Control Method for Reducing the Motor Loss of Dual-inverter Fed Open-end winding Induction Motor in the Low-speed Region

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This paper proposes a method for improving the output voltage waveform of dual-inverter fed open-end winding induction motors in the low-speed region. The system consists of two isolated DC power supplies, an open-end winding induction motor, and two voltage source inverters, which are connected to the opposite terminals of the open-end winding with unequal DC-link voltages. The motor losses, which are caused by the harmonics of the voltage applied to the winding, increase when the inverter outputs a pulse-width modulated voltage at a low modulation index. The aim of this paper is to reduce the motor loss at low motor speeds by improving the winding voltage waveform. In the proposed control method, the use of the output voltage difference from the inverters constituting the dual-inverter lowers the peak voltage. In order to obtain the proposed waveform, a method for the space-vector pulse-width modulation of the dual-inverter is presented. Furthermore, this paper presents the commutation sequence of the PWM pattern considering the dead-time and switching device characteristics, especially the turn-off delay time. The proposed method is verified experimentally by driving a dual-inverter fed open-end winding induction motor. The experimental results confirm that the harmonic component of the output voltage, which is related to the switching frequency (5 kHz), is reduced from 18.9 V to 4.4 V by using the proposed method at a motor speed of 300 rpm and load torque of 1.43 Nm.

Keywords: open-end winding induction motor, space-vector modulation, dual-inverter, low modulation index

1. Introduction

Recently, motor drive systems are required a high-efficiency characteristic in a wide speed range not only at a rated speed. Currently, various approaches are being pursued to reduce the motor losses. Multilevel inverters are an effective solution for improving the efficiency of motor drive systems[5]. Since the multilevel inverter outputs a stair-step output voltage, the harmonic components of the output voltage are small as compared with the 2-level inverter. Another approach is to reduce the peak value of the pulse-width modulated (PWM) voltage by using the DC/DC converter, which is also effective for improving the motor efficiency[6]. The multilevel waveforms and lower peak value of the voltage can reduce the output harmonics and the motor’s harmonic losses[7,8].

A motor drive system using a dual-inverter is one of the circuit topologies used to achieve motor loss reduction, which is also effective for improving the waveform of the winding voltage with multilevel operation[8,9]. The dual-inverter system consists of two voltage source inverters, which are connected respectively to the opposite terminals of the open-end winding motor.

For the dual-inverter, various circuit topologies and PWM strategies have been considered, achieving multilevel operation[10,11], common-mode voltage and zero-sequence current reduction[12,13], and wide speed range operation[14]. In particular, the dual-inverter topology, which has two 2-level inverters with isolated DC-links in each inverter, can generate a five-level waveform in the winding phase voltage by keeping each DC-link voltages in a ratio of 2 : 1[8]. Through these researches, motor drive systems can achieve a high efficiency at rated power, in other words high modulation index, due to generating a multilevel waveform. In the case of high modulation index, the PWM pattern, which operates in either inverter, is explained in (5). However, while the PWM modulation index is low, the harmonic distortion of the inverter’s output voltage is high, because the harmonic component of the switching frequency increases compared with the fundamental component[9,10]. The voltage distortion is the cause of the increase in motor losses. Recent motors have efficiency of more than 80% at rated speed because of the development of motor structures or materials. However, the efficiency decreases at low speeds (e.g. the efficiency of 44% at 300 rpm in a 3 kW standard squirrel-cage induction motor reported in (19)), and one of the cause of motor losses is voltage distortion. For variable-speed applications, such as electric vehicles and flywheel energy storages, etc., the high efficiency in a wide speed range is important in terms of the energy saving. Especially in the case of a high rated voltage motor, the more iron losses occur at low speeds, because the harmonic components increase than the fundamental component. Using the dual-inverter, the number of switching devices increases. However, switching devices which have a lower voltage rating can be used in the dual-inverter, because the effective motor winding voltage is the sum of each inverter’s output voltage at high speed region. Therefore, even
if the number of switching devices increases, improving the winding voltage waveform is important for increasing the total efficiency, which is comprised of the inverter and motor efficiency.

This paper proposes an improvement method of the output voltage waveform at low motor speeds. The original point of this paper is that, in order to reduce the peak PWM value at low motor speeds, the dual-inverter outputs the voltage difference between the two inverters. In the proposed method, the output voltage difference of the inverters constituting the dual-inverter lowers the peak value of the output voltage at low motor speeds by changing the DC-link voltage of the secondary inverter using a DC/DC converter. This paper also proposes a commutation sequence to improve the output voltage waveform. The 3-step commutation pattern is presented to avoid the effect of the dead-time, which is caused by freewheeling diodes and phase-current direction (20). The proposed method is effective to reduce the motor losses in the low-speed region, whereas in the high-speed region, the dual inverter can realize wide range variable-speed drive by supplying the sum of respective output voltages to the motor winding.

This paper is organized as follows; firstly, the configuration of the dual-inverter fed open-end winding induction motor drive system and space-vector diagrams of the dual-inverter are introduced. Secondly, the principle of relationship between voltage harmonics and modulation index in a space-vector modulated inverter, and the proposed PWM method, which realizes the reduction of the PWM peak value, are described. Next, the commutation sequence of the PWM pattern is explained in order to achieve the proposed PWM waveform which is affected by the dead-time depending on the characteristic of switching devices, in particular turn-off delay time. Finally, the experimental results demonstrate that the dual-inverter drives the open-end winding induction motor using the proposed PWM method with lower peak PWM winding voltage. Furthermore, motor loss reduction is verified at low motor speeds compared with the single inverter drive.

2. Dual-inverter Fed Open-end Winding Induction Motor

2.1 Circuit Configuration of the Dual-inverter

A configuration of the open-end winding induction motor fed from a dual-inverter is shown in Fig. 1. The output voltage difference between the inverters is supplied to the stator windings. Thus, the U-phase winding voltage \( v_u \) (as well as V-phase voltage \( v_v \) and W-phase voltage \( v_w \)) is expressed as follows:

\[
v_u = v_{u1} - v_{u2} \tag{1}
\]

Here, \( v_{u1} \) and \( v_{u2} \) are the output voltage of each inverter. From the equation, a higher or lower voltage than that of each inverter can be supplied to the motor winding by using the phase difference between the two inverters.

2.2 Space-vectors of the Dual-inverter

This paper applies Space-vector Modulation (SVPWM) on the individual 2-level inverters shown in Fig. 2. In this paper, the switching states of INV.1 and INV.2 are respectively numbered as 0, 1, 2, ..., 7 and 0', 1', 2', ..., 7'. Here, a “+” means that the top switch of a leg is turned on, while a “-” means that the bottom switch of the leg is turned on. Thus, the dual-inverter, which consists of two 2-level inverters, has 64 possible switching states (18). Because the winding voltage is expressed as the difference voltage between two inverters (from equation (1)), the resulting vector combinations of the dual-inverter are as shown in Fig. 3, when the ratio of the DC-link voltages is \( V_{dc1} : V_{dc2} = 2 : 1 \) and \( 3 : 2 \) for example. These figures show that the dual-inverter outputs various voltage vectors depending on the DC-link voltage ratio of the inverters. In particular, Fig. 3(a) shows that the 7-level phase voltage waveform can be generated due to the high redundancy of the switching pattern (18-27).

2.3 Conventional Operation at Low Motor Speeds

In the conventional method, which is reported in (18), the converter losses can be reduced when the motor is driven at low speeds (low modulation index) because of the dual-inverter outputting the PWM waveform by using only the INV. 1. In this paper, it is called single-inverter operation. A problem of this operation is that, the winding voltage distortion increases because the peak voltage of PWM is fixed at \( 2V_{dc1}/3 \). Therefore, the motor losses caused by voltage distortion will increase when the motor is driven at low speed.

3. Concept of Voltage Harmonics Reduction

3.1 Relationship Between Voltage Harmonics and Modulation Index in 2-Level Inverter

According to Ref. (18), the iron losses of a motor fed by a pulse-width modulated inverter change depending on the modulation index when the motor is driven at constant speed with different DC-link voltage. Therefore, the result indicates that the iron losses are related to the winding voltage harmonics, which are caused by switching. In order to clarify the voltage harmonics when the 2-level voltage source inverter is operated with space-vector pulse-width modulation (SVPWM), a normalized RMS value of the voltage harmonics is theoretically

Fig. 1. The dual-inverter circuit which is considered in this paper

Fig. 2. Space-voltage vector diagram of INV.1 and INV.2
and voltage references harmonics components. In this paper, a normalized value of the voltage indicated in this paper.

In general, the RMS value of a voltage waveform is defined as follows:

\[
V_{\text{rms}} = \sqrt{\frac{1}{T} \int_0^T (v(t))^2 dt} \quad \cdots (2)
\]

Here, \(T\) is the period of fundamental component, and \(v(t)\) is the U-phase voltage which changes depending on the time \(t\). The \(V_{\text{rms}}\) includes a fundamental component and harmonic components. In this paper, a normalized value of the voltage harmonics \(V_{h_{\text{normalized}}}\) is defined as follows:

\[
V_{h_{\text{normalized}}} = \frac{V_{\text{rms}} - V_{1_{\text{rms}}}}{V_{1_{\text{rms}}}} \quad \cdots (3)
\]

\[
V_{1_{\text{rms}}} = \sqrt{\frac{1}{T} \int_0^T (v_1^*(t))^2 dt} \quad \cdots (4)
\]

Here, \(V_{1_{\text{rms}}}\) is the RMS value of the fundamental component of the U-phase voltage which is expressed as equation (4) by using a voltage reference \(v_1^*(t)\).

The case of using the SVPWM which is shown in Fig. 4 and voltage references \(v_1^*, v_1^*\) and \(v_0^*\) are expressed as equation (5), the output time of each vectors \(T_1, T_3\) and \(T_7\) are expressed as equation (6). Furthermore, when the DC-link voltage of the 2-level inverter is \(V_{dc}\), the U-phase voltage values are as in equation (7).

![Fig. 3. Space-vector combinations of the dual-inverter when the DC voltage ratio is (a) \(V_{dc1}: V_{dc2} = 2:1\) and (b) \(V_{dc1}: V_{dc2} = 3:2\)](image)

\[
\begin{bmatrix}
V_u^* \\
V_v^* \\
V_w^*
\end{bmatrix} = \frac{V_{dc}}{\sqrt{3}} \begin{bmatrix}
\cos \omega t \\
\cos (\omega t - \frac{2\pi}{3}) \\
\cos (\omega t - \frac{4\pi}{3})
\end{bmatrix} \quad \cdots (5)
\]

\[
\begin{bmatrix}
T_1 \\
T_3 \\
T_7
\end{bmatrix} = \begin{bmatrix}
m \cos (\omega t + \frac{\pi}{6}) \\
m \sin \omega t \\
1 - m \sin (\omega t + \frac{\pi}{3})
\end{bmatrix} \quad \cdots (6)
\]

\[
v_u(t) = \begin{cases}
\frac{2}{3} V_{dc}, & \text{when output vector is } V_1 \\
\frac{1}{3} V_{dc}, & \text{when output vector is } V_2 \\
0, & \text{when output vector is } V_0
\end{cases} \quad \cdots (7)
\]

Here, \(m\) is the modulation index \((0 \leq m \leq 1)\), \(\omega\) is the angular frequency of fundamental component \((=2\pi/T)\) and \(T_s\) is the sampling period. From Eqs. (6) and (7) into (2) and (4), the \(V_{\text{rms}}\) and \(V_{1_{\text{rms}}}\) are given by

\[
V_{\text{rms}} = V_{dc} \sqrt{\frac{2m}{3\pi}} \quad \cdots (8)
\]

\[
V_{1_{\text{rms}}} = \frac{V_{dc} m}{\sqrt{6}} \quad \cdots (9)
\]

Therefore, the \(V_{h_{\text{normalized}}}\) is given by

\[
V_{h_{\text{normalized}}} = \frac{2}{\sqrt{\pi m}} - 1 \quad \cdots (10)
\]

The \(V_{h_{\text{normalized}}}\) indicates the harmonic components of the U-phase voltage waveform which is caused by the SVPWM and it is related to modulation index \(m\). Therefore, a reduction of voltage harmonics can be achieved by keeping the modulation index \(m\) at a higher value, when the rotor speed and the torque are constant.

3.2 Proposed Operation for Dual-inverter In the proposed method, the dual-inverter outputs the PWM waveform, which has a low peak voltage by using INV. 1 and INV. 2 voltage vectors which have the same angle, while the DC/DC converter regulate the DC-link voltage \(V_{dc}\) depending on the motor speed keeping \(V_{dc1} > V_{dc2}\). The DC-link voltage of the left-side inverter (INV. 1) is higher than the right-side inverter (INV. 2). In the proposed method, the peak value of the PWM supplied to the winding is \(2(V_{dc1} - V_{dc2})/3\) by outputting synchronized pulses in both inverters, which means that switching states 0°, 11°, 22°, 33°, 44°, 55°,
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Fig. 5. The output voltage vectors of the proposed method

Fig. 6. An image of the Proposed PWM waveforms

66', and 77' are only used as shown in Fig. 5. Therefore, the output voltages are expressed as \( \pm 2(V_{dc1} - V_{dc2})/3 \), \( \pm (V_{dc1} - V_{dc2})/3 \), and 0. PWM waveform images of the proposed method are shown in Fig. 6. It indicates the PWM peak value can be related to only the difference between each DC-link voltage of INV.1 and INV.2. Therefore, the proposed method improves the voltage distortion because the peak value of the PWM is reduced compared with the conventional method. Hence, the proposed method improves the motor losses at low speeds.

The value of the DC-link voltage \( V_{dc2} \) is changed depending on the motor speed in this proposed method. If a high-speed operation is required, by making \( V_{dc2} \) the same value as \( V_{dc1} \) and outputting the sum voltage of dual-inverter, a twice voltage can be obtained than the single-inverter. Furthermore, if a multilevel operation is required, by making the ratio of \( V_{dc1} \) and \( V_{dc2} \) 2 : 1, a 4-level operation can be achieved same as 4-level single-inverter.

4. PWM Method to Realize Proposed Waveform

4.1 Current Commutation for Proposed Method

In the proposed PWM method, the peak value of the winding voltage pulses can be reduced by outputting synchronized pulses in both inverters ideally as shown in Fig. 7. In general, voltage source inverters require the dead-time because the peak value of the PWM is reduced compared with the conventional method. Hence, the proposed method improves the motor losses at low speeds.

![Fig. 7. The U-phase circuit of the dual-inverter operated in the proposed PWM method](image)

![Fig. 8. A switching sequence and current flow during the dead-time](image)

![Fig. 9. Improved commutation to avoid the effect of turn-off delay time](image)

dead-time, an error voltage is applied to the winding because the freewheeling diodes \( D_{n1} \) and \( D_{n2} \) (or \( D_{p1} \) and \( D_{p2} \)) start conducting as shown in Fig. 8.

Figure 8 shows the effect of the dead-time on switching states, when the switching states of both inverters are changed from “+” to “-”. The diodes, which turn on in the dead-time, depend on the direction of the phase current. Therefore, the current commutation is considered to avoid the effect of the dead-time reported in (8) as shown in Fig. 9(a).

The commutation pattern is achieved by detecting each phase current and it is performed in 3-steps operation. On the first step, \( S_{sp1} \) is turned off as the freewheeling diodes start conducting. The switching states of both inverters are not changed because the freewheeling diode is still conducting. On the second step, \( S_{sp2} \) is switched off and \( S_{sn2} \) is turned on at the same time. The switching states of both inverters are changed because the freewheeling diode is still conducting. On the third step, \( S_{sp1} \) is switched off and \( S_{sn2} \) is turned on at the same time. The switching states of both inverters are changed because the freewheeling diode, which is located on the opposite side of \( S_{sp1} \), starts conducting in this step. Finally, the commutation operation is finished by turning on \( S_{sn1} \). Figure 9(a) shows an example of the current commutation sequence when the switching states of both inverters are changed from “+” to “-”.

4.2 Compensated Switching Pattern Considering Turn-off Delay Time of the Switching Device

In order to achieve the proposed operation, the changing timing of the switching states of each inverter should be synchronous at
the second step as mentioned in the previous section. However, the current commutation does not work well when the instantaneous value of the phase current is close to zero, due to the turn-off delay time of the switching device. Figure 10 shows the switching waveforms of the dual-inverter; it indicates that the delay time, which is the period from the end of State 1 to the beginning of State 2, is 50 ns at 4.82 A and 970 ns at 158 mA.

In general, the turn-off delay time is inversely proportional to the phase current. In this paper, the dead-time insertion timing is compensated depending on phase current as shown in Fig. 9(b). In Fig. 9(b), the falling timing of the gate-source voltage of \( S_{n2} \) and the rising timing of the gate-source voltage of \( S_{p2} \) are made \( T_{\text{delay}} \) earlier from Fig. 9(a). Therefore, the period from the end of State 1 to the beginning of State 2 will be zero, thus the error of the phase voltage can be eliminated. In this commutation method, when the phase current is close to zero crossing, the delay time equals to the dead-time, which means that there is no commutation step. On the other hand, as the phase current increases, the delay time is close to zero.

5. Experimental Results

5.1 Experimental Setup and Conditions  
The experimental setup is shown in Fig. 11, whereas the experimental conditions and parameters of the open-end winding induction motor are shown in Tables 1 and 2. The proposed method is only related to modulation, thus the voltage reference and the phase-current values are needed. Therefore, there is no design restriction on the motor. A general-purpose induction motor (TFO-FK 0.75KW 4P 200 V, made by HITACHI) is used for the experiment. In this experiment, INV. 1 DC-link voltage is regulated at 180 V by using a DC power supply, while INV. 2 DC-link voltage can be changed by using a DC electronic load. Figure 12 shows the control block diagram of the proposed dual-inverter drive. In this proposed method, some control methods, which calculates the voltage reference such as an MPTA control and V/F control, can be used. In this paper, the slip frequency type field oriented control (FOC) is used in order to keep the mechanical output power and the fundamental component of the output voltage at constant value when the difference of the DC-link voltage changes. The switching patterns of the dual-inverter are selected using the switching table depending on the voltage reference \( V_{\text{ref}} \) and each phase current value \( i_a, i_c, i_w \). For the verification of
the proposed method, the open-end winding induction motor is driven at a constant speed by the load motor and inverter, and is loaded with constant torque using the dual-inverter.

5.2 Turn-off Characteristic of MOSFETs and the Result of the Compensation

A look up table is created from the delay time characteristic in order to compensate the dead-time. The relationship between the turn-off delay time and the phase current value is experimentally measured when MOSFETs (FK30SM-6) are used as the switching device and it is shown in Fig. 13. From the characteristic, the look up table of $T_{\text{delay}}$ is calculated following equation of the fitted curve;

$$T_{\text{delay}} = \frac{2.463I_u + 141.4}{I_u}$$

Using this look up table, the switching timing is compensated. Figure 14 shows the voltage waveforms (a) without turn-off delay time compensation and (b) with turn-off delay time compensation when $V_{\text{dc},2} = 120$ V.

5.3 Waveforms and Harmonics Analysis

Figure 15 shows the winding voltage and current waveforms of the U-phase with the conventional method and proposed method. The conventional method operates using only INV. 1, having the DC-link voltage at 180 V (Fig. 15(a)). On the other hand, the difference of the DC-link voltage is 50 V in the proposed method (Fig. 15(b)). It is verified that the proposed method reduced the peak voltage from 180 V to 120 V.

Figure 16 shows the results of the harmonics analysis of the U-phase voltage shown in Fig. 15. The harmonic components of the output voltage, which are related to the switching frequency (5 kHz), is reduced from 18.9 V to 4.39 V by the proposed method compared with the conventional method while the fundamental component is the same.

5.4 Normalized Voltage Harmonics and Modulation Index Characteristic

Figure 17 shows the characteristics of normalized voltage harmonics and modulation index; experimental results are given by calculating the RMS value from the voltage waveform, while the fundamental RMS value is calculated using the modulation index. Figure 17 indicates that the experimental results tend to the theoretical line, which is calculated from Eq. (10). It is verified that the proposed PWM method can reduce the voltage harmonics by keeping the modulation index high.
5.5 Motor losses and Modulation Index Characteristic

The relationship between motor losses and modulation index is experimentally verified. In order to control the modulation index, the DC voltage difference $V_{dc1} - V_{dc2}$ is regulated as shown in Fig. 18, and the modulation index is kept less than 1.0 in order to avoid occurring the voltage saturation. Here, the fundamental component of the voltage is kept constant. Figure 19 shows the motor losses characteristics when the modulation index is controlled as in Fig. 18 at 200 rpm and 300 rpm. It is verified that the motor losses are reduced 2.58 W compared with the conventional method at 200 rpm. Furthermore, the motor efficiency is improved 0.39% at 300 rpm and 0.27% at 200 rpm as shown in Fig. 20.

In this experiment, the modulation index is calculated by average of the voltage reference, which is generated by a current controller in order to keep the output power constant. Therefore, as increasing the modulation index, voltage saturation is more likely to occur instantaneously. Because of the instantaneous voltage saturation, low-order harmonics are generated thus the motor efficiency decreases when the modulation index is higher than approximately 0.8 as shown in Fig. 20(b).

5.6 Total Efficiency of the Motor and Converter

The relationship between converter efficiency and modulation index is experimentally verified same as motor efficiency. Figure 21 shows the characteristics of converter efficiency. Because the switching devices of INV.2 is not operated in the conventional method, the converter efficiency is higher than the case of using proposed method. In the proposed method, the converter efficiency decreases with the increase of modulation index because of the increase of $V_{dc2}$. The total efficiency, which is comprised of the efficiency of motor and converter, is shown in Fig. 22. The results indicate that the total efficiency increases compared with the conventional method, due to keeping the modulation index high.

6. Conclusions

This paper proposes a control method for the dual-inverter fed open-end winding induction motor at low motor speeds. Furthermore, the RMS value of the voltage harmonics caused
The proposed method is effective to reduce the motor losses in the low-speed region. Furthermore, in the high-speed region, the dual inverter can realize wide range variable-speed drive by supplying the sum of respective output voltages to the motor winding.

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