Control of Single-Phase to Three-Phase AC/DC/AC PWM Converters for Induction Motor Drives

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Abstract—This paper proposes a control scheme of single-phase to three-phase PWM converters for low power three-phase induction motor drives, where a single-phase half-bridge PWM rectifier and a two-leg inverter are used. With this converter topology, the number of switching devices is reduced to six from ten in case of full-bridge rectifier and three-leg inverter systems. In addition, the source voltage sensor is eliminated with a state observer which controls the deviation between the model current and the system current to be zero. A simple voltage modulation method is used for a two-leg inverter and a new technique to eliminate the effect of the dc link ripple voltage on the output current is proposed. Though the converter topology itself is of lower cost than the conventional one, it retains such functions as sinusoidal input current, unity power factor, dc-link voltage control, bidirectional power flow, and VVVF output voltage. The experimental results for the V/F control of 3Hp induction motor drives controlled by a DSP TMS320C31 chip have verified the effectiveness of the proposed scheme.

Index Terms—Ac/dc/ac PWM converter, two-leg inverter, induction motor, source voltage estimation, dc-link voltage ripple.

I. INTRODUCTION

In view of the efficiency, power factor, and torque ripples, a three-phase induction motor is preferable to a single-phase induction motor. So, it is desirable to replace the three-phase induction motor drives by the single-phase induction motor drives in residential appliance, farming, and low power industrial applications [1].

Where only a single-phase utility is available, a single-phase to three-phase power converter system is required to feed the three-phase induction motor drives. Conventionally, a full-bridge diode rectifier plus three-leg PWM inverter has been used. However, the diode rectifier produces harmonic currents to flow into the supply and have no capability of regenerative operation [2].

A few recent papers dealt with the half-bridge PWM rectifier connected to the front end of the two-leg inverter [3]-[8]. However, it has not been studied so much. Moreover, the sensor elimination technique for this kind of converter is very rare [9]-[11].

In this paper, a low-cost half-bridge PWM rectifier plus two-leg PWM inverter for the three-phase induction motor drives are studied, and an algorithm of the source voltage sensor elimination technique is presented, which makes the system cheaper and robust to measurement noises. Also, a compensation scheme of the dc-link ripple voltage effect on the inverter output is proposed. In addition, a balancing control of the neutral voltage in the dc link is applied for the symmetrical output voltage of the inverter. The proposed algorithm has been implemented for the V/F control of the three-phase induction motor drives controlled by a DSP TMS320C31.

II. CONVENTIONAL SINGLE-PHASE TO THREE-PHASE AC/DC/AC CONVERTERS

The simplest circuit of ac/dc/ac converter topology converting from a single-phase supply to three-phase VVVF system is a single-phase full bridge diode rectifier and three-leg PWM inverter system. This circuit is simple and relatively of low cost. However, it has significant disadvantages such non-reversible power flow and source current distortion with poor power factor [1].

Fig 1 shows the power circuit structure with a dc boost chopper intervening between the diode rectifier and the three-leg PWM inverter. With the dc chopper, the source current can be controlled to be sinusoidal with unity power factor. However, the power flow is still unidirectional, not reversible [2].
Fig. 3. Single-phase half-bridge plus two-leg three-phase ac/dc/ac PWM converter topology.

Fig. 2 shows the single-phase full-bridge PWM rectifier with three-phase full-bridge PWM inverter system. This converter gives an excellent performance such as sinusoidal control of source current, unity power factor control of the source-side, constant dc voltage control, and bi-directional power flow. However, it requires ten active switching devices, so, it is expensive [2]-[4].

III. SINGLE-PHASE HALF-BRIDGE PWM RECTIFIER

Fig. 3 shows the single-phase half-bridge PWM rectifier and two-leg three-phase PWM inverter. Compared with the circuit of the Fig. 2, the number of the switching devices is reduced to six from ten, while it almost retains the merits of Fig. 2 [3]-[5]. The circuit topology shown in Fig. 3 itself is nothing new. However, it has been usually applied to the two-phase induction motor drives, where one output is fed to the main winding and the other output is for the auxiliary winding [12].

In this paper, application of this converter to the V/f control of the three-phase induction motor drives is studied. In addition, the source voltage elimination technique is proposed to reduce the system cost and better reliability to the sensing noise.

A. Half-bridge PWM rectifier

There are two operating modes, that is, charging and discharging modes in half-bridge PWM rectifier as shown in Fig. 4 [4],[5], where $e_s$ and $i_s$ is the source voltage and current, respectively, and $v_{dc1}$ and $v_{dc2}$ are the capacitor voltages in the dc side, and $v_s$ is the converter input voltage.

For the upper part of the rectifier when $i_s>0$, the current and the voltage are given by

$$i_c = i_{dc1} - i_l$$
$$i_{dc1} = S_a \cdot i_s$$
$$v_{dc1} = \frac{1}{C} \int i_s dt = \frac{1}{C} \int (i_{dc1} - i_l) dt$$

where, $S_a$ is a switching state, 1 or 0. For the lower part, the similar equations are obtained as

$$i_c = i_{dc1} - i_l$$
$$i_{dc1} = S_a \cdot i_s$$

Expressing (9) in a discrete domain, gives

$$v_s(n-1) = R_i i_s(n) + \frac{L_i}{T} v_s(n-1) + v_s(n-1)$$

The source voltage in (10) is expressed as

$$v_s(n-1) = V \cos \theta$$

where $V$ and $\theta$ are the magnitude and the phase angle of the source voltage, respectively. The phase angle at the n-th sampling instant is given by

$$\theta(n) = (n-1)\omega T + \theta_0$$

where $\theta_0$ is an initial angle at (n-1)th sampling. From (10), the source current in a discrete domain can be written as
current control block

\[ i_s^* = \frac{\int \Delta V}{L_s} \]

On the other hand, the estimated source voltage in the rectifier model can be expressed as

\[ v_M(n) = V_M \cos \theta_M \]

where \( V_M \) and \( \theta_M \) are the magnitude and the phase angle of the source voltage in the rectifier model, where the subscripts "M" mean the variable or parameter for the rectifier model.

Similar to (10), the model current is given by

\[ i_M(n) = i_s(n-1) + \frac{T}{L_s} (v_M(n-1) - R_s i_s(n-1) - v_r(n-1)) \]

where \( L_M \) and \( R_M \) are the model parameters.

It is assumed that the inductance and resistance in the real system and the model are the same, which is usually reasonable in the rectifier system. Then, subtracting (15) from (13) gives

\[ \Delta i_s(n) = i_s(n) - i_M(n) \]

\[ = \frac{T}{L_s} (\Delta V \cos \theta_M - \Delta \theta \cdot V \sin \theta_M) \]

It should be noted that the deviation of \( \Delta i_s \) stems from the estimation error of the source voltage, that is, the magnitude error and the phase angle error.

IV. PWM IN TWO-LEG INVERTERS

Fig. 7 shows the two-leg inverter which gives three-phase VVVF output voltage.

For the induction motor drive, the three-phase voltage references are given in a balanced set as

\[ v_{ac} = V_m \cos \omega t \]

\[ v_{ac} = V_m \cos(\omega t - \frac{2\pi}{3}) \]

\[ v_{ac} = V_m \cos(\omega t + \frac{2\pi}{3}) \]

There are two kinds of voltage modulation schemes for two-leg three-phase PWM inverter [7]. One is named a vector modulation using a space voltage vector concept. The other is called a scalar modulation, which uses phase voltages in calculating the switching time. Here, the scalar modulation scheme is adopted since it is simple and straightforward.

Since the two phases of the induction motor are connected to the inverter legs and the third phase is connected to the neutral point of the dc link, the line-to-line voltage can be used for the PWM instead of the phase voltage.

Let the c-phase be connected to the neutral point. Then, from (21)-(23), the two line-to-line voltage references are given by

\[ v_{ac}^* = v_{ac}^* - v_{cs}^* \]

\[ = \sqrt{3} V_m \cos(\omega t - \frac{\pi}{6}) \]
\[ v_{hc}^* = v_{hs}^* - v_{cs}^* = \sqrt{3} V_m \cos(\omega t - \pi/2) \]  
(25)

The magnitude of these two voltages is \( \sqrt{3} \) times that of the phase voltage and displaced by \( \pi/3 \) each other. By using the proportionality of the triangle as shown in Fig. 8, the switching time can be calculated as

\[ T_1 = T_s \frac{v_{ac}}{v_{ac}} + \frac{v_{ac}}{v_{ac}} T_s \]  
(26)

\[ T_2 = T_s \frac{v_{ac}}{v_{ac}} + \frac{v_{ac}}{v_{ac}} T_s \]  
(27)

When \( v_{dc1} \) and \( v_{dc2} \) are exactly equal each other, the time duration of \( T_1 \) and \( T_2 \) in (26) and (27) gives precise modulated voltages. However, due to the dc link voltage ripples and the dead time effect, the output currents as well as the output voltages are unbalanced and distorted. The deterioration of these waveforms can be eliminated by compensating for the dc link voltage ripples and the dead time effect.

From (26) and (27), the averaged voltages over one switching period are calculated as

\[ \overline{v_{ac}} = \left( T_1 - T_s \right) \frac{v_{ac}}{v_{ac}} + T_s \frac{v_{ac}}{v_{ac}} / T_s \]  
(28)

\[ \overline{v_{dc}} = \left( T_2 - T_s \right) \frac{v_{dc}}{v_{dc}} + T_s \frac{v_{dc}}{v_{dc}} / T_s \]  
(29)

Even though the average value of \( v_{dc1} \) is the same as that of \( v_{dc2} \), the instantaneous values of them are not the same due to the ac ripple component. This ac ripple component causes the voltage modulation error in (26) and (27). So the instantaneous difference of the two dc voltages should be compensated.

According to the switching state, the output voltages of the inverter are given in Table 1. For example, when the switching state is given as shown in Fig. 9, the output voltages in case of the unequal \( v_{dc1} \) and \( v_{dc2} \) are modified from (28) and (29) as

\[ \overline{v_{ac}} = \left( T_1 - T_s \right) \left( -v_{dc2} \right) + \left( T_2 - T_s \right) v_{dc1} / T_s \]  
(30)

\[ \overline{v_{dc}} = \left( T_2 - T_s \right) \left( -v_{dc2} \right) + \left( T_2 - T_s \right) v_{dc1} / T_s \]  
(31)

For a precise modulation, a voltage component to be compensated can be derived from the difference between the real output voltage and its reference as

\[ v_{comp} = \overline{v_{ac}} - \overline{v_{dc}} = \overline{v_{hc}} - \overline{v_{bc}} \]  
(32)

Including this term in the calculation of the switching time,

\[ T_1 = T_s \frac{v_{ac}}{v_{ac}} + \frac{v_{ac}}{v_{ac}} T_s \]  
(33)

\[ T_2 = T_s \frac{v_{dc}}{v_{dc}} + \frac{v_{dc}}{v_{dc}} T_s \]  
(34)

On the other hand, the dead time is required to prevent the shoot-through of the inverter switching leg. This dead time should be compensated for eliminating the voltage distortion. The dead time effect will be considered in the next section.

V. EXPERIMENT RESULTS

Fig. 10 shows the configuration of the experimental set-up. The source voltage is a single-phase, 110[V], 60[Hz], and the boost inductance and its resistance are 2[mH] and 0.06[\Omega], respectively, and the dc link capacitor is 3300[\mu F]. The switching frequency of the IGBT devices is 5 [kHz] and the dc link voltage is controlled to be 540[V]. The induction motor is of a three-phase, 220[V], 4P, and 3[Hp], which is operated by V/F constant control mode. For load application to the induction motor, the separately-excited dc motor coupled is
operated in torque control mode. The main controller is a DSP TMS320C31.

Fig. 11 shows the estimated source voltage waveform. With the proposed estimation strategy, the source voltage is well performed so that the source voltage sensor can be eliminated.

Fig. 12 shows the source voltage and current waveforms in steady-state. The source current is controlled to be sinusoidal with unity power factor kept for bidirectional power flow operation. Fig. 13 shows the dc-link voltages of $v_{dc_1}$, $v_{dc_1}$, and $v_{dc_2}$. The $v_{dc_1}$ coincides with its reference value of 540[V] with low ripple components. With the balancing control, $v_{dc_1}$ and $v_{dc_2}$ are well balanced.

Fig. 13 shows the induction motor currents at half a load and at 10[Hz] operation. Without any compensation in (a), the motor current is much distorted. In (b) and (c) with the compensation of either the dc-link ripple component or the dead time effect, it is seen that there still exists the distortion of the waveforms. With the compensation for these two effects in (d), the waveform becomes a balanced set of sinusoidal currents. Fig. 14 shows the same results as in Fig. 13 except in operating at 60[Hz]. Since the output voltage at high speeds is larger than that at low speeds, the degree of distortion without proper compensation is less than that at low speeds. However, there still exist some distortion and unbalance. With compensation, the current has been made sinusoidal and balanced. Fig. 15 shows that the transient response of the motor current at impact loading. Due to the compensation, the motor currents are well balanced and sinusoidal in transient state.

VI. CONCLUSIONS

In this paper, a control algorithm for a single-phase half-bridge PWM rectifier and two-leg PWM inverter for three-phase induction motor drives has been studied. Here, adverse influences of a dc link ripple voltage and the dead time of the inverter switching legs on the inverter output current have been investigated and a solution was proposed, which resulted in a balanced set of three-phase output current. Also, a novel algorithm to eliminate the source voltage sensor in the half-bridge PWM rectifier was presented, which proved that the estimated source voltage coincides well with the measured voltage. In addition, a balancing control of the neutral voltage in the dc link was applied for the symmetrical output voltage of the inverter. The validity of the proposed algorithm has been verified by experimental results. The proposed digital control algorithm for the given circuit topology can be applied to the residential appliance, farming, and low power industrial applications with a low cost but high performance DSP as TMS2407.
Fig. 14. Induction motor currents at half a load condition (10 Hz) with compensation for (a) none (b) dc-link ripple voltage (c) dead time effect (d) both.

Fig. 15. Induction motor currents at half a load condition (60 Hz) with compensation for (a) none (b) dc-link ripple voltage (c) dead time effect (d) both.

Fig. 16. Transient response of induction motor currents for load variation. (a) impact loading, (b) induction motor currents

REFERENCES


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