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A New Direct Single Phase To Three Phase Converter For Powering Ac Motors

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Abstract – This paper presents a new approach to convert an available single-phase power directly to a three-phase power suitable for powering an AC motor. A control strategy is developed so that a balanced three-phase voltage is provided across the motor terminals and the gain between the fundamental output voltage and the input voltage of the converter is maximized. Voltage and frequency at the converter output can be adjusted by a PWM technique for variable speed operation of the motor load. For high power operation the four-step switching technique is applied to the proposed scheme for safe commutation of inductive load current from one bidirectional switch to another. The proposed approach does not employ any reactive components and snubber circuits across the bidirectional switches, and hence is robust and is small in size and weight. Simulation results verify key concepts. Finally, experimental results are provided from a laboratory proto-type converter fed induction motor drive from a single phase supply.

1. Introduction

It has been known that three-phase electrical equipment such as three-phase induction motors are significantly more efficient, economical and easy to control than their single-phase counterpart. Also, the size of three-phase motors are significantly smaller than its single phase version with the same kVA rating [1]. However, in many cases three-phase power is rarely available in the rural or light industrial areas, due to the high cost of extending three-phase service [1]. So the only available power in these areas is in the form of single-phase and there exists a need for phase converters for powering induction motor driven fans, pumps, and other irrigation equipment.

Some approaches have been proposed to convert a single-phase power to three-phase power in reference [1-5]. These systems are either rotary or static and some uses electromechanical or reactive elements to achieve phase conversion. And most of the early approaches produce balanced voltage only at one specific load condition. The later approaches employ static converters with rectifier, dc-link and inverter stage.

In this paper, a new approach to directly convert an available single-phase power to a three-phase power suitable for induction motor drives is proposed (Fig. 1). The proposed converter employs four bi-directional semiconductor switches to define two voltages across the three terminals of the motor load without employing any reactive components. Two anti-series MOSFET's or IGBT's are employed to form one bi-directional switch. For high power operation the four-step switching technique[6] is applied to the proposed scheme. The bi-directional switch pairs are suitably phase controlled and/or pulse width modulated to respect the three phase laws. A control strategy is provided to maximize the gain between the fundamental output voltage and the input voltage. Variable speed operation of the proposed approach is also analyzed.

The advantage of the proposed converter are,

- Conversion from single phase to three phase is directly performed without any intermediate stage (No energy storage components are used).
- Variable speed operation of induction motor is possible with the proposed approach.
- Four-step switching strategy of the bi-directional switches results in smooth commutation of load current (No snubbers are employed).
- Improved input power factor and current harmonics

The resulting system is compact, small in size and weight.

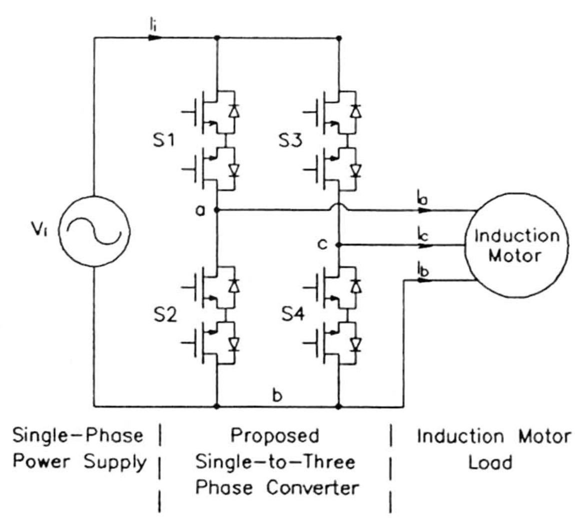

Fig. 1 The proposed phase converter to power three phase induction motor from a single phase supply

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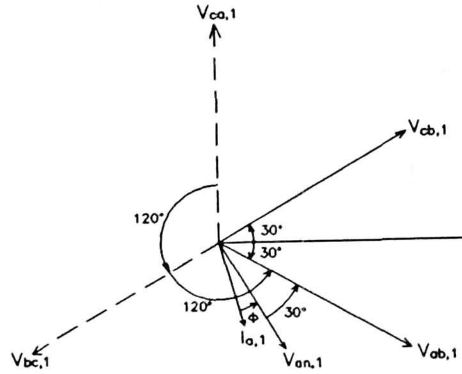


Fig. 2 Vector diagram of the converter

The followings are the disadvantages of the proposed converter, and possible solutions to overcome these are,

- At rated output (60Hz) the voltage gain for the fundamental component from the input and output is 63%. This necessitates derating of the motor. However, if a lower voltage motor is powered from an available single phase higher voltage this disadvantage can be overcome. Alternatively, an autotransformer can be employed to boost the input supply.
- Harmonic currents in the motor contribute to increased heating and possible torque pulsations. The motor impedance suitably limits the harmonic current magnitudes, and therefore slight derating is acceptable and not seen as a disadvantage particularly in low performance fan and pump systems.

2. Principles of Operation

The proposed converter configuration is shown in Fig.1. Bi-directional switch pair S1 and S2 defines voltage V_{ab} and S3 and S4 define voltage V_{cb}. Then, voltage Vca is defined by the phase relationship of $V_{ca} = V_{cb} - V_{ab}$. Fig. 2 shows the vector diagram of the input voltage and the line to line output voltages. It can be seen from Fig. 2 that three output voltages V_{ab},1, V_{cb},1, and V_{ca},1 are balanced. This ensures the generation of the rotating magnetic field in the induction motor.

2.1 Calculation of Firing Angles α and β

Switch pair S1 and S2 is phase controlled with a delay angle α (Fig. 3b) so that the fundamental component of V_{ab}, V_{ab},1, laggs input voltage V_i by 30 degree.

Line to line output voltage V_{ab} (Fig. 3b) is defined by,

$$V_{ab}(\omega t) = \begin{cases} 0 & (S1\ is\ on)\ \ 0 \le \omega t < \alpha,\ \pi \le \omega t < \pi + \alpha \\ V_m \sin(\omega t) & (S2\ is\ on)\ \ 0 \le \omega t < \pi,\ \pi + \alpha \le \omega t < 2\pi \end{cases} \quad (1)$$

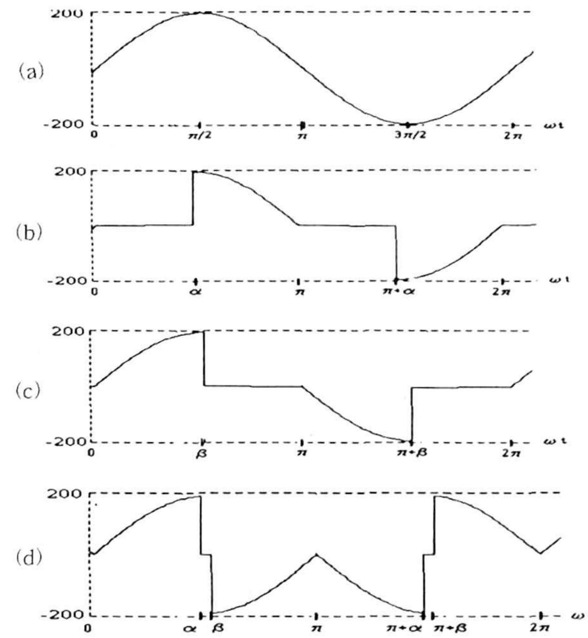


Fig. 3 (a) Input voltage waveform V_i
(b) Converter output voltage waveform V_{ab}
(c) Converter output voltage waveform V_{cb}
(d) Converter output voltage waveform V_{ca}

The Fourier series representation of V_{ab} is given by,

$$V_{ab}(\omega t) = A_0 + \sum_{n=1}^{\infty} (A_n \sin(n\omega t) + B_n \cos(n\omega t)) \quad (2)$$

where,

$$A_0 = 0$$

$$A_n = \frac{V_m}{\pi}\left(\frac{\sin(n+1)\alpha}{n+1} - \frac{\sin(n-1)\alpha}{n-1}\right) \quad n \ge 3,\ \text{odd}$$

$$B_n = \frac{V_m}{\pi}\left(\frac{\cos(n+1)\alpha - 1}{n+1} - \frac{\cos(n-1)\alpha - 1}{n-1}\right) \quad n \ge 3,\ odd \quad (3)$$

Then, the phase angle of V_{ab},1, $\varphi_{ab,1}$, becomes,

$$\varphi_{ab,1} = \tan^{-1}\left(\frac{B_1}{A_1}\right) = \tan^{-1}\left(\frac{\cos(2\alpha) - 1}{\sin(2\alpha) + 2(\pi - \alpha)}\right) \quad (4)$$

For the three-phase balanced operation, we need

$$\varphi_{ab,1} = 30^{\circ} \quad (5)$$

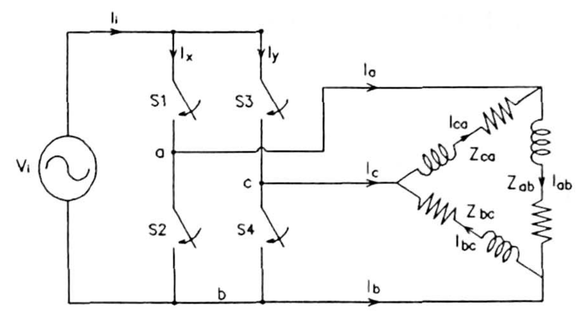


Fig. 4 The circuit notation used in the analysis

Solving for α from equations (4) and (5) yields,

$$\alpha = 85.7^\circ \tag{6}$$

Similarly, line to line voltage V_{cb} (Fig. 2c) is defined by,

$$V_{cb}(\omega t) = \begin{cases} V_m \sin(\omega t) & (S3\ is\ on)\ \ 0 \le \omega t < \beta,\ \pi \le \omega t < \pi + \beta \\ 0 & (S4\ is\ on)\ \ \beta \le \omega t < \pi,\ \pi + \beta \le \omega t < 2\pi \end{cases} \tag{7}$$

In the same manner the phase angle of the fundamental component of V_{cb} is given by,

$$\varphi_{cb,1} = \tan^{-1}\left(\frac{B_1}{A_1}\right) = \tan^{-1}\left(\frac{1-\cos(2\beta)}{2\beta - \sin(2\beta)}\right) \tag{8}$$

β can be obtained by setting $\varphi_{cb,1} = 30^\circ$

$$\beta = 94.3^\circ \tag{9}$$

The magnitude of the fundamental component of V_{ab} can be obtained by,

$$|V_{ab,1}| = \sqrt{A_1^2 + B_1^2} = 0.633 * V_m \tag{10}$$

Similiarly, the magnitude of the fundamental component of V_{cb} is also

$$|V_{cb,1}| = 0.633 * V_m \tag{11}$$

This result illustrates that setting the angles α and β appropriately to produce leading and lagging vectors by 30° from the reference vector yields the three phase balanced output vectors.

2.2 Analysis of Input Current Ii

We extend our analysis to get input current I_i as follows. Assuming the balanced three-phase load in Fig. 4, it can be

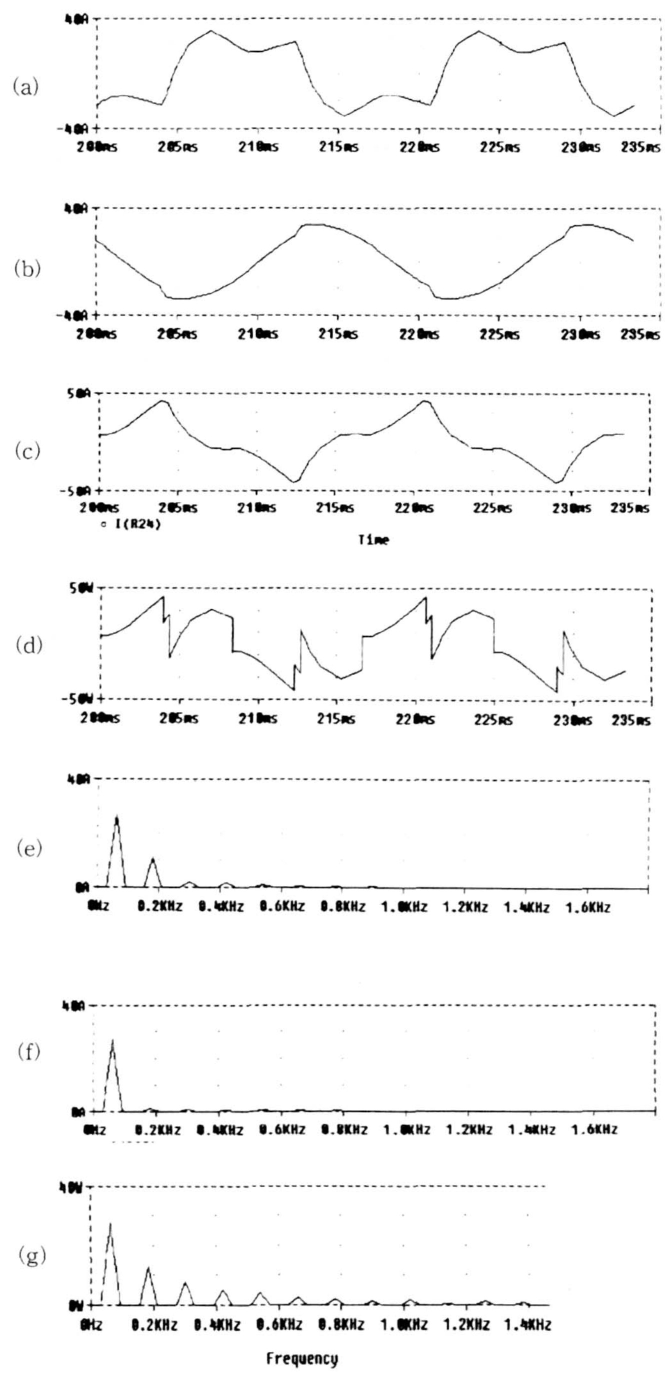


Fig. 5 Simulation results (a) output current I_a, (b) I_b, (c) I_c, (d) input current I_i, (e) frequency spectra of currents I_a and I_c, (f) frequency spectrum of I_b, (g) frequency spectrum of I_i

seen that

$$I_a = I_{ab} - I_{ca} = \frac{V_{ab} - V_{ca}}{Z} \tag{12}$$

$$I_c = I_{ca} - I_{bc} = \frac{V_{ca} + V_{cb}}{Z} \tag{13}$$

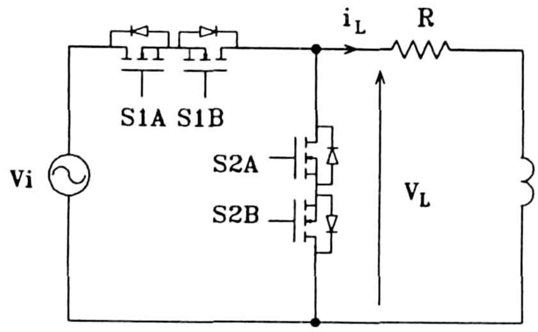


Fig. 6 Simplified circuit illustrating the four-step switching of the proposed converter

and

$$I_x = I_a \cdot F_\alpha = \frac{(V_{ab} - V_{ca}) \cdot F_\alpha}{Z} \tag{14}$$

$$I_y = I_c \cdot F_\beta = \frac{(V_{ca} + V_{cb}) \cdot F_\beta}{Z} \tag{15}$$

where F_α and F_β are the switching function for S1 and S3, respectively and Z is the load impedance. Then the input current I_i can be expressed as,

$$I_i = I_x + I_y = \frac{(V_{ab} - V_{ca}) \cdot F_\alpha + (V_{ca} - V_{cb}) \cdot F_\beta}{Z} \tag{16}$$

Substituting the Fourier series of V_{ab}, V_{cb}, V_{ca}, F_α and F_β into (16) yields,

$$I_i = \sum_{n=1,3,5,\ldots}^{\infty} \frac{K_n}{Z_n} \sin(n\omega t - \phi_{L,n}) \tag{17}$$

where K_n is the magnitude coefficients, $|Z_n| = \sqrt{R^2 + (n\omega L)^2}$ is the load impedance magnitude and $\varphi_{L,n} = \tan^{-1}(\frac{n\omega L}{R})$ is the load phase angle for n-th component. Equation (17) illustrates that the input power factor directly correspond with the load power factor .

3. Simulation Results

Fig. 3 and Fig. 5 illustrate the simulation results on PSPICE for the circuit topology shown in Fig. 4. Fig. 3 (a) to (d) show the converter input and output voltages, respectively. Fig. 5 (a) to (c) show converter output currents I_a, I_b and I_c, respectively. Fig. 5 (d) shows the wave shape of input current I_i. Fig. 5(e) to (f) illustrate the frequency spectra of the currents.

4. Four-Step Switching Technique

Fig. 6 shows the simplified circuit drawn for one leg of the proposed converter. Each bidirectional switch is composed of two unidirectional switches such as MOSFETs connected in anti-series. Now, suppose S1A or S1B is conducting the load current. If an attempt is made to turn off S1A and S1B and turn on S2A and S2B at the same time, the commutation problem dicussed in [6] will be encountered. To overcome this situation, the four-step switching strategy is applied to the proposed single-phase to three-phase converter.

A cycle of sinusoidal input voltage can be divided into four modes according to the direction of inductive load current i_L. These are,

Mode 1 : Vi > 0 and $i_L > 0$
Mode 2 : Vi > 0 and $i_L < 0$
Mode 3 : Vi < 0 and $i_L > 0$
Mode 4 : Vi < 0 and $i_L < 0$

The four-step switching technique according to each mode is described below,

Mode 1 : $V_i > 0$ and $i_L > 0$ (S1A is conducting)

1. Turn on S2B - Nothing happens since S2B is reverse biased.
2. Turn off S1A - i_L is transferred from S1A to S2B.
3. Turn on S2A - Nothing happens.
4. Turn off S1B - Nothing happens.

Mode 2 : $V_i > 0$ and $i_L < 0$ (S1B is conducting)

1. Turn on S2B - Nothing happens since S2B is reverse biased.
2. Turn off S1A - Nothing happens.
3. Turn on S2A - i_L is transferred from S1B to S2A.
4. Turn off S1B - Nothing happens.

Mode 3 : $V_i < 0$ and $i_L > 0$ (S1A is conducting)

1. Turn on S2A - Nothing happens since S2A is reverse biased.
2. Turn off S1B - Nothing happens.
3. Turn on S2B - i_L is transferred from S1A to S2B.
4. Turn off S1A - Nothing happens.

Mode 4 : $V_i < 0$ and $i_L < 0$ (S1B is conducting)

1. Turn on S2A - Nothing happens since S2A is reverse biased.
2. Turn off S1B - i_L is transferred from S1B to S2A.
3. Turn on S2B - Nothing happens.
4. Turn off S1A - Nothing happens.

A small delay (1 or 2μsec) is introduced between each of the switching steps. Note that Mode 1 and Mode 2 have identical switching sequences and Mode 3 and Mode 4 also do. This switching method is also applied to swich pair S3 and S4 in the same manner.

5. Variable Speed Operation

The proposed converter (Fig. 1) can be controlled to generate a variable-frequency, variable-voltage three-phase output for adjustable speed operation of the induction motor drive. Since for variable speed operation V/f should remain constant, at lower frequencies it is necessary to generate lower voltage. Thus the converter does not suffer from a voltage derating. Further, at lower frequencies the converter switches can be operated with PWM control to facilitate voltage control.

Let ω_i = input frequency (60Hz)

ω_s = converter modulation frequency

$\omega_o = \omega_s - \omega_i$ = converter output fundamental frequency

and single-phase input voltage be

$$V_i(\omega_i t) = V_m \cos(\omega_i t) \tag{18}$$

$A_i \cos(\omega_s t - 30°)$ is employed as a low frequency modulating function for trianglular carrier wave. The resulting PWMed pulses are employed as a gating signal for switches S1 and S2. A similar modulating function $A_i \cos(\omega_s t + 30°)$ is used to control S3 and S4. The switching function $F_\alpha(\omega_s t)$ (gating signal for S1 and S2) can be represented as,

$$F_\alpha(\omega_s t) = A_o + A_1 \cos(\omega_s t - 30°) + \sum_{n=2}^{\infty} A_n \cos[n(\omega_s t - 30°)] \tag{19}$$

Therefore the voltage $V_{ab}(\omega_o t)$ now becomes,

$$\begin{aligned} V_{ab}(\omega_0 t) &= V_i(\omega_i t) * F_\alpha(\omega_s t) \\ &= \frac{1}{2} A_1 V_m \cos(\omega_o t - 30°) + A_0 V_m \cos(\omega_i t) \\ &+ \frac{1}{2} A_1 V_m \cos(\omega_s t + \omega_i t - 30°) \\ &+ \sum_{n=2}^{\infty} A_n V_m \cos[n(\omega_s t - 30°)] \cos(\omega_i t) \end{aligned} \tag{20}$$

Similarly, for switch pair S3 and S4

$$F_\beta(\omega_s t) = A_o + A_1 \cos(\omega_s t + 30°) + \sum_{n=2}^{\infty} A_n \cos[n(\omega_s t + 30°)] \tag{21}$$

and

$$\begin{aligned} V_{cb}(\omega_0 t) &= V_i(\omega_i t) * F_\beta(\omega_s t) \\ &= \frac{1}{2} A_1 V_m \cos(\omega_o t + 30°) + A_0 V_m \cos(\omega_i t) \\ &+ \frac{1}{2} A_1 V_m \cos(\omega_s t + \omega_i t + 30°) \\ &+ \sum_{n=2}^{\infty} A_n V_m \cos[n(\omega_s t + 30°)] \cos(\omega_i t) \end{aligned} \tag{22}$$

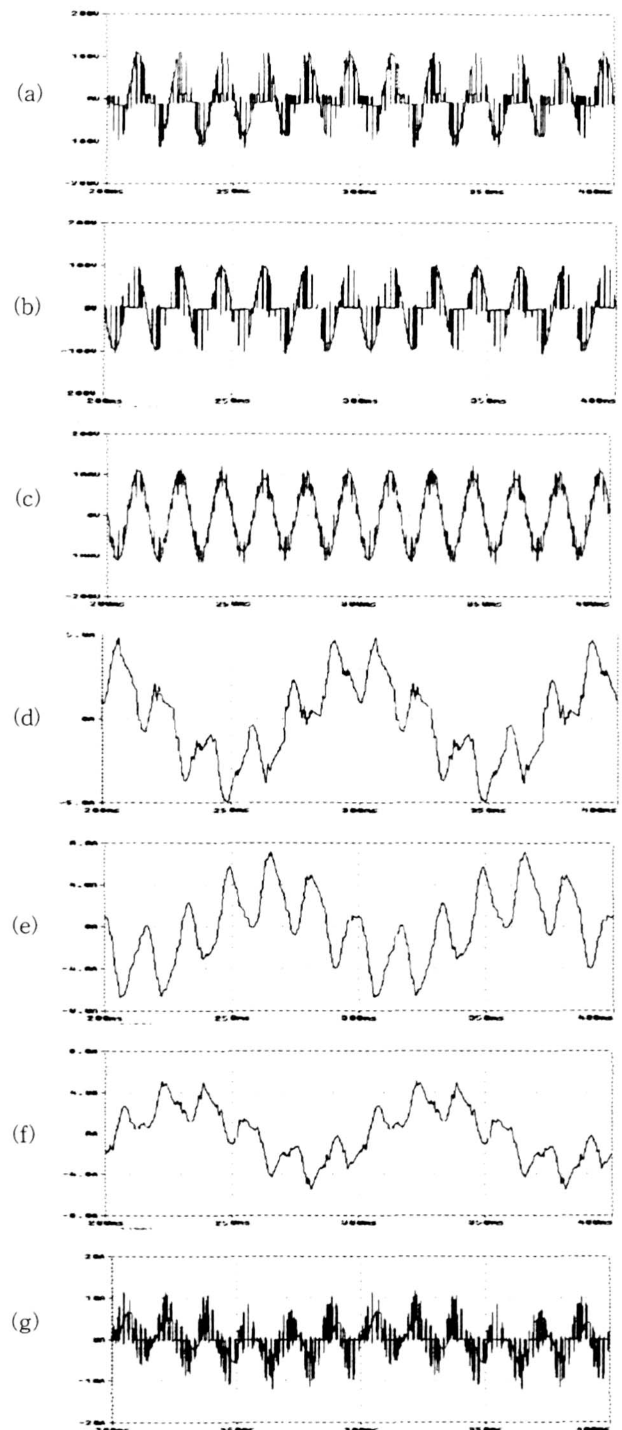
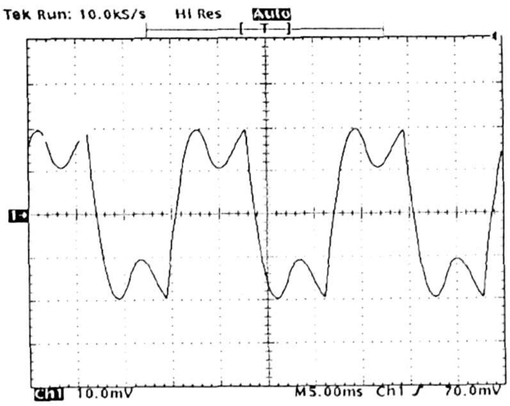
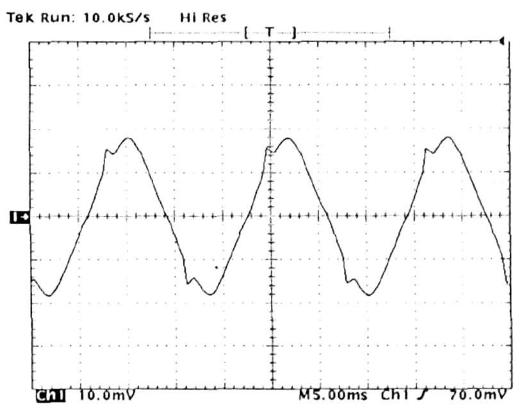


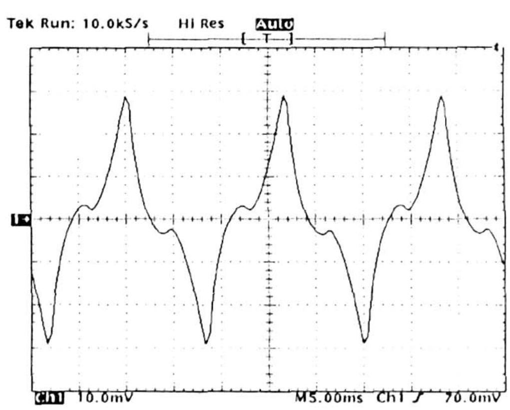
Fig. 7 Output voltage waveforms (a) V_{ab}, (b) V_{cd}, (c) V_{ca}, Output current waveforms (d) I_a, (e) I_b, (f) I_c, Input current waveform (g) I_i



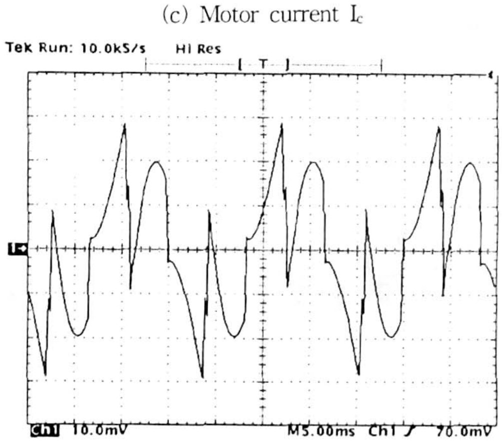
(a) Motor current I_a



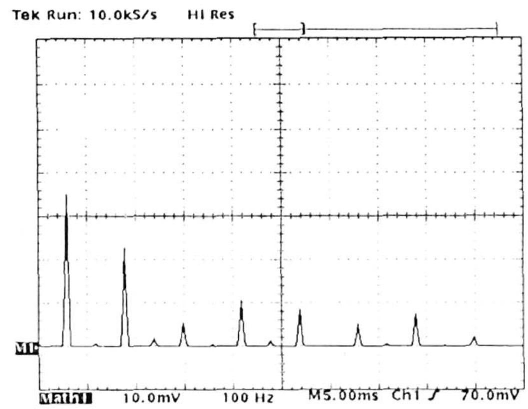
(b) Motor current I_b



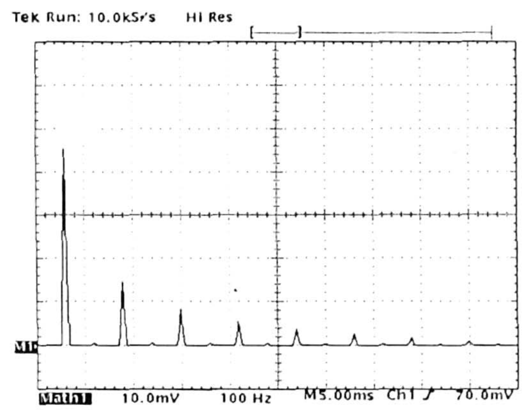
(c) Motor current I_c



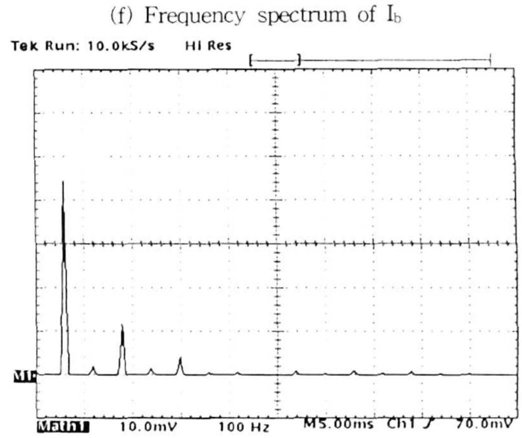
(d) Input current I_i



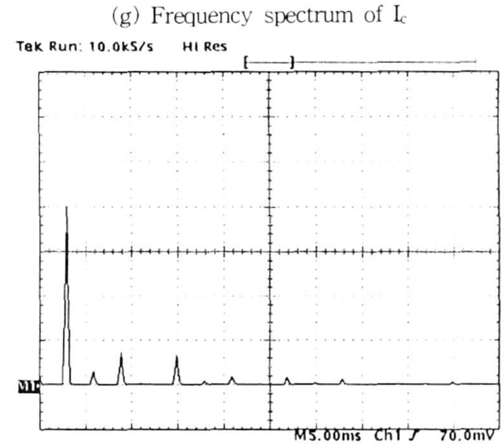
(e) Frequency spectrum of I_a



(f) Frequency spectrum of I_b



(g) Frequency spectrum of I_c



(h) Frequency spectrum of I_i

Fig. 8 Experimental results

For example, f_o= 10Hz is required and f_i= 60Hz then the modulating frequency becomes f_s= f_o+f_i = 70Hz from Eqns. (19) and (21). The above described control strategy results in controllable variable frequency output. The output voltage is adjusted by varying A_1. Fig. 7(a) to (c) show output voltage V_{ab}, V_{cb} and V_{ca} for f_o = 10Hz. Fig. 7(d) to (f) illustrate respective currents I_a, I_b and I_c. Finally Fig. 7(g) shows input current I_i.

6. Experimental Results

To demonstrate the feasibility of the proposed single to three phase converter, a prototype converter which drives a three phase induction motor was implemented. Fig. 8 shows the waveforms of the output currents I_a, I_b, I_c and the input current I_i and their respective frequency spectra. It can be seen that the harmonic current magnitudes are limited by the motor leakage inductance.

7. Conclusion

A new approach to directly convert an available single-phase power to a three-phase power suitable for powering induction motor has been proposed. The proposed converter does not employ any reactive components and snubbers, and hence is robust and is small in size and weight. Voltage and frequency at the converter output can be adjusted by a PWM technique for variable speed operation of the motor. For high power operation the four-step switching technique is applied for safe commutation of inductive load current from one bidirectional switch to another. Simulation results verified key concepts. Experimental results were provided from a laboratory proto-type converter fed induction motor drive from a single phase supply.

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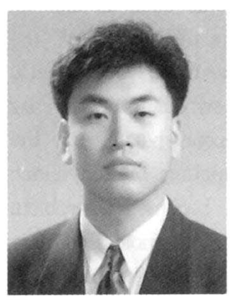

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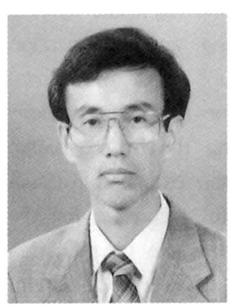

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