a rate of 300 rpm per second from 1.7 s to 2.9 s, and then, another unbalanced three-phase resistive load with resistances of 20 Ω, 20 Ω, and 30 Ω in phases a, b, and c is added to the system at 3.08 s.

Fig. 5(a) presents the PW voltage and CW current during the whole process. Fig. 5(b) shows the details of the PW voltage and CW current at around 3.08 s, when the second unbalanced load is applied. From Figs. 5(a) and (b), it can be seen that the PW voltage is unbalanced and the CW current contains abundant harmonics caused by the unbalanced loads. As shown in Figs. 5(d) and (f), at the end of the rotor speed increase, the error of rotor speed observed by the basic RSO reaches -19.3 rpm.

Therefore, during the rapid rotor speed change, although the improved RSO includes a ripple of frequency 100 Hz and peak-to-peak value 30 rpm, as shown in Figs. 5(c) and (d). The ripple of the observed rotor speed can be significantly reduced to about 12 rpm by the improved RSO, as shown in Figs. 5(e) and (f). Therefore, during the rapid rotor speed change, although the improved RSO has a slightly poorer dynamic performance than

Fig. 6. Experimental results under an unbalanced three-phase load without compensation of the negative-sequence PW voltage, where (a) PW line voltage, (b) CW phase current, and (c) UF of PW voltage.

Fig. 7. Experimental results under a single-phase load without compensation of the negative-sequence PW voltage, where (a) PW line voltage, (b) CW phase current, and (c) UF of PW voltage.
the basic one, it is with a much smaller steady state error, which is very important for the control of standalone BDFIG. When another unbalanced three-phase load is added at 3.08 s, the error of the rotor speed observed by the improved RSO is almost zero under the 17% transient drop of the PW voltage amplitude and the 30% increase of the CW current amplitude as shown in Figs. 5(a) and (b), which demonstrates good dynamic performance of the improved RSO during the load variation.

C. Results without Compensation of Negative-Sequence PW Voltage

To clearly illustrate the performance of the proposed control scheme, the system characteristics under unbalanced three-phase load and single-phase load without compensation of the negative-sequence PW voltage are presented in Experiments 1 and 2. The single-phase load is a special case of the unbalanced three-phase load, which usually causes more serious unbalance of the output voltage. In all the experiments, the amplitude and frequency references of the PW line voltage are set as 380 V and 50 Hz, respectively. The experimental results are presented in Figs. 6 and 7.

Experiment 1: The rotor speed of the BDFIG is kept constant at 885 rpm. The initial load is a balanced three-phase resistive load with resistance of 25 Ω in each phase, and an unbalanced three-phase resistive load with resistances of 12 Ω, 12 Ω, and 6 Ω in phases a, b, and c is added to the system at 0.78 s. As shown in Figs. 6(a) and (b), the unbalanced load causes the PW voltage unbalanced with an unbalance factor (UF) of 12%, and distorts the CW current through the coupling of the BDFIG rotor, as shown in Fig. 6(c).

Experiment 2: The rotor speed is kept constant at 555 rpm. The BDFIG is initially operated at no load, and at 0.94 s, a single-phase load with resistance of 12 Ω is connected between phases a and b of the PW, which causes the UF of PW voltage increase dramatically to 32% and the distortion of CW current is much severer than that in Experiment 1, as shown in Fig. 7.

D. Results with Compensation of Negative-Sequence PW Voltage

With the compensation of negative-sequence PW voltage, Experiments 3 and 4 are carried out to verify the performance of the proposed control scheme under variable loads and rotor speeds when supplying unbalanced three-phase and single-phase loads. The experimental results are shown in Figs. 8 and 9.

Experiment 3: The rotor speed is initially 885 rpm, and from 3.63 s to 4.50 s, it drops to 690 rpm at a rate of 225 rpm per second. The load change occurs at 1.65 s, and the balanced and unbalanced three-phase loads used in this experiment are the same as those in Experiment 1. The overall PW line voltage waveforms are presented in Fig. 8(a). Fig. 8(b) illustrates the overall CW phase current waveforms, in which a smooth change of the CW current phase sequence across the natural synchronous speed (750 rpm) of the BDFIG can be seen at about 4 s. As shown in Fig. 8(c), the UF of the PW voltage under the unbalanced load can be reduced to about 5%, which is
much less than that without compensation of the negative-sequence PW voltage as shown in Fig. 6(c). During the rapid rotor speed variation between 3.63 s and 4.5 s, the UF of PW voltage is not significantly changed. The PW active power during the whole process is shown in Fig. 8(d). The initial PW active power under the balanced load is 5.8 kW. After the unbalanced three-phase load is applied, the PW active power rises rapidly to 12.3 kW. As the PW voltage converges to its reference value, the PW active power gradually increases to 17.5 kW. During the rotor speed drop, the PW active power rises to 24.1 kW, which is consistent with the relationship between the PW active power and rotor speed as shown in (4).

Figs. 9(e), (f) and (g) show the details of the PW line voltage, PW phase current and CW phase current after the compensation, respectively. Although the negative-sequence PW voltage can be successfully compensated, the current becomes unbalanced, as shown in Fig. 8(f), due to the effects of the unbalanced three-phase load. As shown in Fig. 8(g), in order to compensate the negative-sequence PW voltage, a harmonic component is injected into the CW current. Fig. 8(h) presents the harmonic spectrum and total harmonic distortion (THD) of the CW current as shown in Fig. 8(g). The fundamental frequency of the CW current is 9 Hz, which is determined by the rotor speed according to (2). The harmonic has the frequency of 109 Hz, which coincides with (37), and the THD of the CW current reaches 48%.

**Experiment 4:** The standalone BDFIG is initially operated under no load, and at 2.48 s, the same single-phase load as that in Experiment 2 is connected to the system. Before 6.3 s, the rotor speed is 555 rpm, and then it rises to 680 rpm with a rate of 180 rpm per second. After 7 s, the rotor speed is kept constant at 680 rpm. The overall PW line voltage and CW phase current waveforms are illustrated in Figs. 9(a) and (b), respectively. Fig. 9(c) presents the UF of PW voltage under the single-phase load. When the single-phase load is connected to the system, the UF of PW voltage is up to about 24%, and then reduced to 4% within 1.2 s. During the rotor speed variation, the UF of PW voltage is almost constant. After the single-phase load is connected to the system, the PW active power gradually increases to about 16.2 kW as the PW voltage converges to its reference value. When the rotor speed increases to 680 rpm, the PW active power reduces to 13.2 kW, which can be explained by the relationship between the PW active power and rotor speed as shown in (4).

The expanded views of the PW line voltage, PW phase current and CW phase current after the compensation are shown in Figs. 9(e), (f) and (g), respectively. Although the negative-sequence PW voltage is almost fully compensated, due to the effect of single-phase load, two-phase PW currents are in the opposite directions, and the third phase current is nearly zero, as shown in Fig. 9(f). Since a single-phase load is a more severely unbalanced load, more harmonics need to be injected into the CW current in order to compensate the negative-sequence PW voltage, as shown in Fig. 9(g). The harmonic spectrum and THD of the CW current between 5 s and 5.5 s are shown in Fig. 9(h). The CW current contains a...
fundamental component with the frequency of 13 Hz, and a significant harmonic component with the frequency of 87 Hz. Because the BDFIG operates at a sub-synchronous speed of 555 rpm, the fundamental frequency of the CW current is 13 Hz (the negative frequency indicating that the phase sequence of the CW current is opposite to that of the CW current at a super-synchronous speed), which shows that the difference between the fundamental and harmonic frequencies of the CW current still matches (37). The amplitude of the CW current harmonic component is greater than that of the fundamental, which results in the THD of CW current being up to 134%.

VI. CONCLUSION

This paper presents a RSO and a sensorless control scheme for the standalone BDFIG feeding unbalanced loads in a ship shaft power generation system. The positive- and negative-sequence components of the unbalanced PW voltage caused by the unbalanced loads are regulated separately in the proposed control scheme. A RSO is designed for the unbalanced operation of the standalone BDFIG, which is independent to machine parameters except the number of pole pairs. Based on the proposed RSO, a sensorless control scheme for the standalone BDFIG feeding unbalanced loads is presented. Comprehensive experiments have been carried out on a BDFIG prototype with and without compensation of the negative-sequence PW voltage. It is confirmed by the experimental results that the proposed control scheme can significantly reduce the unbalance of the output voltage and thus improve the output voltage quality under unbalanced loads. Furthermore, the harmonic elimination method of standalone BDFIG ship shaft power generation systems with nonlinear loads would be another important work in the future.

APPENDIX

\[ \alpha' = \frac{a_0 (a_0 - p_0 \omega)L_2 (L_3 + L_4) - L_4 R_s}{L_0 (a_0 - p_0 \omega)} \]  

\[ \alpha'' = \frac{-L_0 L_2 s}{L_0} \]  

\[ \alpha' = \frac{-\sigma L_2 L_0 \omega}{L} \]  

\[ \alpha'' = \frac{-L_0 \omega}{L} \]  

\[ \beta' = \frac{\omega_1 L_1 - \omega_0 L_0}{L_1} \]  

\[ \beta'' = \frac{\omega_1 L_1}{L_1} \]  

\[ \alpha' = \frac{-a_0 (a_0 + p_0 \omega) L_2 (L_3 + L_4) + L_4 R_s}{L_0 (a_0 + p_0 \omega)} \]  

\[ \alpha'' = \frac{-L_0 L_2 s}{L_0} \]  

\[ \alpha' = \frac{a_0 (a_0 - p_0 \omega)L_2 (L_3 + L_4) - L_4 R_s}{L_0 (a_0 - p_0 \omega)} \]  

\[ \alpha'' = \frac{-L_0 L_2 s}{L_0} \]  

\[ \alpha' = \frac{-\sigma L_2 L_0 \omega}{L} \]  

\[ \alpha'' = \frac{-L_0 \omega}{L} \]  

\[ \beta' = \frac{\omega_1 L_1 - \omega_0 L_0}{L_1} \]  

\[ \beta'' = \frac{\omega_1 L_1}{L_1} \]

REFERENCES


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