Square-Wave Operation for a Single-Phase-PFC Three-Phase Motor Drive System Without a Reactor

Jun-ichi Itoh and Nobuhiro Ohtani

***Abstract*—This paper proposes a square-wave control strategy for a single-phase power factor correction with a new boost-up converter which uses the leakage inductance of a motor instead of a boost-up reactor. Since the power supply is connected to the neutral point of the motor, the current distortion in the power supply occurs when the inverter outputs square waveforms. First, this paper describes the characteristic of the proposed circuit and the problems in the square-wave operation. Next, a current control method for the square waveform is proposed to suppress the current distortion. Finally, the validity of the proposed converter and its control strategy are demonstrated by experimental results.**

***Index Terms*—Boost-up converter, leakage inductance, neutral point, square-wave operation.**

I. Introduction

RECENTLY, single-phase motors have often been used in consumer electronics. However, in terms of size and weight, it is better to use a three-phase motor instead of a single-phase motor [1]. Single-phase input consumer electronics need a single- to three-phase converter to drive a three-phase motor. The converter needs to achieve high efficiency and small size. Furthermore, an input current is required to meet the harmonic standard [2].

Fig. 1 shows a full-bridge pulsewidth modulation (PWM) rectifier that is commonly applied in a single- to three-phase converter [1], [3], [4]. A PWM rectifier has high efficiency and a good solution for power factor correction (PFC); however, all PFC circuits require a large boost-up reactor in the rectifier, and therefore, the cost is high. Another low-cost structure single- to three-phase converter is composed by a diode-bridge rectifier, a boost chopper circuit, and a three-phase inverter. Although a diode rectifier has low cost, it cannot meet the input current harmonic standard since the input current contains large distortion [5]–[7]. In order to reduce the harmonic, a dc reactor with a diode rectifier is used; however, the dc reactor cannot suppress the harmonic current under a large output power. The boost chopper works as a PFC converter [8], [9].

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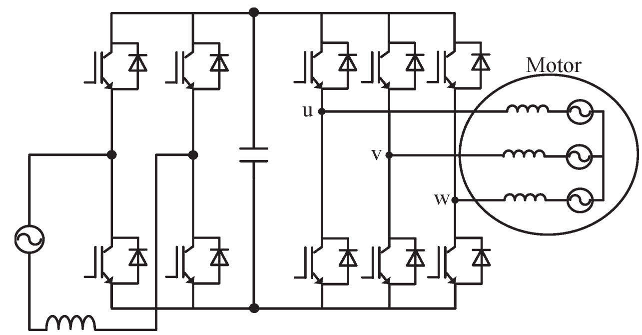


Fig. 1. Conventional full-bridge circuit.

The authors have proposed a method that a leakage inductance of the motor is used instead of a boost-up reactor [9], [10]. Since it consists of a simpler circuit structure than the conventional circuit, the proposed circuit can achieve a smaller size and lower cost application. The motor is controlled by the proposed converter using PWM.

On the other hand, a square-wave operation is popularly used in many applications such as high-speed motor drive, electric vehicle, and rail way system. The square-wave operation of the inverter can increase the voltage utilization and can achieve higher efficiency than a PWM operation since the switching frequency of a square-wave operation is lower than the PWM drive [11]. In the proposed circuit, when the square-wave operation is applied to the inverter control, the voltage of the neutral point of the motor has fluctuation that contains three times of the output frequency. Since the power supply is connected to the neutral point of the motor in the proposed circuit, the input current contains distortion that results from the fluctuation of the neutral point of the motor.

This paper proposes a square-wave operation strategy for the proposed circuit, which uses the neutral point of a motor. The objective of this paper is to suppress the input current distortion under a square-wave operation. The input current distortion is suppressed by an automatic current regulation (ACR) and a feedforward compensation. First, this paper describes the characteristic of the proposed circuit and the principle operation. Thus, when the output frequency is high, it is presented that the dc-link voltage is two times higher than the peak input voltage. Next, this paper discusses the problem of the input current distortion which is caused by the fluctuation of the neutral point of the motor under the square-wave operation. In addition, a high power factor control for suppressing the input current distortion with a square-wave operation is proposed. Finally, the validity of the proposed circuit will be demonstrated by experimental results.

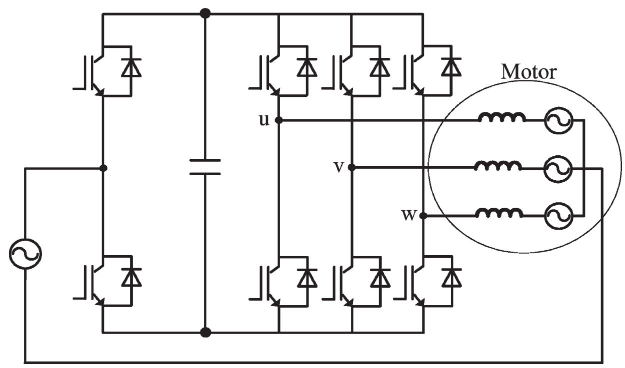


Fig. 2. Proposed circuit using the neutral point motor.

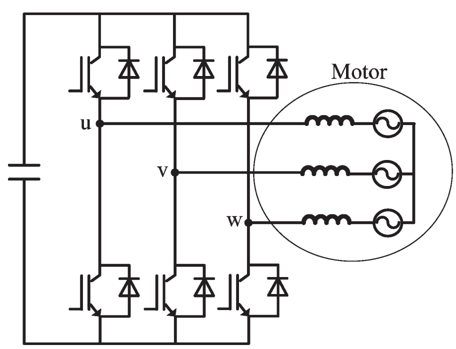


Fig. 3. Equivalent circuit of the positive phase in the proposed circuit using the neutral point of the motor.

II. Reactor-Free PFC Converter

A. Circuit Configuration

Fig. 2 shows the proposed circuit that connects to the neutral point of the motor. The proposed circuit consists of a three-phase inverter, an electrolytic capacitor, and a switching leg using a series insulated-gate bipolar transistor (IGBT). The input power supply is connected between the center of the leg and the neutral point of the motor. The input reactor of the proposed circuit is replaced with the zero-phase inductance, which is also known as the leakage inductance of the motor [9], [10]. The required leakage inductance for the control is varied according to the type of the motors. In an induction motor, a minimum of 3% of the inductance is enough for the proposed control method. On the other hand, in the case of a permanent-magnet motor, a minimum of 10% of the synchronous reactance is sufficient for the proposed control method. Therefore, the input reactor is unnecessary in the proposed circuit. In addition, the number of switching elements can be decreased in comparison to a conventional single-phase full-bridge PWM rectifier and inverter system because the inverter parts in the proposed circuit can substitute as another leg in the rectifier. As a result, the gate drive circuits or peripheral control units in the rectifier can be reduced.

Fig. 3 shows the equivalent circuit of the positive phase sequence in the proposed circuit. In this case, the power supply and the rectifier leg do not appear in the equivalent circuit since it is in zero-phase components. This figure also shows the equivalent circuit of the motor current control which is similar to a conventional three-phase inverter. It is noted that

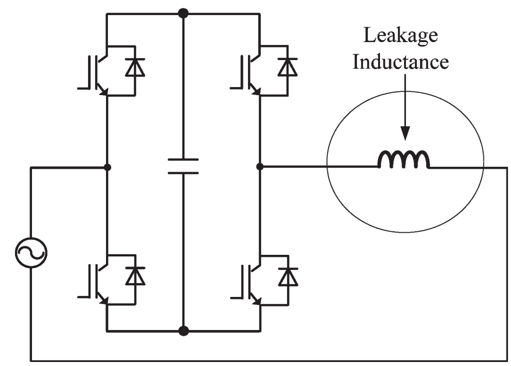


Fig. 4. Equivalent circuit of the zero phase in the proposed circuit using the neutral point voltage.

the zero-phase current in the motor does not generate torque because the zero-phase flux denies each other. The control for the input current of the proposed circuit is achieved by a zero-phase component of the output voltage, and the control for motor current is achieved by a positive phase at the same time. The conduction loss in the rectifier side can be greatly reduced because of using two IGBTs. However, note that the motor loss will be increased subsequently since the output current contains of the rectifier current during the zero-phase sequence.

B. Series–Parallel Compensation

PWM Operation: The proposed circuit is initially controlled with a PWM drive and delivers sinusoidal output waveforms. Then, a transition control is added to the controller to transform the output voltage from sinusoidal wave to square wave while operating at high output frequency. In PWM drive, the zero-phase sequence occurs at the zero vector periods of the inverter side. Fig. 4 shows the zero-phase sequence equivalent circuit. The motor is equaled to a leakage inductance, and the inverter can be considered as a single-leg topology, where this equivalent circuit is similar to a conventional single-phase full-bridge circuit topology. The back electromotive force (EMF) does not appear in the zero-phase equivalent circuit but only the leakage inductance occurs. It is noted that the EMF includes multiple third-order harmonics; those harmonic components appear in the zero-phase equivalent circuit.

The inverter output voltages v_u, v_v, and v_w are expressed in the following:

$$\begin{cases} v_u = a\frac{E_{\mathrm{dc}}}{2}\sin\omega t + v_o \\ v_v = a\frac{E_{\mathrm{dc}}}{2}\sin(\omega t - \frac{2\pi}{3}) + v_o \\ v_w = a\frac{E_{\mathrm{dc}}}{2}\sin(\omega t - \frac{4\pi}{3}) + v_o \end{cases} \tag{1}$$

where a is the modulation index of the inverter phase voltage, $0 < a < 1$, v_o is the neutral point voltage, and ω is the inverter output angler frequency.

Square-Wave Operation and DC-Link Voltage Analysis: A square-wave operation for the three-phase inverter is applied to reduce the switching loss of the inverter in comparison to a PWM. During the square-wave operation, the switching frequency agrees with the output frequency. The output line

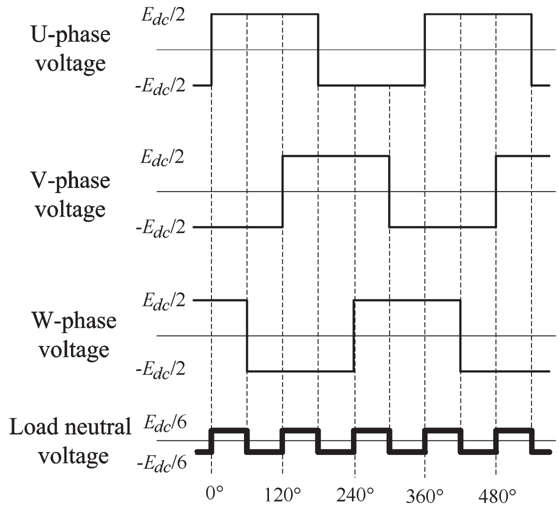


Fig. 5. Relations between each of the phase voltage and the load neutral point.

voltage becomes a 120° square waveform, and then, the fundamental voltage of the output is given by

$$V_{\mathrm{out}} = \frac{\sqrt{6}}{\pi} E_{\mathrm{dc}} \tag{2}$$

where V_{out} is the output line voltage and E_{dc} is the dc-link voltage.

Fig. 5 shows the relations of the phase voltage and the neutral point voltage during the square-wave operation. When the inverter is operated at the square wave, the neutral point of the motor has a fluctuation that is plus or minus one-sixth of the dc-link voltage. As a result, each of the phase voltage becomes the square wave of ± one-sixth of the dc-link voltage.

In the proposed circuit, the neutral point of the motor is connected to the power supply. That is, the voltage fluctuation of the neutral point disturbs the current control of the power supply. Subsequently, the input current contains distortion that is three times of the output frequency. When the output voltage of the inverter leg fluctuates with ± one-sixth of the dc-link voltage, the rectifier leg has to output the sum of ± one-sixth of the dc-link voltage and peak voltage of the input power supply. Therefore, in order to compensate the disturbance by the square-wave operation, the dc-link voltage is constrained by

$$\begin{aligned} \frac{E_{\mathrm{dc}}}{2} &\geq \frac{1}{6} E_{\mathrm{dc}} + \sqrt{2} V_{\mathrm{in}} \\ E_{\mathrm{dc}} &\geq 3\sqrt{2} V_{\mathrm{in}} \end{aligned} \tag{3}$$

where E_{dc} is the dc-link voltage and V_{in} is the power supply voltage.

On the other hand, if the frequency of the fluctuation in the neutral point of the motor is higher than the input current control response, the voltage fluctuation can be neglected. Then, the dc-link voltage requires only two times of the power supply voltage because the control of the rectifier is the same as the half-bridge PWM rectifier.

Fig. 6 shows a control block diagram of the current control in order to discuss the condition of the required dc-link voltage. A proportional–integral (PI) controller is used as a current regulator, and the neutral point of the motor is considered as a voltage disturbance. When the influence of the disturbance voltage can be disregarded for the input current, the dc-link voltage may be twice of the peak voltage of the input voltage. The transfer function from a disturbance voltage to the input current is obtained by

$$\frac{I_{\mathrm{in}}}{V_{\mathrm{dis}}} = s\frac{T_i}{K_p} \times \frac{1}{s^2 \frac{LT_i}{K_p} + sT_i + 1} \times \sqrt{\frac{3}{2}} \tag{4}$$

where K_p is the proportional gain, T_i is the integration time constant of the PI regulator, L is the leakage inductance, and s is the Laplace operator.

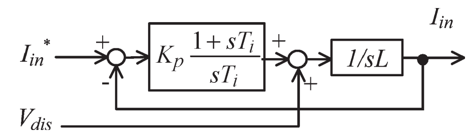


Fig. 6. Control block diagrams of the current control.

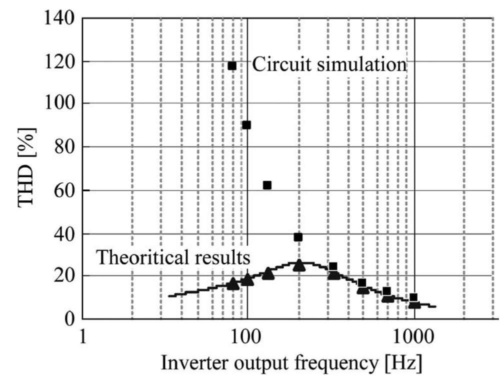


Fig. 7. Relations between the input current THD and inverter output frequency.

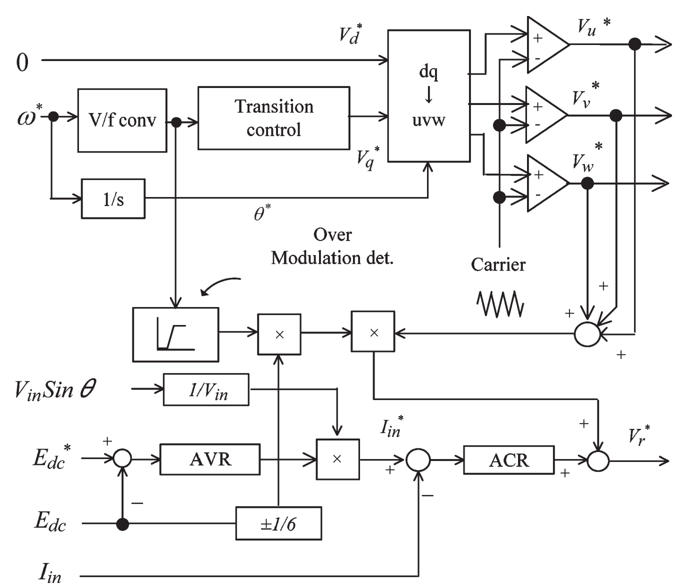
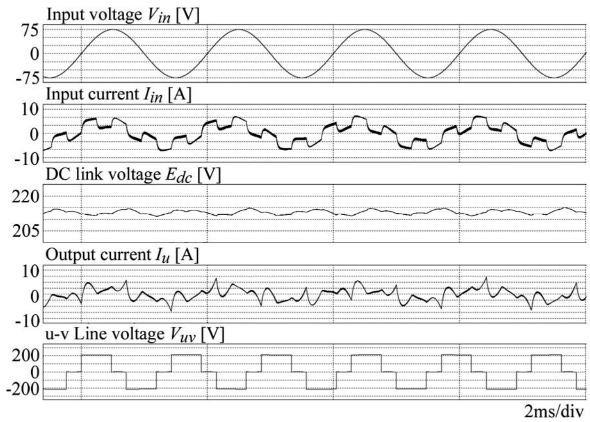
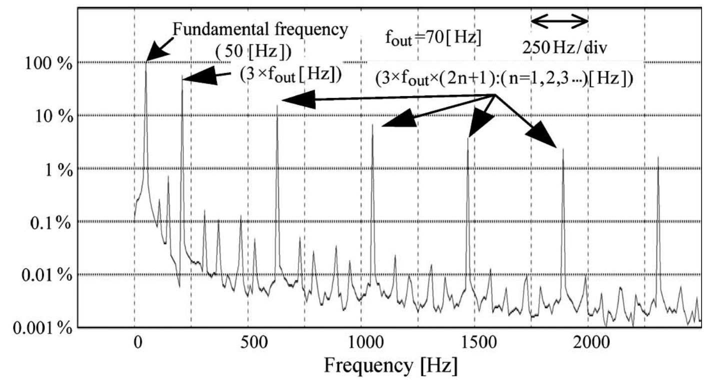


Fig. 8. Block diagram of the proposed circuit.

Fig. 7 shows the relations between the input current total harmonic distortion (THD) and the inverter output frequency. The natural angular frequency ω_n is 4000 rad/s, and the damping factor ζ is $1/\sqrt{2}$. It is noted that the disturbance frequency



(a)



(b)

Fig. 9. Simulation result without the proposed control. (a) Simulation waveform. (b) Harmonic analysis results of the input current.

is three times of the inverter output frequency and the dc-link voltage is set to twice of the peak voltage of the input voltage. The symbol "▲" shows the theoretical curve calculated by (4), and the symbol "■" shows the current THD calculated by a circuit simulator. When the inverter frequency is low, THD increases in the simulation results because the disturbance voltage cannot be neglected in the current controller. That is, the dc-link voltage is not enough to control the current. In order to overcome the problem, the dc-link voltage is required to have at least two times or more than the input peak voltage. In this condition, the limit of the inverter output frequency is set to 330 Hz. That is, before 330 Hz, the required dc-link voltage is three times of the input peak voltage, and after that, the dc-link voltage has to increase according to the inverter output frequency. In this paper, the dc-link voltage is set to three times higher than the input peak voltage with an output frequency at 70 Hz in the simulation and also tested experimentally with an output frequency of 44 Hz.

III. Proposed Control Strategy

Fig. 8 shows the block diagram of the proposed circuit. V_d^* and V_q^* are the $d{-}q$-axes in the three-phase transformation, respectively. V_u^*, V_v^*, and V_w^* are the pulse patterns for inverter side, and V_r^* is the pulse pattern for the rectifier. The current control of the power supply is the same as the conventional PWM rectifier. It is noted that the leakage inductance is used instead of the boost reactor; however, the volume of the inductance decreases to one-third because the leakage inductance of the motor is connected in parallel in the zero-phase equivalent circuit.

As discussed in the previous chapter, the neutral point voltage of the motor has a fluctuation of $\pm 1/6E_{\mathrm{dc}}$ with three times of the inverter output frequency in the square-wave operation. In order to compensate for the voltage fluctuation, a feedforward control is applied to the ACR. The feedforward signal is obtained from the sum of the inverter pulse patterns and adjusted in a reverse polarity of the actual fluctuation waveform at the neutral point of the motor. This reverse signal is then added with the estimated dc-link voltage and sent into the rectifier control to cancel off the fluctuation waveform. Note that the control can distinguish between the PWM and square-wave modulation from the control of the "Over modulation det." The feedforward control sends zero signal to the rectifier control while PWM is operating.

For the inverter side, the PWM control is slowly changed into a square-wave control via the trapezoidal pulse modulation. The transition control block shown in Fig. 8 is the control of V_q^* axis with regard to the output frequency. The V_q^* axis will be linearly extended into the overmodulation region where the amplitude of output voltage reference is nonlinearly increasing, following the modulation index. As a result, the sinusoidal wave of the output voltage will be turned into a square wave completely when the V_q^* axis has finished extending in the overmodulation area.

The PI controller is designed according to (4). That is, the natural angular frequency ω_n is set to 4000 rad/s, and the damping factor ζ is set to $1/\sqrt{2}$.

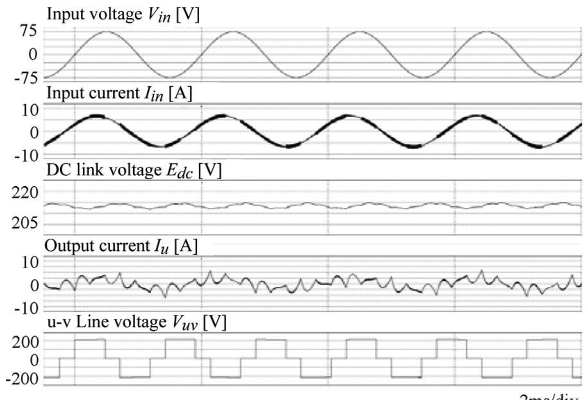
IV. Simulation Results

Fig. 9 shows the simulation results during a square-wave operation. The motor model is expressed in the EMF, the leakage inductance, and the armature resistance. The input voltage is 50 V, the input frequency is 50 Hz, and the output frequency is 70 Hz. From Fig. 9, it is confirmed that the input current contains the third-harmonic components of the output frequency. The THD of the input current is 56.5% in Fig. 9.

Fig. 10 shows that the distortion of the input current is suppressed by the proposed control. The THD of the input current THD is 1.3%. In contrast to Fig. 10, the distortion of the input current is suppressed drastically by the feedforward compensation.

V. Experimental Results

Fig. 11 shows the experimental results where the inverter uses the PWM operation mode. The input voltage, input current, line current, and $u - v$ line-to-line voltage were examined. The parameters of the induction motor used in the experiment are shown in Table I. The input voltage is 50 V and 50 Hz. In Fig. 11, the output frequency of the positive phase is 44 Hz. A sinusoidal input current waveform is obtained because the



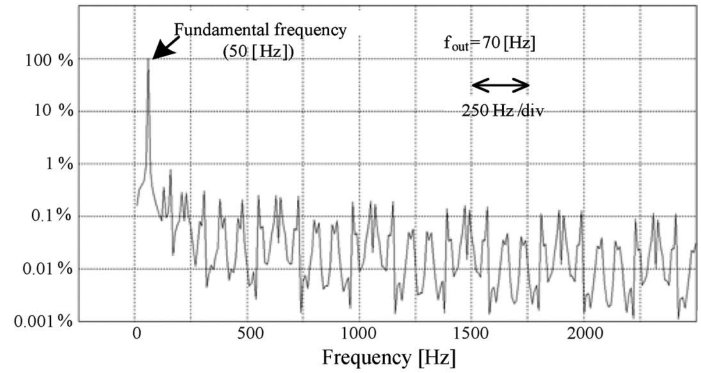
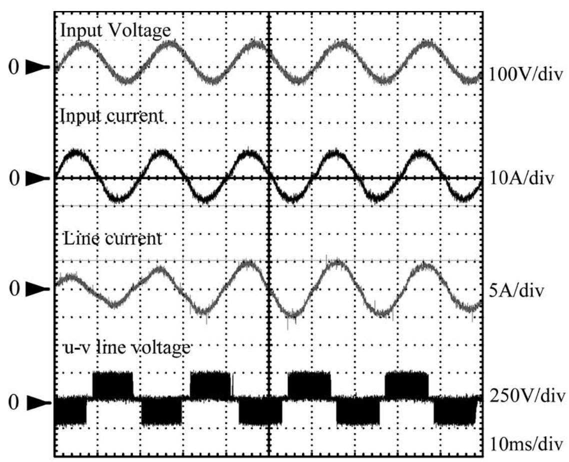


Fig. 10. Simulation result with the proposed control. (a) Simulation waveform. (b) Harmonic analysis results of the input current.



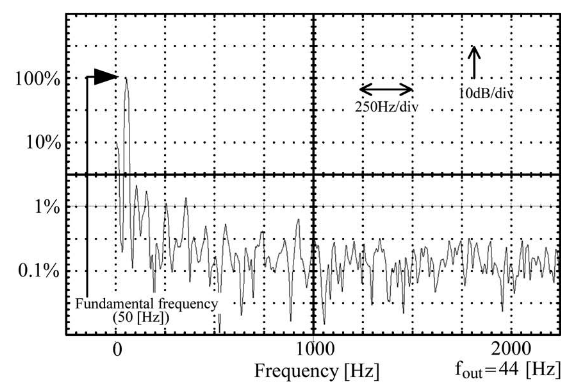


Fig. 11. Experimental results using the inverter in the PWM operation mode. (a) Operation waveform. (b) Harmonic analysis results of the input current.

TABLE I
MOTOR PARAMETERS (FUJI: MLH6085M)

Motor Power	750 W
Poles	4/50 Hz
RPM	1420
Rated current	3.6 A
Rated voltage	200 V
Leakage inductance	4.42 mH

neutral point of the motor voltage is constant during the PWM operation. The THD of the input current is 3.9%. The amplitude of the line current is changed because the line current consists of an input current of 50 Hz and a positive current of the motor of 44 Hz.

Fig. 12 shows the acceleration characteristics from the PWM operation to square-wave operation. The PWM waveform is gradually changed into a square-wave control that can be confirmed by the $u - v$ line voltage. The pulse mode moves from PWM to the square mode without rush current. It is noted that the fluctuation of the line current is generated due to the difference between the input and output frequencies.

Fig. 13 shows the experimental results during a square-wave operation. In Fig. 13, the harmonic distortion of the input current is suppressed, and the dc-link voltage is kept constant that can be confirmed by the $u - v$ line voltage. In addition, the input current is confirmed to contain the third-harmonic components of the output frequency. The THD of the input current is 29.1%. It is noted that the output current includes ripple due to the square-wave operation.

Fig. 14 shows that the distortion of the input current is suppressed by the proposed control. The THD of the input current THD is 6.1%. In comparison with Fig. 13, the distortion of the input current is suppressed to less than 1/20 times. In addition, the THD of the input current is 23% reduced. The input current harmonics of the proposed control method meet the standard of IEC61000-3-2.

Overall, the experimental results confirmed the validity of the proposed feedforward compensation control method.

VI. CONCLUSION

A control strategy for a reactor-free converter has been proposed to apply in a square-wave operation with an adjustable-speed drive motor. The known problem is where the current distortion in the power supply occurs when the inverter outputs square waveforms. In order to overcome the problem, a feed-forward control has been proposed to suppress the input current distortion with a square-wave operation. The proposed control method successfully suppresses the input current distortion to

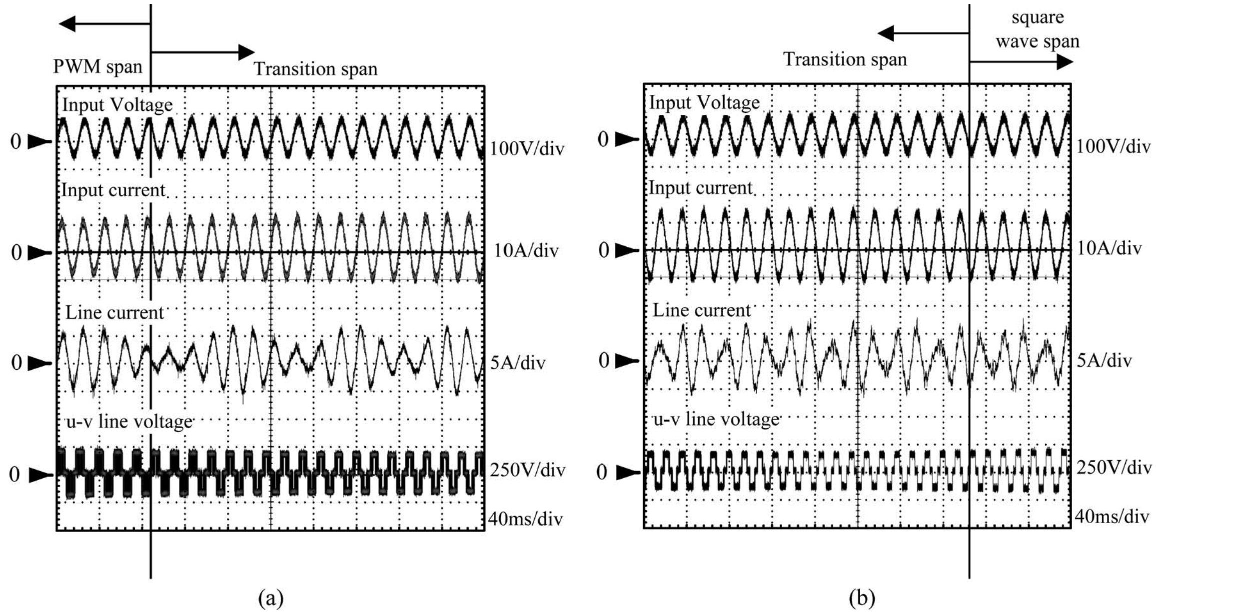


Fig. 12. Acceleration characteristics. (a) Operation waveform from PWM transition span. (b) Operation waveform from span to transition span to square wave span.

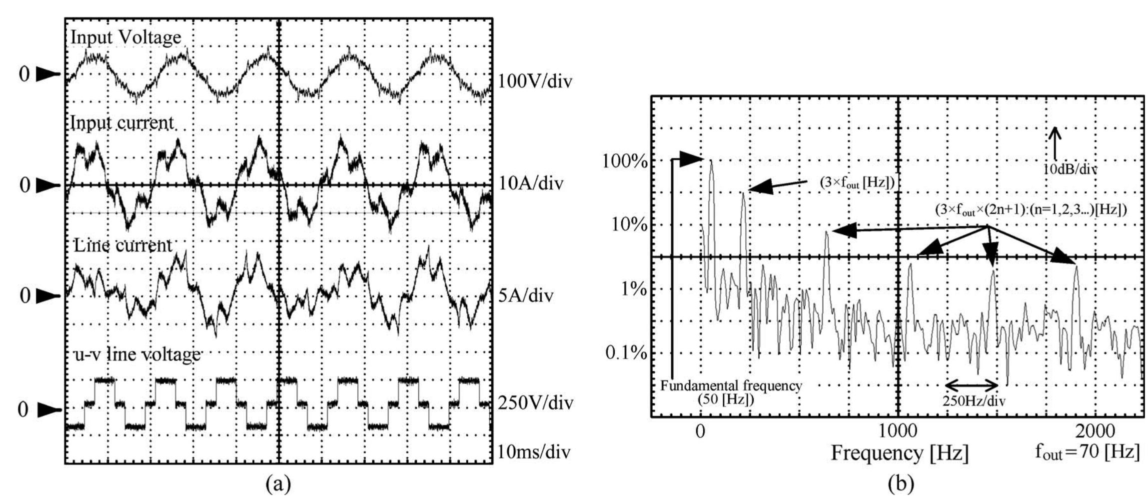


Fig. 13. Experimental results without the proposed control. (a) Operation waveform. (b) Harmonic analysis results of the input current.

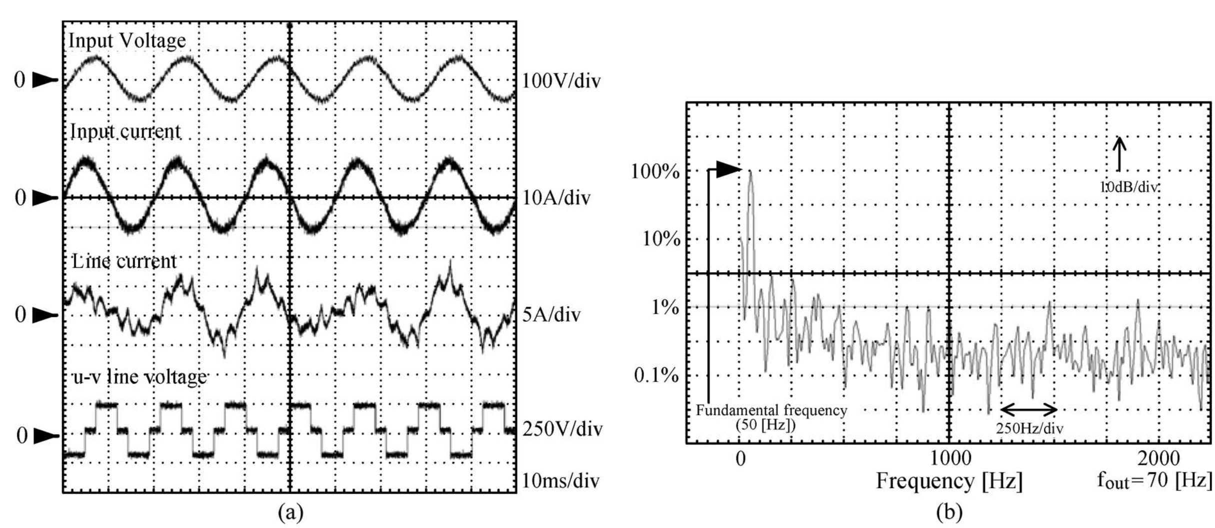


Fig. 14. Experimental results with the proposed control. (a) Operation waveform. (b) Harmonic analysis results of the input current.

1/20 than the ordinary control. In addition, the THD of the input current is reduced by 23%.

In future works, the proposed converter and controller will be applied and examined in a high-speed permanent-magnet motor.

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