Control of Wound Field Synchronous Motor Integrated with ZSI

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This paper proposes a novel variable speed motor drive, a wound field synchronous motor (WFSM) drive integrated with a Z-source inverter (ZSI). The proposed WFSM is a nonpermanent magnet machine and employs two field coils as dc field mmf sources. On the other hand, the ZSI has an impedance network that consists of two dc reactors and two capacitors. In the proposed drive, the two field coils of the WFSM act not only as dc field mmf sources but also as dc reactors in the ZSI, resulting in size and cost reductions in the drive system. In order to achieve voltage boost-up and field current controls simultaneously, the control algorithm for this integrated system is examined. Experimental results using a test drive system show that the proposed drive under the proposed control works properly.

Keywords: boost-up/field current control, integrated motor and drive circuit system, wound field synchronous motor, z-source inverter

1. Introduction

Currently, applications of motors using rare-earth permanent magnets increasing with the aim of increasing efficiency, while reducing size and weight. However, supply constraints and price instability of rare-earth materials have become a problem in recent years. Therefore, research and development of less- or non-rare-earth permanent magnet motors is receiving much attention from industries.

In response to this demand, we proposed a wound field synchronous motor (WFSM) as a non-rare-earth-permanent magnet motor⁽¹⁾. In a WFSM, field fluxes are generated by two field toroidal coils that are wound on stationary field pole cores located at both ends of the motor in the axial direction. This structure enables it to feed current into the field coils without any brushes and slip rings. However, the motor size and cost increase because the field pole cores and the field coils are net additional components that are not present in conventional motors.

On the other hand, in the case of a variable speed drive system where a motor is designed as a high-speed machine under a low-voltage battery drive, a DC-DC boost-up converter inserted between a battery and an inverter is often employed⁽²⁾. In this case, the size and weight of the motor are expected to reduce while improving its efficiency. However, it is a requirement in the system design to take into account efficiency degradation and cost increase owing to the addition of a DC-DC converter.

In this paper, a variable speed drive system shown in Fig. 1

Fig. 1. Proposed drive system configuration

is proposed, where a WFSM is integrated with a Z-source inverter (ZSI)⁽³⁾. The ZSI has an impedance-network, which consists of two DC reactors L_f and two capacitors C and is inserted between a DC voltage source and a three-phase inverter. The ZSI is functioning not only as a three-phase inverter, but also as a DC-DC boost-up converter. However, the size and weight of the bulky DC reactors are disadvantages of the ZSI. This integrated system can solve the drawbacks of the WFSM and the ZSI simultaneously. Specifically, two DC reactors of the ZSI are replaced by two DC field coils of the WFSM as the sharing component in the motor and converter and hence, the proposed drive system is expected to be a solution for size, weight and cost reductions as well as for loss minimization of the total drive system. In the proposed drive system, however, a difficulty lies in its control since the voltage boost-up and field current regulation required for a given operating condition must be satisfied harmoniously. Therefore, the control algorithm for the proposed system is examined as a main objective in this paper. Some experimental studies using a test drive system are conducted and verify that the proposed control algorithm can operate the proposed drive system properly.

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2. Basic Working Principle of WFSM under Proposed Integrated Drive System

2.1 Structure and Basic Working Principle of WFSM

Figure 2 illustrates a general view of a 20-pole/24-slot WFSM employing fractional-slot concentrated windings. The main machine part in the middle of the figure is composed of a laminated rotor core with 10 salient poles and a laminated stator core with three-phase armature concentrated windings accommodated in 24 slots. The back yoke core is made of soft magnetic composites (SMC) and forms axial field flux paths flowing into the field pole SMC cores located at both ends of the motor. Each field pole SMC core located at each end of the motor in the axial direction has the toroidal field coil. Figure 3 depicts the field flux paths when only the field coils are energized by a certain DC current. Although the airgap flux density is unipolar, it varies with the rotor position displacement owing to the doubly salient-pole machine nature. As the spatial distribution of gap permeance with respect to the rotor position has AC components, three-phase symmetrical AC induced voltage waveforms arise in the three-phase symmetric armature windings by the modulation of the field flux with respect to the rotor position displacement. As a result, the WFSM generates torque by feeding three-phase armature currents with an adequate phase angle against the three-phase induced voltage.

2.2 WFSM Model Given that the AC component of the airgap flux density distribution is sinusoidal, a d - q axis model can be introduced, similar to that in conventional AC motors under three-phase symmetrical sinusoidal current drive.

Assuming that the direction of the field current i_f shown in

Fig. 2. General view of WFSM

Fig. 3. Field flux paths in WFSM

Fig. 3 is positive and the origin of the rotor position is defined as one where the flux linkage in the U-phase winding reaches its positive maximum, the voltage equation of the WFSM can be expressed in the following manner through the conventional d - q transformation.

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \begin{bmatrix} R + pL_d & -\omega L_q \\ \omega L_d & R + pL_q \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \sqrt{\frac{3}{2}} \omega K_{if} \begin{bmatrix} 0 \\ 1 \end{bmatrix} \dots \dots (1)$$

The first term in the expression follows the custom in interior permanent magnet motors. The second term K_{if} in the expression is the amplitude of the AC component of the flux linkage in each phase winding generated by field coil mmf.

The torque equation can be obtained in the following manner:

$$\tau = \sqrt{\frac{3}{2}} P K_{if} i_q + P(L_d - L_q) i_d i_q \dots \dots \dots (2)$$

The first term expresses the torque produced by the field flux and the q -axis current similar to the magnet torque in permanent magnet motors. The second term is the reluctance torque. The WFSM shown in Figs. 2 and 3 employs fractional concentrated windings, where the reluctance torque expressed in the second term is negligibly smaller than the torque in the first term. Note that the polarity of the torque is determined by the product of the polarities of the field current i_f and the q -axis current i_q .

2.3 Motoring and Regenerating Control of WFSM Integrated with ZSI When the rotation direction of the motor is positive ($\omega > 0$), the WFSM integrated with the ZSI shown in Fig. 1 works in the motoring mode. In this mode, the field current i_f flows in a direction from the DC power source to the WFSM through the switch S_w . Given that the direction of this field current is positive ($i_f > 0$), the WFSM, under the motoring mode, generates positive torque by controlling the q -axis current so as to be positive based on (2). In the regenerating mode of the WFSM, the field current i_f flows in a direction from the WFSM to the DC power source through the switch S_w . In this case, the direction of this field current is negative ($i_f < 0$) and the WFSM, under the regenerating mode, produces negative torque by controlling the q -axis current to be positive based on (2). Thus, the motoring and regenerating modes in the integrated system can be switched by controlling the combination of polarities of both the field and the q -axis currents and OFF/ON of the switch S_w .

3. Operating Modes of ZSI

The operating modes of the ZSI are divided into three states in accordance with the switching pattern of the three-phase PWM inverter and OFF or ON of the switch S_w shown in Fig. 1 for the motoring and regenerating modes, respectively. Here after, it is noted that the inductance of each field coil is L_f , the resistance of each field coil is R_f , the capacitance of the capacitors is C , each voltage across each impedance-network component is $L_f di_f/dt$, $R_f i_f$ and v_c , respectively. Furthermore, the voltage and the current of the DC source are V_0 and i_{in} , respectively, and the field current is i_f , the capacitor current is i_c , the voltage across the inverter leg is v_i , the current flowing into the inverter is i_i ,

Fig. 4. Equivalent circuit at *active state* (motoring)

respectively. Each operating mode can be expressed as a corresponding voltage equation regardless of motoring or regenerating. Therefore, the behaviors of the field current and the capacitor voltage are determined accordingly. In the following section, the behavior of the ZSI at each operating mode under a motoring operation is analyzed on the assumption that the PWM carrier frequency of the inverter is sufficiently high compared with the resonance frequency of the impedance-network.

3.1 Active State Figure 4 shows the equivalent circuit at the *active state* in the motoring mode. Of three possible states, the *active state* is defined by the inverter outputting a non-zero voltage vector and supplying power to the WFSM. The load side is approximated as a constant current source I_i and therefore, the current fed to the inverter i_i should be equal to I_i ($i_i = I_i$). With regard to time t , the beginning time of the *active state* is defined as $t = 0$.

From the voltage equation of impedance-network side at the *active state*, the time response of field current can be expressed as,

$$\frac{di_f}{dt} = \frac{V_0 - v_c - R_f i_f}{L_f} \quad (3)$$

The time response of capacitor voltage can be similarly expressed in the following equation.

$$\frac{dv_c}{dt} = \frac{i_f - i_i}{C} = \frac{i_f - I_i}{C} \quad (4)$$

On the other hand, the current of DC voltage source i_{in} can be expressed in the following manner according to the relationship between the currents in the ZSI.

$$i_{in} = 2i_f - I_i \quad (5)$$

In addition, the following condition has to be satisfied because the switch S_w is always in the OFF in the motoring mode.

$$2i_f - I_i > 0 \quad (6)$$

If (6) is not satisfied, the voltage across the inverter leg v_i becomes zero because a part of the current fed to the motor flows through some freewheeling diodes in some switching devices. As a result, this state becomes identical with the *shoot through state* mentioned later leaving the inverter unable to output the non-zero voltage vector. This state is called the *freewheeling diode shoot through state*⁽⁴⁾ to distinguish it from an intentionally executed *shoot through state*. In order

Fig. 5. Equivalent circuit at *active zero state* (motoring) Fig. 6. Equivalent circuit at *shoot through state* (motoring)

to avoid this state, one of the control restrictions of the ZSI is that the field current i_f has to control to meet (6). As long as the field current i_f is controlled to meet (6), the voltage across the inverter leg v_i meets the following relation.

$$v_i = 2v_c - V_0 \quad (7)$$

3.2 Active Zero State The equivalent circuit at the *active zero state* in the motoring mode is shown in Fig. 5. As the inverter outputs the zero voltage vector in the *active zero state*, the inverter is disconnected from the DC side ($i_i = 0$). With regard to time t , the changing time to the *active zero state* is redefined as $t = 0$. In this case, as the voltage equation of the impedance-network side is identical with the previously mentioned *active state*, the time response of field current can be expressed as similar to (3). The time response of capacitor voltage is similarly expressed in (4), but it can be rearranged as the following equation by considering the inverter disconnection.

$$\frac{dv_c}{dt} = \frac{i_f}{C} \quad (8)$$

The voltage across the inverter leg v_i can be also similarly expressed in (7).

3.3 Shoot Through State The equivalent circuit at the *shoot through state* in motoring is shown in Fig. 6. The *shoot through state* is the state of an inverter arm in short circuit. With regard to time t , the time switching to the *shoot through state* is redefined as $t = 0$.

From the voltage equation of impedance-network side, the time response of field current can be expressed as,

$$\frac{di_f}{dt} = \frac{v_c - R_f i_f}{L_f} \quad (9)$$

The time response of capacitor voltage is obtained from the

following equation.

$$\frac{dv_c}{dt} = -\frac{i_f}{C} \dots \dots \dots (10)$$

The voltage across the inverter leg at this state can be expressed as $v_i = 0$.

4. Control Algorithm for Field Current and Capacitor Voltage

4.1 Control Principles for Field Current and Capacitor Voltage in Proposed Drive System The basic control principle of the ZSI is examined from the time response expression of field current i_f and capacitor voltage v_c under each operating state described in the preceding chapter.

The duty ratios d_A , d_Z , and d_S of the *active state*, the *active zero state* and the *shoot through state* have the following relation.

$$d_A + d_Z + d_S = \frac{T_A}{T} + \frac{T_Z}{T} + \frac{T_S}{T} = 1 \dots \dots \dots (11)$$

The time response of field current at both the *active state* and the *active zero state* is determined by (3), and the one at the *shoot through state* is obtained from (9). The averaging time response of field current during one control interval T is derived from (3), (9) and (11) and given in

$$\frac{di_f}{dt} = \frac{(2v_c - V_0)d_S + V_0 - v_c - R_f i_f}{L_f} \dots \dots \dots (12)$$

(12) means that the field current can be controlled by the *shoot through state* duty ratio d_S based on a feedback of field current. Similarly, the averaging time response of capacitor voltage during one control interval T is derived from (4), (8), (10) and (11) and given in

$$\frac{dv_c}{dt} = \frac{-2i_f d_S - I_i d_A + i_f}{C} \dots \dots \dots (13)$$

(13) implies that the capacitor voltage can be controlled by the *active state* duty ratio d_A based on a feedback of the capacitor voltage under the given *shoot through state* duty ratio d_S as decided by the field current controller. On the other hand, the *active state* duty ratio d_A also has to be determined to meet the requirement for the power control of the WFSM simultaneously. In the proposed drive system, therefore, the capacitor voltage v_c and the motor power have to be controlled by only one control input d_A . In general, this control might be impossible and therefore, we have to introduce another independent control input for the capacitor voltage control. In the right side of (13), the inverter load current I_i corresponds to the input power of the WFSM for the given torque and speed condition. As the input power includes not only the output, but also the inverter loss, motor copper and iron losses, the inverter load current I_i can be adjusted by the motor copper loss which is controlled by the d -axis current with negligible torque change.

4.2 Proposed Capacitor Voltage Control The field and the d - and q -axis current controllers are designed so as to respond faster than the capacitor voltage controller because these current responses directly dominate the torque and motor output response. Under this condition, the controller design for achieving the capacitor voltage and the motor output

controls simultaneously is examined based on the AC small signal modeling method⁽⁵⁾.

Initially, the relation between the input power and the motor output is examined. The input power can be expressed in $v_i i_i d_A$. To explain the control algorithm in simple terms, it can be assumed that the inverter loss and the motor iron loss are negligible compared to the motor copper loss. As a result, the following relation is obtained from (1), (2) and (7).

$$(2v_c - V_0)i_i d_A = R \{i_d^2 + i_q^2\} + \omega \tau \dots \dots \dots (14)$$

Based on the AC small signal modeling, the variables in the above equation can be defined by using the target value (X^*) and the deviation ($\hat{x}$). In addition, the fine adjustment value of d -axis current for the capacitor voltage control mentioned earlier is defined as Δi_d . Under these definitions, each variable is expressed as follows.

$$v_c = V_c^* + \hat{v}_c, \quad i_q = I_q^*, \quad i_f = I_f^*, \quad i_d = I_d^* + \Delta i_d \dots \dots \dots (15)$$

Finally, the transfer function from Δi_d to $\hat{v}_c$ is derived from (2), (12), (13) and (14) by substituting the variables expressed in (15) and given in,

$$G_{v_c} = \frac{\hat{v}_c(s)}{\Delta i_d(s)} = -\frac{2RI_d^* + \omega P(L_d - L_q)I_q^*}{sC(2V_c^* - V_0)} \dots \dots \dots (16)$$

(16) implies that the deviation of the capacitor voltage $\hat{v}_c$ is controllable by the fine tune control of the d -axis current and thus, it is possible to support both the controls of capacitor voltage and motor output together.

4.3 Control System Configuration of Proposed Drive System

Figure 7 illustrates the control block diagram configured based on the examination in the preceding section. The field current controller employs a PI controller based on the error between its reference and detected values. The capacitor voltage controller is also composed of a PI controller based on the deviation between its reference and detected values. In addition, Δi_d as the output of the capacitor voltage controller is delivered to the d -axis current controller so that the independent control of the capacitor voltage and the motor torque and output can be achieved. The WFSM controller takes control of each axis current with a PI controller by employing the decoupling control as with conventional AC machines. The proposed system has to be controlled by the appropriate control reference parameters I_f^* , V_c^* , I_d^* , I_q^* taking into account the control constraint conditions under the given operating point (τ^* , ω). For this reason, the sets of control reference parameters for the operating region are stored in a look-up table in advance.

5. Control Constraint Conditions and Reference Parameter Search

This chapter explains how to search for the control reference parameters to control the field current, the capacitor voltage and the motor output simultaneously at given operating points in the proposed system taking into account its constraints.

5.1 Constraint of Input-output Balance Equation in ZSI

The steady-state input-output relation in the ZSI is expressed in the following equation with the control reference parameters of given operating points from energy conservation.

Fig. 7. Block diagram of the proposed controller

Fig. 8. Duty ratio of each operating state

$$V_0 I_f^* = 2R_f I_f^{*2} + R(I_d^{*2} + I_q^{*2}) + \omega \tau^* \dots \dots \dots (17)$$

The left-hand side of (17) expresses the input from the DC power source. The first term on the right side expresses the copper loss of the field coils. The second and third terms are the copper loss of armature windings and the motor output, respectively.

5.2 Constraint of Field Current Reference The field current reference I_f^* and the current I_i being fed to inverter bridge at the *active state* have to satisfy the following equation to avoid the *freewheeling diode shoot through state*.

$$2I_f^* > I_i \dots \dots \dots (18)$$

5.3 Constraint of Operable Duty Ratio The ZSI is controlled by the duty ratios of the three states. The control reference parameters have to be chosen such that the sum of the duty ratios at operating points does not exceed 1. The following equation is obtained by rearranging (11).

$$m^* = D_A^* + D_{Z-min}^* \leq 1 - D_S^* \dots \dots \dots (19)$$

Where D_{Z-min}^* is the minimum duty ratio reference of the *active zero state*. m^* is the sum of D_A^* and D_{Z-min}^* and expressed by another equation below for the reason that the duty ratio equals the area ratio as shown in Fig. 8.

$$m^* = \frac{2\pi}{3\sqrt{3}} D_A^* \dots \dots \dots (20)$$

Fig. 9. Flowchart of control reference parameters search

Rearranging (19) and (20), the following equation is derived.

$$D_A^* \leq \frac{3\sqrt{3}}{2\pi} (1 - D_S^*) \dots \dots \dots (21)$$

Thus, the ZSI has to be controlled such that the duty ratios D_A^* and D_S^* satisfy (21).

5.4 Control Parameter Search at Given Motor Operating Points

The flowchart of control reference parameters search at given operating points of the drive system of the WFSM integrated with the ZSI is shown in Fig. 9. The field current I_f^* , the armature current I_d^* , I_q^* and the capacitor voltage V_c^* as the control reference parameters satisfying the constraint conditions of (17), (18) and (21) is searched at given torque reference and motor speed condition. They are eventually stored in a look-up table.

6. Experimental Results

Experimental verifications of the proposed control algorithm are conducted using a test WFSM and a test ZSI as shown in Figs. 10 and 11. The measured WFSM parameters and the capacitance C of impedance-network used in the ZSI appear in Table 1. The dimensions of the main machine part of the WFSM are the outer diameter of 100 mm and the stack length of 26.6 mm, respectively. The rated torque of the test

Fig. 10. Photographs of tested WFSM

Fig. 11. Photographs of fabricated ZSI for test

Table 1. Measured parameters of WFSM and ZSI

WFSM machine parameters	R [Ω]	0.36@20°C
	L_d [mH]	0.82(nominal)
	L_q [mH]	0.98(nominal)
	R_f [Ω]	1.13@20°C
	L_f [mH]	30(nominal)
ZSI	C [μ F]	9.9

WFSM is approximately 1.5 Nm under ambient air cooling. The rated speed of the motor is 3,000 r/min when the conventional three-phase inverter with the rated DC bus voltage of 283 V is used. In experiments, the DC voltage source is set to 60 V constant in order to check voltage boost-up capability. Since the capacitor voltage v_c is fluctuated owing to the use of a small capacitance of 9.9 μ F, the maximum capacitor voltage limit is set to 130 V considering the withstand voltage of the switching devices in the inverter (600 V). This results in

Fig. 12. Experimental result operating at 0.94 Nm–250 r/min

Fig. 13. Experimental result operating at 1.34 Nm–1,000 r/min

Fig. 14. Experimental result operating at 0.63 Nm–2,000 r/min

the maximum voltage across the inverter leg v_i at the *active state* is 200 V as determined by (7). Thus, the rated motor speed is reduced by 2,000 r/min. The test WFSM is in torque control mode under speed control by the dynamo side.

Figure 12 demonstrates the experimental result at 0.94 Nm under 250 r/min. The command values of d -axis current I_d^* and q -axis current I_q^* are set to 0 A and 6.8 A, respectively. The field current command I_f^* is set to 8.0 A. In this experiment, the maximum capacitor voltage limit is intentionally changed from 130 V to 100 V to verify the effectiveness of the proposed capacitor voltage control. It is found that all currents converge to their references. The actual capacitor voltages V_{c1} and V_{c2} are boosted more than 100 V, resulting in the voltage across the inverter leg v_i being more than 140 V. The detected capacitor voltage V_{c-det} in the controller follows the maximum capacitor voltage limit $V_{c-limit}$. Thus, it is concluded that the proposed capacitor voltage control works well. The two traces from the bottom in the figure show the three-phase current waveforms for the enlarged time scale during 1 to 1.03 s and 3 to 3.03 s before and after the maximum capacitor voltage limit change, respectively. As apparent from the waveforms, the three-phase currents are controlled pure sinusoidal. The torque ripple, which can be

observed after starting the capacitor voltage control, would be due to low-frequency capacitor voltage ripple.

Figure 13 demonstrates the experimental result at an approximately rated torque of 1.34 Nm under 1,000 r/min. The command values of d -axis current I_d^* , q -axis current I_q^* and field current I_f^* are -5.2 A, 9.0 A and 10.0 A, respectively. From the figure, it can be seen that all currents follow their references.

Figure 14 shows the experimental result at 0.63 Nm under 2,000 r/min. The command values of d -axis current I_d^* , q -axis current I_q^* and field current I_f^* are -3.5 A, 6.0 A and 5.0 A, respectively. From the figure, it can be found that all currents follow their references. Consequently, it is concluded that the proposed control algorithm is effective for operating the proposed drive system.

Here, Table 2 summarizes the comparisons of the motor efficiency and the losses of the tested motor at each operating point corresponding to Figs. 13 and 14 when the proposed drive system and a non-integrated drive system are used. The non-integrated drive system is composed of separated circuits such as a DC chopper for field coil excitation and a three-phase inverter for feeding current to the armature windings. P_{min} is measured by a Yokogawa Digital Power Meter WT

Table 2. Efficiency and losses with WFSM and ZSI-WFSM

		1.34Nm-1,000r/min		0.63Nm-2,000r/min	
		WFSM	ZSI-WFSM	WFSM	ZSI-WFSM
Motor input P_{min}		425 W	429 W	212 W	220 W
Motor output P_{mout}		137 W	139 W	123 W	131 W
Copper loss	Armature w.d.g. P_{ca}	38 W	39 W	17 W	17 W
	Field coils P_{cf}	224 W	231 W	57 W	60 W
Mechanical loss P_{mech}		4 W	4 W	6 W	6 W
Residual loss P_{res}		22 W	17 W	9 W	6 W
Motor efficiency		32.2 %	32.4 %	58.0 %	59.5 %
Motor power factor		0.892	0.928	0.919	0.935

1600. P_{mout} is calculated from the motor speed and the average torque measured by the dynamo setup, Ono Sokki, TS-7700. P_{ca} and P_{cf} are calculated by the resistances in Table 1 and the measured rms currents of the armature windings and DC current of the field windings. P_{mech} is also premeasured by the dynamo setup. The motor efficiency and the motor power factor of the proposed integrated WFSM are slightly higher than those of the non-integrated WFSM. In the table, the residual loss P_{res} is calculated from the following equation.

$$P_{res} = P_{min} - P_{mout} - P_{ca} - P_{cf} - P_{mech} \dots \dots \dots (22)$$

Generally, the residual loss is regarded as the iron loss. The residual loss of the WFSM coupled in the proposed drive system is slightly smaller than that in the non-integrated drive system. As a result, it is reasonable to say that the WFSM coupled in the proposed integrated drive system can keep equivalent drive performances with those in the non-integrated drive system.

7. Conclusions

This paper has proposed a variable motor drive system, a WFSM integrated with a ZSI. The basic working principle of the proposed integrated drive system has been explained. To solve the problem of reduced control freedom in the proposed integrated drive system, a control algorithm employing d -axis current control has been proposed. Some experimental tests have been conducted that verified the validity of the proposed control algorithm. In future studies, we will clarify the design procedure of the proposed integrated drive system for a specific application and will evaluate the advantages from the viewpoints of system efficiency as well as cost and size reduction.

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