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# A New Current Phasor-Controlled ZVS Twin Half-Bridge High-Frequency Resonant Inverter for Induction Heating

Tomokazu Mishima, *Member, IEEE*, Chikanori Takami, *Non-Member, IEEE*, and Mutsuo Nakaoka, *Member, IEEE*

**Abstract**—A novel soft-switching high-frequency (HF) resonant (HF-R) inverter for induction heating (IH) applications is presented in this paper. By adopting the current phasor control of changing a phase shift (PS) angle between two half-bridge inverter units, the IH load resonant current can be regulated continuously under the condition of wide range soft-switching operations. In addition to this, the dual mode power regulation scheme based PS angle control & asymmetrical pulse-width-modulation (PWM) in one inverter unit is proposed for improving the efficiency in low output power settings. The essential performances on the output power regulation and soft-switching operations are demonstrated in an experiment using its 1 kW-60 kHz HF-R inverter prototype, then the topological validity is evaluated from a practical point of view.

**Index Terms**—induction heating (IH), high-frequency (HF) resonant (HF-R) inverter, zero voltage soft-switching (ZVS), current phasor, phase shift (PS) angle control, asymmetrical PWM.

## I. INTRODUCTION

IH has been applied widely in a variety of electric power conversion processing such as home appliances (IH cookers), fluid heating systems, induction hardening, and superheated steamer system owing to its unique and practicable properties of direct and local heating, high efficiency and safety.

As an example of up-to-date IH technologies, the developments of HF inverters for IH metallic pans/vessels fabricated from low resistivity and low permeability materials such as aluminum and copper are gathering much attention in pursuing for energy saving home appliances amid global warming preventions. This type of IH process demands HF inverters operating in high efficiency, higher switching frequency and minimized switching power losses [1]–[10].

In order to attain the reduction of switching loss, low Electro-Magnetic Interference (EMI) noise and high power density effectively, the introduction of soft-switching technologies is useful in the HF-R inverter. The soft-switching HF-R

inverter which have been developed so far have attractive features such as the low cost and simple control schemes based on pulse-frequency-modulation (PFM) and PWM. However, the HF-R inverters controlled by PFM have the inherent technical issue, i.e. switching frequency limitation for the low-medium output power settings, thereby the wide range power regulation of the IH load cannot be ensured [3]. In addition, improvement of power density cannot be expected in the asymmetrical half-bridge HF inverter without the complicated power regulation schemes [5][6], and so in the phase shift PWM (PS-PWM) full-bridge (FB) inverters [11]–[15] because of the narrow soft-switching range. In particular, the conventional ZVS PS-PWM FB HF inverter as depicted in Fig. 1 suffers from severe soft-switching limitation in the controlled-phase active switches under the condition of low output power settings. The dual half-bridge inverter suitable for the coupled working coils is proposed in [4], but performances on the soft switching operations are not clearly demonstrated. Therefore, improvement of the power density is a significantly technical challenge for those previous type HF inverters while constraining the switching frequency.

As a HF inverter operating with higher switching frequency conditions, the triplex resonant frequency (TRF) inverter has been developed for a 2 kW-60 kHz IH cooker as reported in [1]. The 60 kHz TRF inverter can be actually utilized without increasing the switching power losses by adopting one-third switching frequency 20 kHz. However, the drawback of the TRF HF inverter is that the fundamental component of the load resonant current is inherently eliminated and not transferred to the IH load. Accordingly, power conversion efficiency might deteriorate in the large and rated output power area, thereby high efficiency as performed in the commercial HF inverters can not be expected.

As a new solution for the technical problem, an innovative soft-switching HF-R inverter is proposed in this paper. In the HF-R inverter proposed herein, ZVS operations can be attained over the wide range of output power variations together with a seamless power regulation by varying the PS angle between the two instantaneous resonant currents of the half-bridge inverter units, which is named as "current phasor control" in this research. Therefore, the soft-switching operations can be maintained even in the low output power conditions, consequently the reductions of switching power losses and EMI noises can be realized effectively for the wide-range output power setting. Moreover, the proposed HF-R inverter

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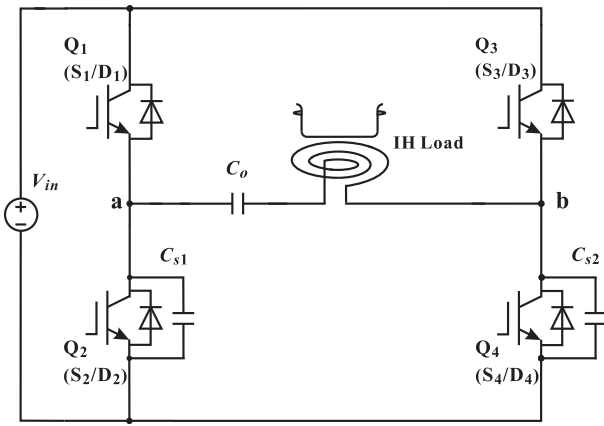


Fig. 1. Conventional ZVS PS-PWM FB HF inverter topology.

can operate as a single asymmetrical half-bridge inverter by exchanging the numbers of operating half-bridge inverter units. Thereby, high efficiency and high reliability power conversions can be expected in the low output power area of the proposed HF-R inverter.

This paper is presented for demonstrating and evaluating the soft-switching operations and steady-state power regulation characteristics of the newly-proposed soft-switching HF-R inverter in an experiment using a 1 kW-60 kHz prototype, then its practical effectiveness is originally clarified.

The remainder of this paper is organized as follows. The circuit configuration and its operation principle of the proposed ZVS HF-R inverter are described in Section II. An analysis of the output power regulation based on the current phasor control is demonstrated by using the simplified equivalent circuit of the proposed HF-R inverter in Section III. Then, the HF-R inverter characteristics of the wide-range power regulation are theoretically clarified in the same section. The experimental verifications on the soft-switching operation, wide-range and seamless power regulation are carried out by using the 1 kW-60 kHz prototype in Section IV. In addition to these, the validity of the dual-mode control scheme for improving the conversion efficiency in the low output power setting is actually evaluated in the experiment. Furthermore, performance comparison between a conventional ZVS PS-PWM FB HF inverter and the proposed HF-R inverter with the single and dual mode controls are demonstrated in terms of soft-switching range and conversion efficiency. Finally, the effectiveness of the proposed HF-R inverter is demonstrated from a practical point of view.

## II. A NEW ZVS HF RESONANT INVERTER

### A. Circuit Description

A schematic diagram of the proposed HF-R inverter is illustrated in Fig.2. The HF-R inverter treated herein is composed of the two asymmetrical half-bridge inverters  $U_1$  and  $U_2$  which share the IH load. The IH load which consist of the IH working coil and a metallic utensil (pan, vessel) are represented by the equivalent effective resistance  $R_o$  and equivalent effective inductance  $L_o$ . The resonant and power factor correction (PFC) tuned capacitor  $C_o$  is connected in series with the IH load [16].

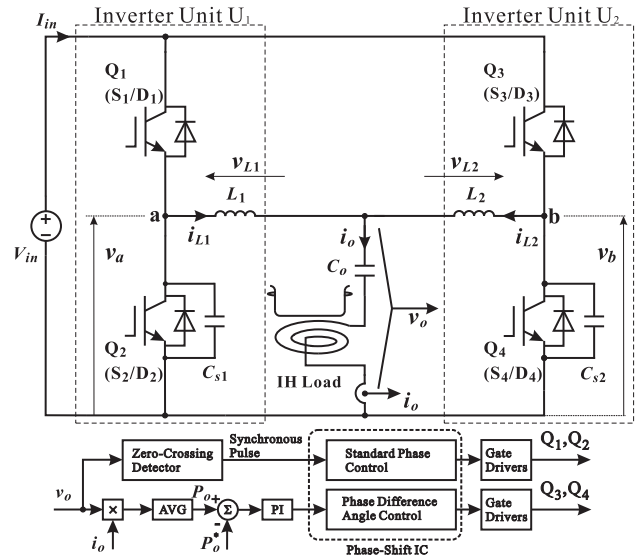


Fig. 2. A proposed current phasor-controlled ZVS HF-R inverter.

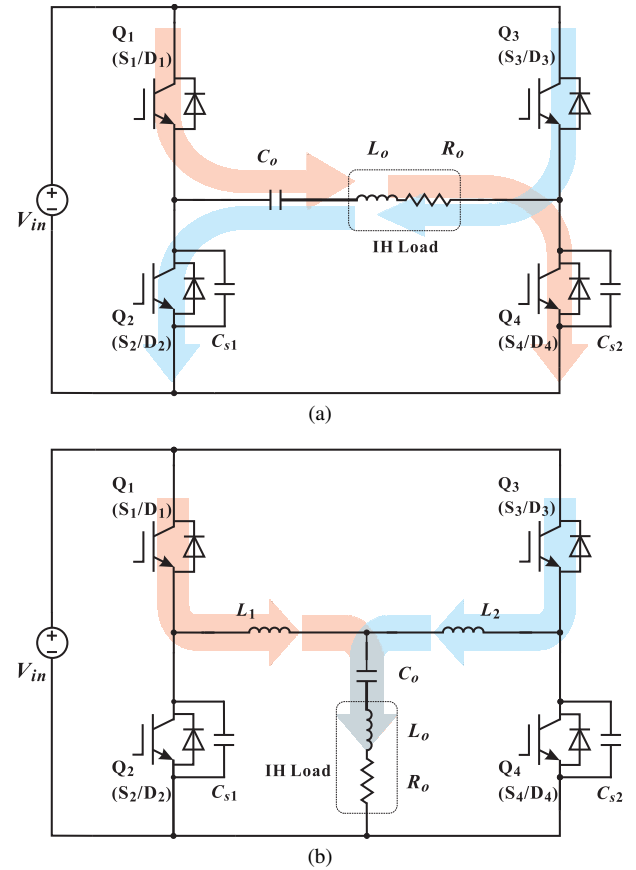


Fig. 3. Current pathway for IH load at full load (PS angle  $\phi = 0^\circ$ ): (a) PS-PWM FB HF inverter, (b) proposed ZVS HF-R inverter.

The two inverters  $U_1$  and  $U_2$  produce the two resonant link currents  $i_{L1}$  and  $i_{L2}$  respectively, which are controlled by changing the PS angle  $\phi$  in the IGBT gate pulse signals between  $Q_1$ - $Q_2$  and  $Q_3$ - $Q_4$  [17] [18]. Synthesizing the resonant link currents  $i_{L1}$  and  $i_{L2}$  yields the load current  $i_o$  with a lower harmonic distortion. Accordingly, the output power

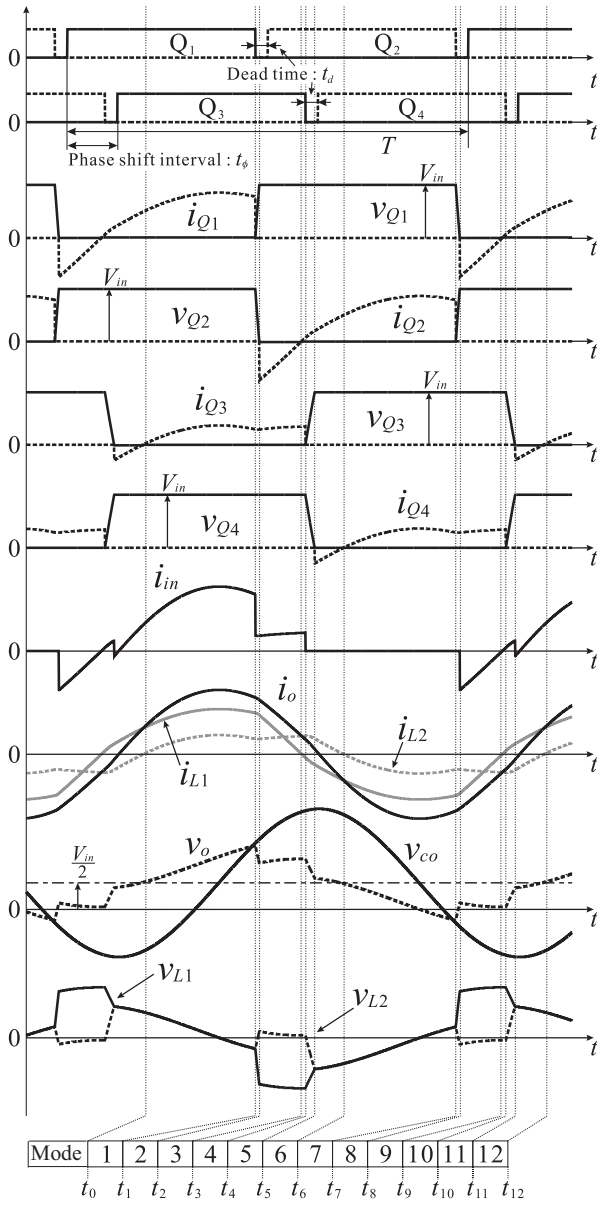


Fig. 4. Relevant voltage and current operating waveforms ( $\phi = \omega t_\phi = 2\pi f_s t_\phi$ ).

injected into the IH working coil and load can be continuously controlled over the wide range of output power variations under the soft-switching condition. Since the IH load current is resulted by the two resonant link inductor currents, the HF inverter topology treated herein is suitable for increasing the power rating while reducing the conduction power losses, especially for the large and rated output power conditions.

The active switches  $Q_1$ - $Q_2$  and  $Q_3$ - $Q_4$  are commutated in ZVS due to the edge resonance by the lossless capacitors  $C_{s1}$  and  $C_{s2}$  with the aids of  $L_1$  and  $L_2$ . The switching frequency  $f_s (= \omega / 2\pi)$  of the two HF inverters can be designed and fixed to the value which is less than the inverter resonant frequency  $f_r$  defined by

$$f_r = \frac{1}{2\pi\sqrt{LC_o}}, \quad (1)$$

$$L = L_o + \frac{L_1 \cdot L_2}{L_1 + L_2}. \quad (2)$$

Since  $U_1$  and  $U_2$  are fundamentally designed to have the same power rating,  $L_2$  is equal to  $L_1$ . Then, the inverter resonant switching frequency  $f_r$  is rewritten as

$$f_r = \frac{1}{2\pi\sqrt{LC_o}}, \quad (3)$$

$$L = L_o + \frac{L_r}{2}, \quad (4)$$

where  $L_r = L_1 = L_2$ .

As compared to the conventional ZVS PS-PWM FB HF inverter which is illustrated in Fig. 1, the proposed HF-R inverter has the advantages as

- The current rating of active switches can be reduced theoretically by half as compared to the ZVS PS-PWM FB HF inverter under the condition of same input dc voltage as illustrated in Fig. 3. Accordingly, the HF-R inverter proposed herein is effective for decreasing the conduction power losses in middle-heavy load power setting.
- No soft-switching limitation exists in the controlled-phase active switches, so minimization of switching power losses as well as the low EMI noise can be guaranteed for the wide-range output power variations.

### B. Operation Principle

The operating waveforms of the proposed HF-R inverter are depicted in Fig. 4. The output power regulation is carried out by adjusting the PS angle  $\phi$  (PS interval  $t_\phi$ ), which can be defined by

$$\phi = \frac{t_\phi}{T} \times 360^\circ. \quad (5)$$

The switch-mode transitions and the resonant currents pathways in the proposed HF-R inverter are illustrated in Fig. 5. The one switching cycle operation is divided into the twelve sub-modes as follows:

**[Model 1 :  $t_0 \leq t < t_1$ ] <positive half-cycle steady-state power delivering mode>:** The resonant link currents  $i_{L1}$ ,  $i_{L2}$  flow from the input dc power source  $V_{in}$  to the IH load through  $S_1$  of  $Q_1$  and  $S_3$  in  $Q_3$ .

**[Model 2 :  $t_1 \leq t < t_2$ ] < $Q_1$  ZVS turn-off mode>:** The gate signal to  $S_1$  of  $Q_1$  is removed at  $t = t_1$ . Then, the voltage  $v_{Q1}$  across  $Q_1$  rises gradually by charging  $C_{s1}$ , while the voltage  $v_{Q2}$  across  $Q_2$  declines with a certain slope from  $V_{in}$ . Thus, the ZVS turn-off operation begins in  $Q_1$ . During this interval, the inductive energy stored in the resonant link inductor  $L_1$  should be greater than the capacitive energy of  $C_{s1}$  for completing ZVS operation due to the edge resonance in  $L_1$  and  $C_{s1}$ . Thus, ZVS condition in  $Q_1$  and  $Q_2$  is defined by

$$\frac{1}{2}L_1 i_{L1}(t_1)^2 > \frac{1}{2}C_{s1} V_{in}^2. \quad (6)$$

Furthermore, the voltage  $v_{Q1}$  across  $Q_1$  during its turn-on transition interval is expressed by

$$v_{Q1}(t) = Z_{q1} i_{L1}(t_1) \sin \omega_{q1}(t - t_1) \approx Z_{q1} i_{L1}(t_1)(t - t_1) \quad (7)$$

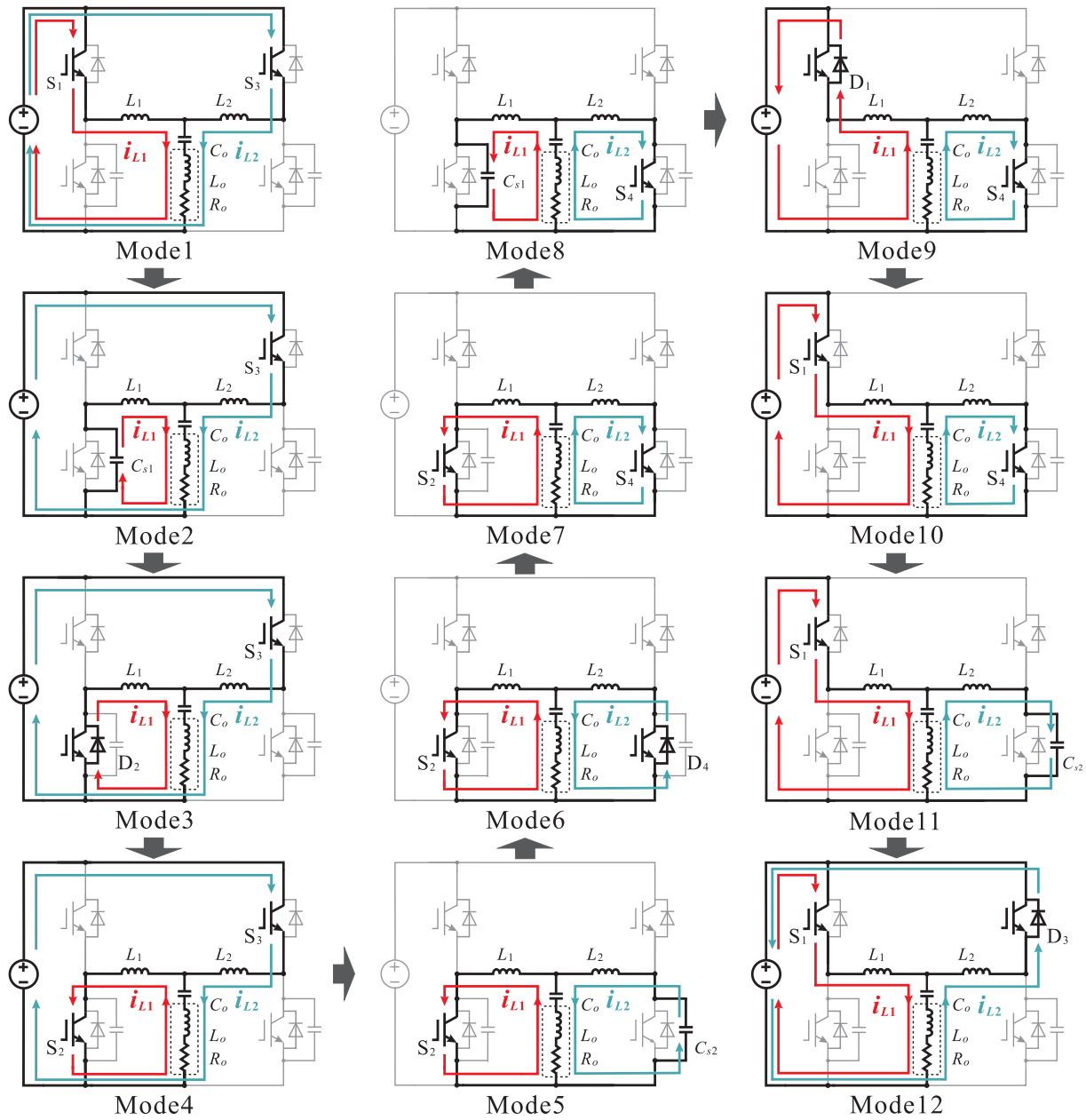


Fig. 5. Mode transitions during one switching cycle ( $\phi = \omega t_\phi = 2\pi f_s t_\phi$ ).

where  $Z_{q1} = \sqrt{L_1/C_{s1}}$  and  $\omega_{q1} = 1/\sqrt{L_1 C_{s1}}$ .

**[Mode3 :  $t_2 \leq t < t_3$  <Q<sub>2</sub> Zero voltage and Zero Current Soft-Switching (ZVZCS) turn-on mode>:** The voltage  $v_{Q1}$  across Q<sub>1</sub> reaches  $V_{in}$  and the ZVS turn-off operation of Q<sub>1</sub> is competed at  $t = t_2$ , while the voltage  $v_{Q2}$  across Q<sub>2</sub> decreases to zero. Then, the anti-parallel diode D<sub>2</sub> in Q<sub>2</sub> is naturally forward-biased, thereby ZVZCS turn-on can be achieved in Q<sub>2</sub>.

**[Mode4 :  $t_3 \leq t < t_4$  <U<sub>1</sub> resonant link current reversely-circulating mode>:** The current through Q<sub>2</sub> commutates from D<sub>2</sub> to S<sub>2</sub> at  $t = t_3$  owing to the series resonance in U<sub>2</sub>. Accordingly, the resonant link current  $i_{L1}$  in U<sub>1</sub> circulates through L<sub>1</sub>, S<sub>2</sub> and the IH load with C<sub>o</sub>. In this transition, the current  $i_{L1}$  and  $i_{L2}$  have the reverse directions each other. This transition contributes for the output power regulation in the proposed HF-R inverter.

**[Mode5 :  $t_4 \leq t < t_5$  <Q<sub>3</sub> ZVS turn-off mode>:** The gate signal to S<sub>3</sub> of Q<sub>3</sub> is removed at  $t = t_4$ . Then, the voltage  $v_{Q3}$  across Q<sub>3</sub> rises gradually by charging C<sub>s2</sub>, while the voltage  $v_{Q4}$  across Q<sub>4</sub> declines with a gradient from  $V_{in}$ . Thus, ZVS turn-off operation begins in Q<sub>3</sub>. During this interval, ZVS condition deriving from edge resonance by L<sub>2</sub> and C<sub>s2</sub> is defined in a similar way to (6) as

$$\frac{1}{2}L_2 i_{L2}(t_4)^2 > \frac{1}{2}C_{s2} V_{in}^2. \quad (8)$$

Furthermore, the voltage  $v_{Q3}$  across Q<sub>3</sub> during its turn-off transition interval is expressed by

$$v_{Q3}(t) = Z_{q2} i_{L2}(t_4) \sin \omega_{q2}(t - t_4) \approx Z_{q2} i_{L2}(t_4)(t - t_4), \quad (9)$$

where  $Z_{q2} = \sqrt{L_2/C_{s2}}$  and  $\omega_{q2} = 1/\sqrt{L_2 C_{s2}}$ .

**[Mode6 :  $t_5 \leq t < t_6$  <Q<sub>4</sub> ZVZCS turn-on mode>:** The voltage



$v_{Q3}$  across  $Q_3$  reaches  $V_{in}$  at  $t = t_5$  and the ZVS turn-off operation of  $Q_3$  is completed, while the voltage  $v_{Q4}$  across  $Q_4$  decreases down to zero. Then, the anti-parallel diode  $D_4$  in  $Q_4$  is naturally forward-biased, thereby ZVZCS turn-on can be achieved in  $Q_4$ .

**[Mode7 :  $t_6 \leq t < t_7$  <negative half-cycle steady-state power delivering mode>]:** The resonant link current  $i_{L2}$  in  $U_2$  naturally commutates from  $D_4$  to  $S_4$  at  $t = t_6$ . Then, the two resonant link currents  $i_{L1}, i_{L2}$  flow through the IH load in the same direction.

**[Mode8 :  $t_7 \leq t < t_8$  < $Q_2$  ZVS turn-off mode>]:** The gate signal to  $S_2$  of  $Q_2$  is removed at  $t = t_7$ . Then, the voltage  $v_{Q2}$  across  $Q_2$  rises gradually by charging  $C_{s1}$ , while the voltage  $v_{Q1}$  across  $Q_1$  declines with a certain slope from  $V_{in}$ . Thus, ZVS turn-off operation begins in  $Q_2$ . During this interval, ZVS condition due to the edge resonance in  $L_1$  and  $C_{s1}$  is defined by

$$\frac{1}{2}L_1 i_{L1}(t_7)^2 > \frac{1}{2}C_{s1} V_{in}^2. \quad (10)$$

Furthermore, the voltage  $v_{Q2}$  across  $Q_2$  during its turn-off transition interval is expressed by

$$v_{Q2}(t) = Z_{q1} i_{L1}(t_7) \sin \omega_{q1}(t - t_7) \approx Z_{q1} i_{L1}(t_7)(t - t_7). \quad (11)$$

**[Mode9 :  $t_8 \leq t < t_9$  < $Q_1$  ZVZCS turn-on mode>]:** The voltage  $v_{Q2}$  across  $Q_2$  reaches  $V_{in}$  at  $t = t_8$  and the ZVS turn-off operation is completed in  $Q_2$ , while the voltage  $v_{Q1}$  across  $Q_1$  decreases to zero. Then, the anti-parallel diode  $D_1$  in  $Q_1$  is naturally forward-biased, thereby ZVZCS turn-on can be achieved in  $Q_1$ . During this interval, a part of the IH load current is fed back to the input dc voltage source  $V_{in}$  via the resonant link inductor  $L_1$ .

**[Mode10 :  $t_9 \leq t < t_{10}$  < $U_2$  resonant link current reversely-circulating mode>]:** The current through  $Q_1$  commutates from  $D_1$  to  $S_1$  at  $t = t_9$  owing to the series resonance in  $U_1$ . Then, the power begins to be delivered to the IH load through the resonant link inductor  $L_1$ , while the resonant link current  $i_{L2}$  in  $U_2$  circulates through  $L_2$ ,  $S_4$  and the IH load with  $C_o$ . In this transition, the resonant link currents  $i_{L1}$  and  $i_{L2}$  have the reverse directions each other. This transition contributes for the output power regulation as well as Mode 4.

**[Mode11 :  $t_{10} \leq t < t_{11}$  < $Q_4$  ZVS turn-off mode>]:** The gate signal to  $S_4$  of  $Q_4$  is removed at  $t = t_{10}$ . Then, the voltage  $v_{Q4}$  across  $Q_4$  rises gradually by charging  $C_{s2}$ , while the voltage  $v_{Q3}$  across  $Q_3$  declines slowly from  $V_{in}$ . Thus, ZVS turn-off begins in  $Q_4$ . During this interval, ZVS condition due to the edge resonance by  $L_2$  and  $C_{s2}$  is defined by

$$\frac{1}{2}L_2 i_{L2}(t_{10})^2 > \frac{1}{2}C_{s2} V_{in}^2. \quad (12)$$

Furthermore, the voltage  $v_{Q4}$  across  $Q_4$  during its turn-off transition interval is expressed by

$$v_{Q4}(t) = Z_{q2} i_{L2}(t_{10}) \sin \omega_{q2}(t - t_{10}) \approx Z_{q2} i_{L2}(t_{10})(t - t_{10}). \quad (13)$$

**[Mode12 :  $t_{11} \leq t < t_{12}$  < $Q_3$  ZVZCS turn-on mode>]:** The voltage  $v_{Q4}$  across  $Q_4$  reaches  $V_{in}$  at  $t = t_{11}$  and the ZVS turn-off operation is completed in  $Q_4$ , while the voltage  $v_{Q3}$  across  $Q_3$  decreases to zero. Then, the anti-parallel diode  $D_3$  in  $Q_3$  is naturally forward-biased, thereby ZVZCS turn-on can be achieved in  $Q_3$ .

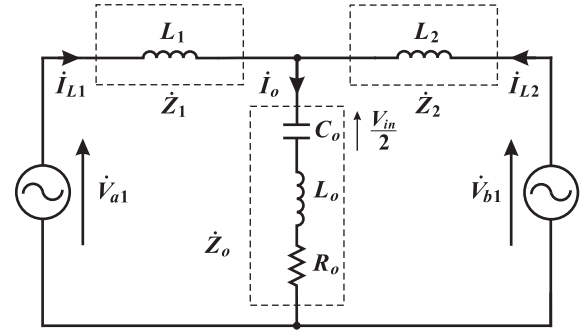


Fig. 6. Frequency-domain equivalent circuit based on two dc-biased fundamental frequency ac voltage sources.

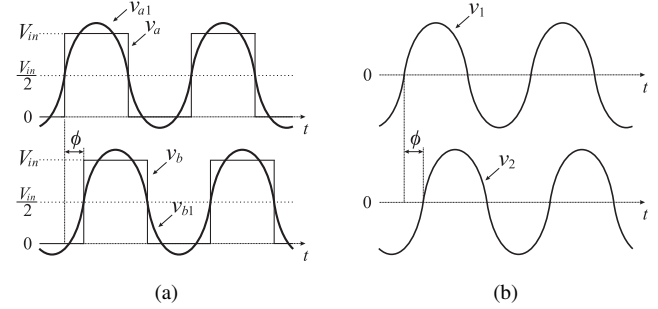


Fig. 7. Rectangular waveform voltage across middle point (a, b) of each inverter leg and negative dc bus line: (a) dc-biased, (b) non-dc biased.

### III. ANALYSIS OF OUTPUT POWER REGULATION BASED ON EQUIVALENT CIRCUIT

The frequency-domain equivalent circuit of the proposed HF-R inverter is illustrated in Fig. 6. The complex voltage vectors  $\dot{V}_{a1}$  and  $\dot{V}_{b1}$  represent the fundamental components of the  $V_{in}/2$  dc biased rectangular waveforms  $v_a$  and  $v_b$  as depicted in Fig. 7 (a), which appear between the middle points  $a, b$  and the negative dc busline on the asymmetrical half-bridge inverter units  $U_1$  and  $U_2$ , respectively. For simplifying the analysis of power regulation principle in the proposed HF-R inverter, the non-dc-biased rectangular waveforms  $v_1$  and  $v_2$  are introduced as depicted in Fig. 7 (b).

The root mean square (RMS) value  $V$  of the fundamental components  $v_{a1}$  and  $v_{b1}$  can be defined by Fourier Series Expansion as

$$V = \frac{\sqrt{2}V_{in}}{\pi}. \quad (14)$$

Accordingly,  $\dot{V}_1$  and  $\dot{V}_2$  can be expressed with the PS angle  $\phi$  as

$$\dot{V}_1 = V, \quad (15)$$

$$\dot{V}_2 = V e^{-j\phi}. \quad (16)$$

In Fig. 8, the two kinds of equivalent circuits, i.e., Superposition Principle model and Thevenin's Theorem model can be derived from Fig. 6. Based on the Superposition model in Fig. 8 (a), the two resonant link currents phasors  $\dot{I}_{L1}$ ,  $\dot{I}_{L2}$  and the IH load current phasor  $\dot{I}_o (= \dot{I}_{L1} + \dot{I}_{L2} \angle \theta, \theta = \angle \dot{I}_{L1} - \angle \dot{I}_{L2})$  can be defined as

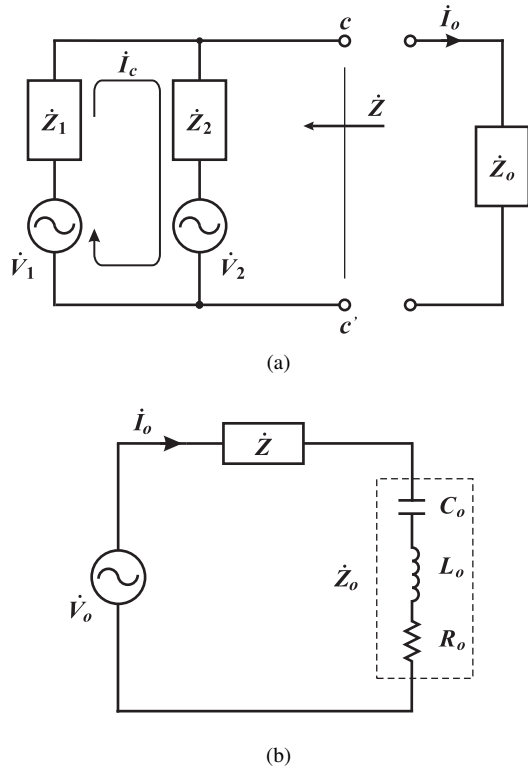


Fig. 8. Equivalent circuits with non-dc biased ac voltage sources: (a) superposition principle, (b) Thevenin's theorem.

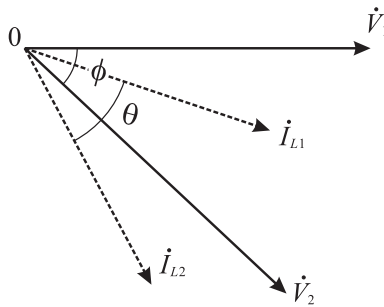


Fig. 9. Two half-bridge inverter units  $U_1, U_2$  voltage and current phasors.

$$i_{L1} = \frac{V\{1 + \frac{Z_o}{Z_1}A(1 - e^{-j\phi})\}}{2\Gamma}, \quad (17)$$

$$i_{L2} = \frac{VA\{e^{-j\phi} - \frac{Z_o}{Z_1}(1 - e^{-j\phi})\}}{2\Gamma}, \quad (18)$$

$$i_o = \frac{V(1 + Ae^{-j\phi})}{2\Gamma}, \quad (19)$$

$$A = \frac{Z_1}{Z_2} = \frac{L_1}{L_2}, \quad (20)$$

$$\Gamma = \left(\frac{1+A}{2}\right)Z_o + \frac{Z_1}{2} = R_o\left\{\left(\frac{1+A}{2}\right) + jQ\left(\frac{\omega}{\omega_r} - \frac{\omega_r}{\omega}\right)\right\} \quad (21)$$

$$\omega = \frac{2\pi}{T}, \quad \omega_r = \frac{1}{\sqrt{LC_o}}, \quad (22)$$

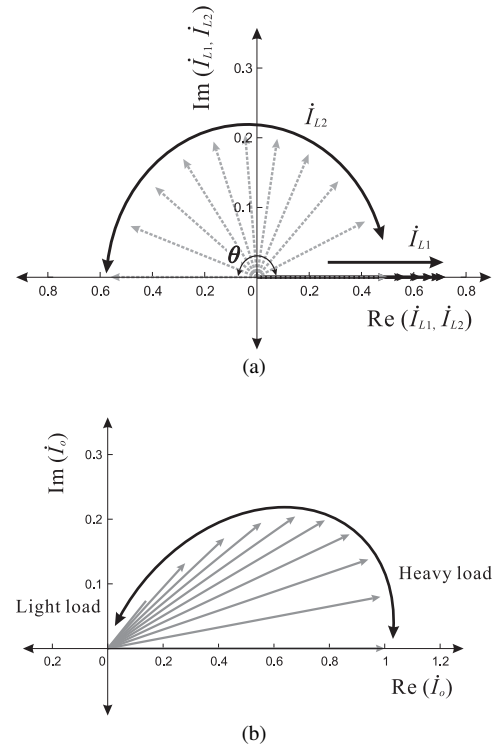


Fig. 10. Phasor diagrams of resonant link inductors and IH load current: (a) resonant link currents, (b) IH load current.

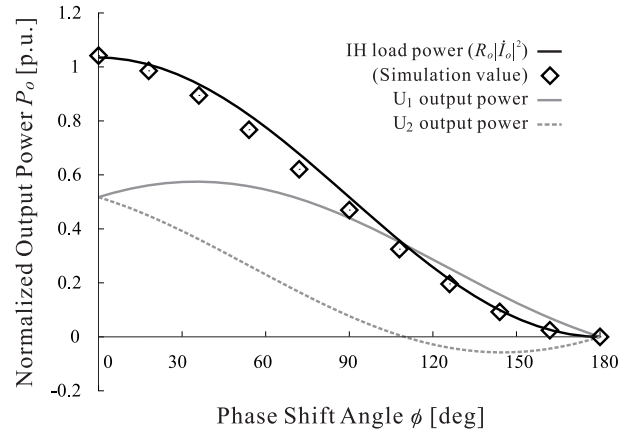


Fig. 11. Relationship between output power and phase-shift angle.

$$Q = \frac{\omega_r\{(1+A)L_o + L_1\}}{2R_o} = \frac{1+A}{2\omega_r C_o R_o}, \quad (23)$$

where  $Q$  denotes the loaded quality factor and  $A$  represents the resonant link inductors ratio, respectively. Accordingly, the IH load power can be determined by

$$P_o = R_o|i_o|^2. \quad (24)$$

The phase angle relationship between the complex voltage vector  $\dot{V}_1, \dot{V}_2$  and current vectors  $\dot{I}_{L1}, \dot{I}_{L2}$  is illustrated in Fig. 9. It should be noted here that the phase difference angle  $\theta$  between  $\dot{I}_{L1}$  and  $\dot{I}_{L2}$  is equal to  $\phi$  when  $L_1$  and  $L_2$  are the same inductance. Based on (17)–(23), the phasor diagrams of the PS-controlled resonant link currents and the resultant IH load current are indicated in Fig. 10. It can be known herein

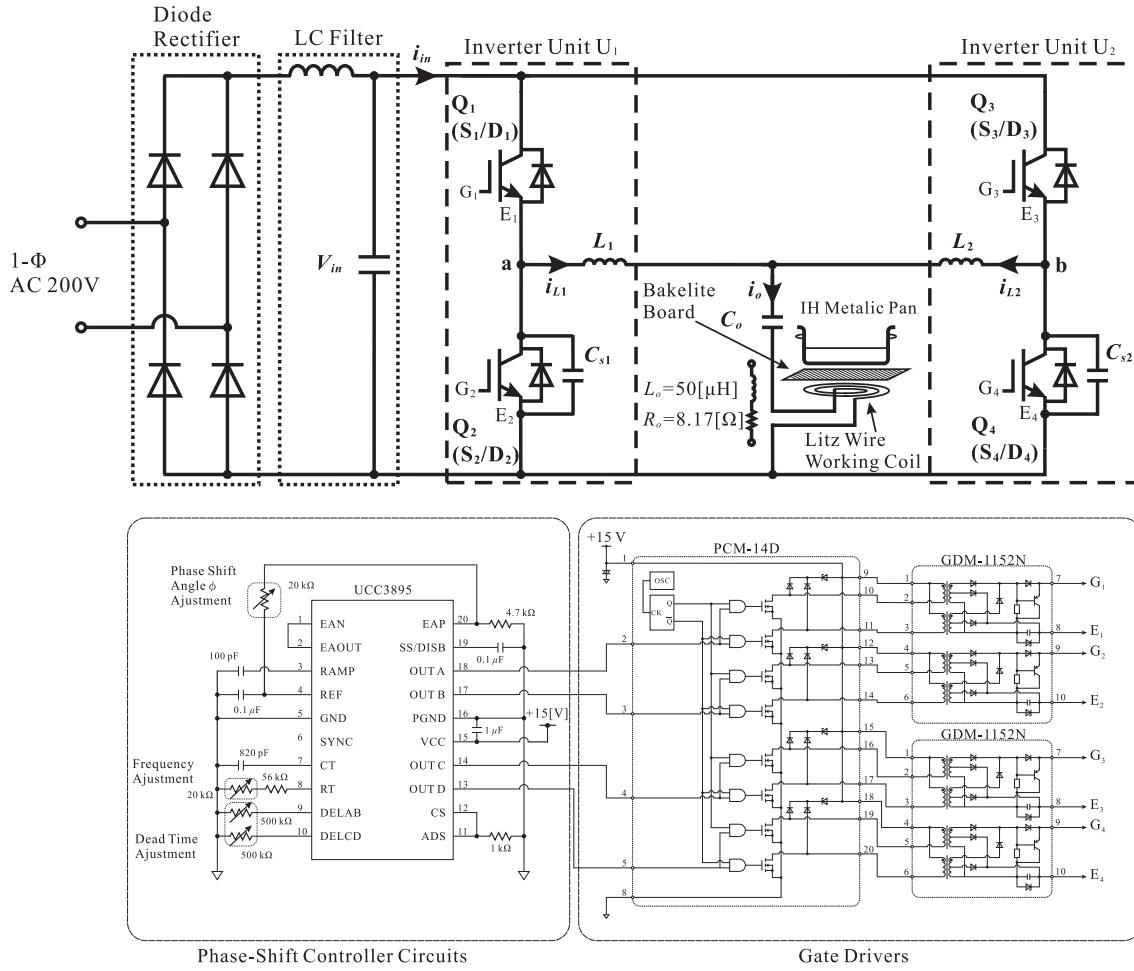


Fig. 12. Schematic diagram of the experimental main circuit and gate pulse generator with gate drivers (in open loop control).

that the load current phasor  $\dot{I}_o$  can be continuously controlled by varying the PS angle  $\phi$ .

The cross current  $\dot{I}_c$  between  $U_1$  and  $U_2$  as depicted in Fig. 8 (a), which is outstanding in the low output power settings can be expressed in RMS by

$$I_c = \left| \frac{\dot{V}_1 - \dot{V}_2}{\dot{Z}_1 + \dot{Z}_2} \right| = \frac{2V}{\omega(L_1 + L_2)} \quad (25)$$

The another equivalent circuit, Thevenin's Theorem model, is shown in Fig. 8 (b). In this equivalent circuit, the two ac voltages sources  $\dot{V}_1$  and  $\dot{V}_2$  can be expressed by just one ac voltage source (Thevenin's equivalent voltage source)  $\dot{V}_o$  as

$$\dot{V}_o = \frac{(1 + e^{-j\phi})\dot{V}}{2}. \quad (26)$$

In addition, the two resonant link inductors  $L_1$  and  $L_2$  can be simply expressed by one internal impedance  $\dot{Z}$  as

$$\dot{Z} = \frac{\dot{Z}_1 \dot{Z}_2}{\dot{Z}_1 + \dot{Z}_2} = j\omega \left( \frac{L_1 L_2}{L_1 + L_2} \right). \quad (27)$$

Thus, the Thevenin's Theorem model is useful for facilitating the analysis of power regulation in the proposed HF-R inverter.

Theoretical curves of active powers in the two inverters  $U_1, U_2$  and the IH load vs. PS angle characteristics are

depicted in Fig. 11, where the relevant simulation values are plotted for showing the accuracy of the theoretical equations through (14) to (24). It can be understood from Fig. 11 that the IH load power as well as output power of each inverter unit can be continuously controlled by varying the PS angle  $\phi$ , which facilitates the seamless output power regulation in the proposed HF-R inverter.

#### IV. EXPERIMENTAL RESULTS AND EVALUATIONS

##### A. Design Procedure for Laboratory Prototype

Performances of the soft switching operation and the output power regulation based on the current phasor control are evaluated in an experiment with the small-scale 1 kW-60 kHz laboratory prototype.

The schematic diagram of the experimental set-up with an open loop controller is illustrated in Fig. 12. The open loop method is employed for the prototype in order to focus on investigating the validity of the proposed HR-R inverter circuit topology in the experiment. The gate pulse signals for the active switches  $Q_1$ – $Q_4$  are generated by a PS-PWM controller *UCC3895* (Texas Instruments) as depicted in Fig. 12. The exterior appearance of the experimental set-up is portrayed in Fig. 13. In this experiment, the actual efficiency



TABLE I  
DESIGN SPECIFICATIONS AND CIRCUIT PARAMETERS OF PROTOTYPE.

Parameter	Value [unit]
Output power rating $P_o$	1 kW
DC input voltage $V_{in}$	240 V
Resonant inductor $L_1, L_2$	44 $\mu$ H
Power factor tuned capacitor $C_o$	0.112 $\mu$ F
Lossless snubbing capacitors $C_{s1}, C_{s2}$	6 nF
IH load (Iron pan)	Equivalent effective resistance $R_o$
	8.17 $\Omega$
	Equivalent effective inductance $L_o$
	50 $\mu$ H
Switching frequency $f_s$	60 kHz
Inverter resonant frequency $f_r$	56 kHz
Dead time $t_d$	0.5 $\mu$ s
Q <sub>1</sub> –Q <sub>4</sub> : IXGN60N60C2D1 600 V, 75 A	
Phase shift PWM IC:UCC3895	
Gate driver: GDM-1152N	

$\eta (= P_o/P_{in})$  is measured for the dc–HF ac power conversion stage by using a precision power analyzer YOKOGAWA WT1800 ( $\pm\{(0.3 \times f_s - 9.5) \% \text{ of reading} + 1 \% \text{ of range}\}$ ) as drawn in Fig. 14. In order to focus on evaluations of the essential performances in the proposed HF-R inverter and simplify the experimental set-up configuration, the iron pan is employed here in stead of aluminum and copper pans which would demand an additional impedance matching transformer.

The design specifications of the proposed HF-R inverter and experimental conditions are listed in TABLE I. The equivalent effective resistance  $R_o$  and inductance  $L_o$  of the IH load can be estimated from the transformer representation of the electromagnetic induction between the IH working coil and the metallic pan[19]. By setting the switching frequency  $f_s = 60$  kHz, the PFC tuned capacitor  $C_o$  can be calculated with  $L_o$  as

$$C_o = \frac{1}{\omega^2 L_o} = \frac{1}{(2\pi f_s)^2 L_o} = 0.112 \mu\text{F}. \quad (28)$$

The loaded quality factor  $Q = 3.0$  is given herein so that the voltage across  $C_o$  does not exceed its rated value 800 Vrms. Accordingly, the inverter resonant frequency  $f_r$  can be set to be 56 kHz from (3) for assuring the inductive load conditions and ZVS operations in Q<sub>1</sub>–Q<sub>4</sub>.

In this experimental evaluation, it is assumed that the inverter units U<sub>1</sub> and U<sub>2</sub> have the symmetric circuit parameters and the same power rating. Accordingly, the resonant link inductors  $L_r = L_1 = L_2$  can be designed from (3) as

$$L_r = 2\left(\frac{1}{\omega_r^2 C_o} - 1\right) = 2\left\{\frac{1}{(2\pi f_r)^2 C_o} - 1\right\} = 44 \mu\text{H}. \quad (29)$$

### B. Inverter Operating Waveforms

The resonant link currents  $i_{L1}, i_{L2}$ , IH load current  $i_o$ , the HF-R inverter output voltage  $v_o$  and the PFC turned capacitor voltage  $v_{co}$  for the full and no load conditions are illustrated in Fig. 15 for simulation results obtained by PSIM Ver.9-2, and in Fig. 16 for the corresponding measured waveforms, respectively. The amplitudes of the fixed and controlled-phase currents  $i_{L1}, i_{L2}$  are continuously regulated in accordance with the PS angle  $\phi$  as illustrated in Fig. 10. It can also be confirmed

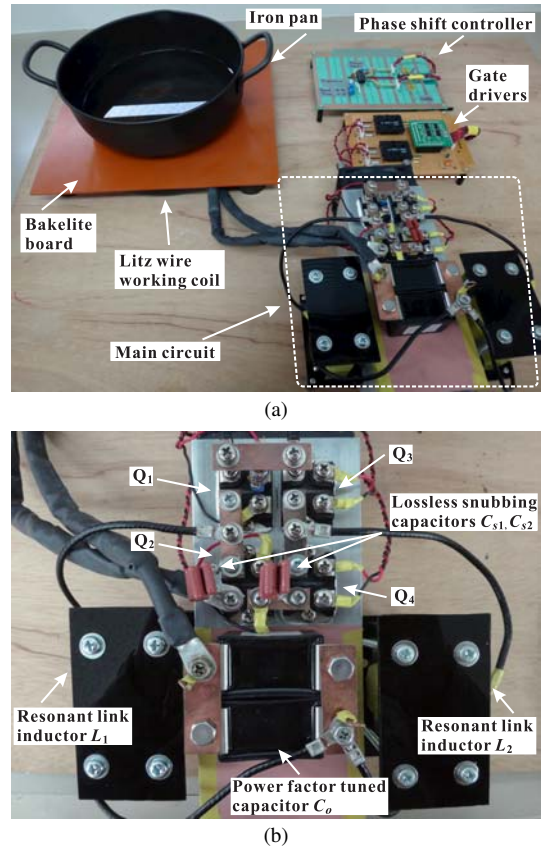


Fig. 13. Exterior appearance of the proposed HF-R inverter prototype: (a) experimental set-up, (b) main circuit.

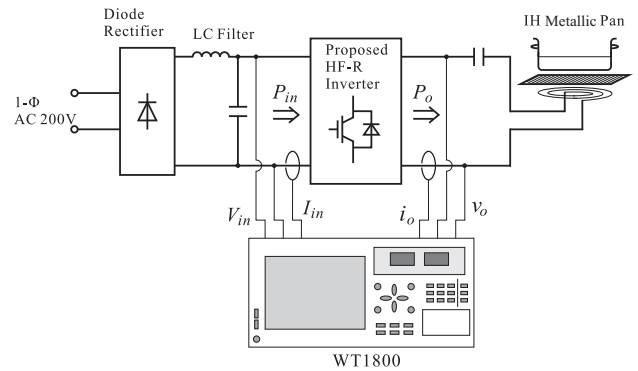


Fig. 14. Measurement set-up of actual efficiency in HF inverter prototype.

from the measured waveforms of  $i_o$ ,  $v_o$  and  $v_{co}$  that more power is supplied into the IH load as the PS angle command gets smaller. The good agreement can be confirmed between the simulation and experimental waveforms of the steady-state operating waveforms in the proposed HF-R inverter.

The switching performances of Q<sub>1</sub> & Q<sub>2</sub> in the fixed-phase inverter U<sub>1</sub> and Q<sub>3</sub> & Q<sub>4</sub> in the controlled-phase inverter U<sub>2</sub> are depicted in Figs. 17 and 18, respectively. In addition, the enlarged waveforms of the fixed-phase and controlled phase switches are depicted for the rated output power condition in Fig. 19, and for the null load condition in Fig. 20. Furthermore, the voltage and current Lissajous figures of Q<sub>1</sub> and Q<sub>3</sub> are

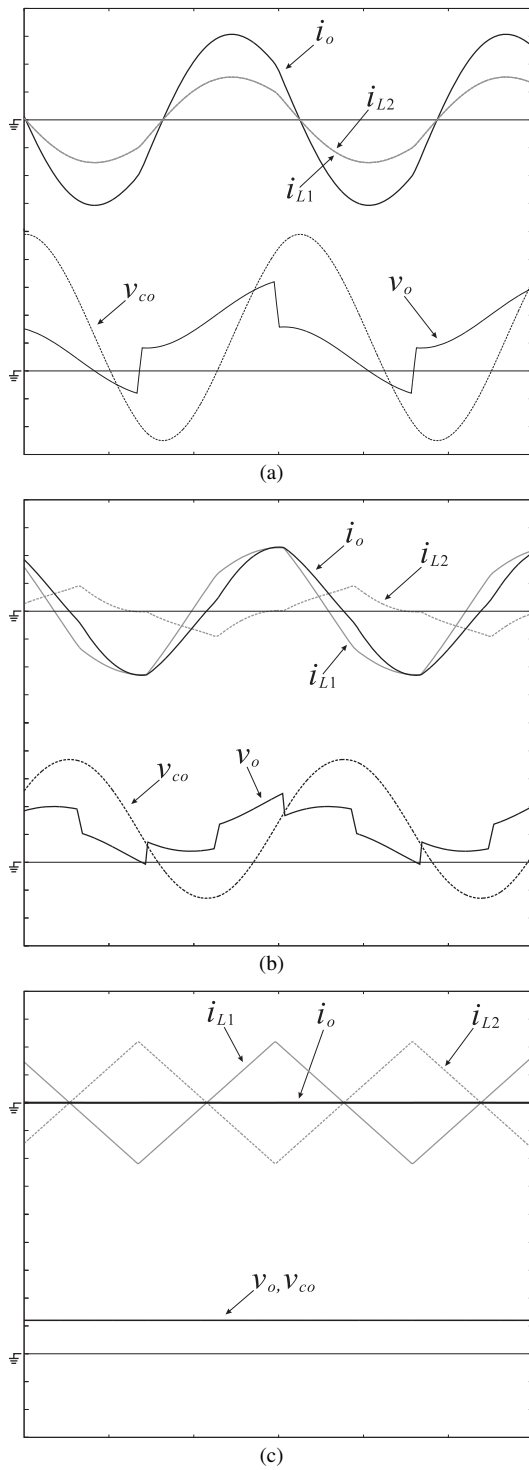


Fig. 15. Simulated steady-state operating waveforms: (a)  $P_o = 1 \text{ kW}$ ,  $\phi = 0 \text{ deg}$ , (b)  $P_o = 400 \text{ W}$ ,  $\phi = 90 \text{ deg}$ , (c)  $P_o = 0 \text{ W}$ ,  $\phi = 180 \text{ deg}$  (100 V/div, 5 A/div, 5  $\mu\text{s}$ /div).

indicated for the large output setting in Fig. 21 and for the no load setting in Fig. 22, respectively. It can be confirmed that ZVS operations are achievable and maintain in all the active switches regardless of the load power settings. ZVZCS turn-on and ZVS turn-off operations are achievable even for the null load condition, thereby the wide-range (100 % – 0 % load) soft-switching performances are actually demonstrated.

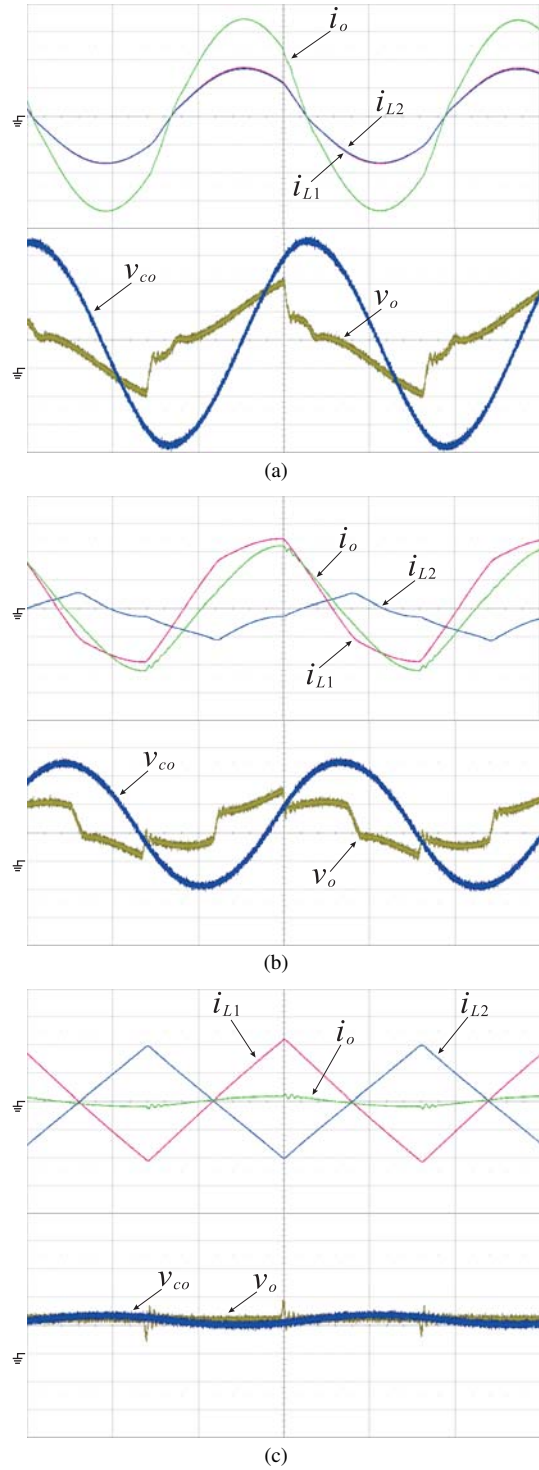


Fig. 16. Experimental steady-state operating waveforms: (a)  $P_o = 1 \text{ kW}$ ,  $\phi = 0 \text{ deg}$ , (b)  $P_o = 400 \text{ W}$ ,  $\phi = 90 \text{ deg}$ , (c)  $P_o = 0 \text{ W}$ ,  $\phi = 180 \text{ deg}$  (100 V/div, 5 A/div, 5  $\mu\text{s}$ /div).

Thus, it can be expected that EMI noise can be reduced over the wide load range in the proposed HF-R inverter.

The switching waveforms of the fixed-phase active switches  $Q_1$  &  $Q_2$  and the controlled-phase active switches  $Q_3$  &  $Q_4$  in the ZVS PS-PWM FB HF inverter under the same conditions of power rating, input dc voltage, switching frequency and IH load parameters as the proposed HF-R inverter prototype

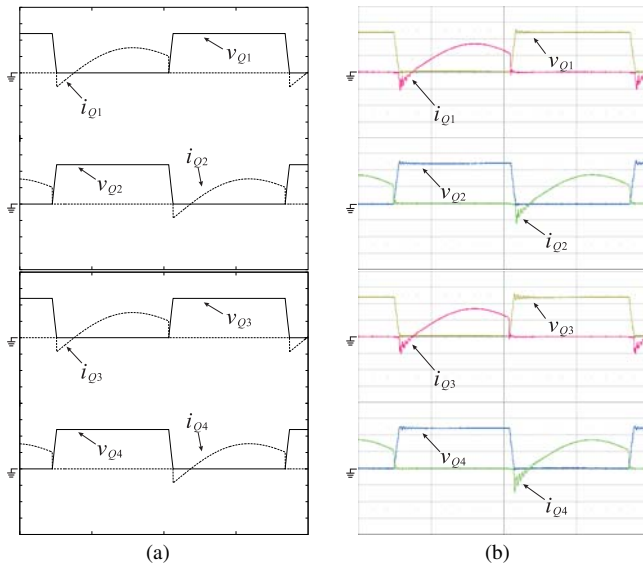


Fig. 17. Switching waveforms in the proposed HF-R inverter at  $P_o = 1 \text{ kW}$ ,  $\phi = 0 \text{ deg}$ : (a) simulated waveforms, (b) experimental waveforms (100V/div, 5A/div,  $5 \mu\text{s/div}$ ).

are drawn in Fig. 23. In addition, the enlarged waveforms of the fixed-phase and controlled-phase switches are depicted for the rated output power condition in Fig. 24, and for almost the null load condition in Fig. 25. The corresponding voltage and current Lissajous figures of the fixed-phase switch  $Q_1$  and the controlled-phase switch  $Q_3$  are portrayed for the rated output power condition in Fig. 26 and for the low output power condition in Fig. 27, respectively. It is clearly demonstrated in Fig. 23 (d), Fig. 25 (b) and Fig. 26 (d) that the complete ZVS operations are lost in the controlled-phase active switches in the low output power setting because the residual voltage exists in the lossless snubbing capacitor  $C_{s2}$ . As a result, the soft-switching range is limited in 100 % – 20 % load setting area.

It can be observed in Fig. 17 (b) and Fig. 23 (a),(b) that the peak current of the active switches are actually reduced by half in the proposed HF-R inverter as compared to the ZVS PS-PWM FB HF inverter. Actually, the RMS current of active switches in the proposed HF-R inverter is 4.25 A, while 8.45 A is measured in the ZVS PS-PWM FB HF inverter. Thus, it is actually proven that the smaller current rating power devices can be employed for the proposed HF-R inverter.

### C. Power Regulation Characteristics

The measured output power  $P_o$  vs. PS angle  $\phi$  curves are depicted in Fig. 28. The experimental characteristics demonstrated herein can well agree with those of the theoretical and simulation analysis in Fig. 11. It is actually proven herein that the wide-range output power regulation can be attained by the current phasor control.

The actual power conversion efficiencies of the proposed HF-R inverter and ZVS PS-PWM FB HF inverter are compared in Fig. 29. The proposed HF-R inverter attains the higher efficiency in the power range of 0.4 kW–1 kW (40 %–100 % load settings). This is essentially due to the lower

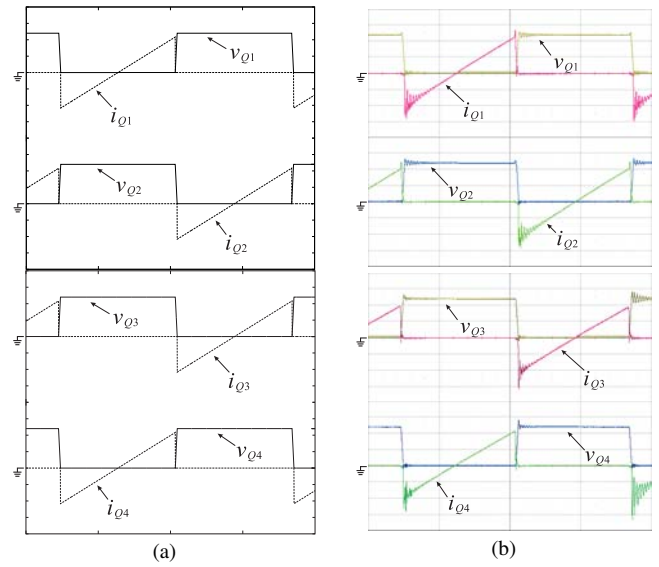


Fig. 18. Switching waveforms in the proposed HF-R inverter at  $P_o = 0 \text{ W}$ ,  $\phi = 180 \text{ deg}$ : (a) simulated waveforms, (b) experimental waveforms (100 V/div, 5 A/div,  $5 \mu\text{s/div}$ ).

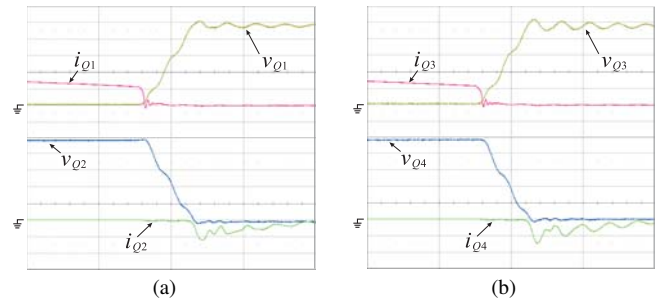


Fig. 19. Enlarged switching waveforms in the proposed HF-R inverter at  $P_o = 1 \text{ kW}$ : (a) fixed-phase switches  $Q_1$  &  $Q_2$ , (b) controlled-phase switches  $Q_3$  &  $Q_4$  (50 V/div, 5A/div, 500 ns/div).

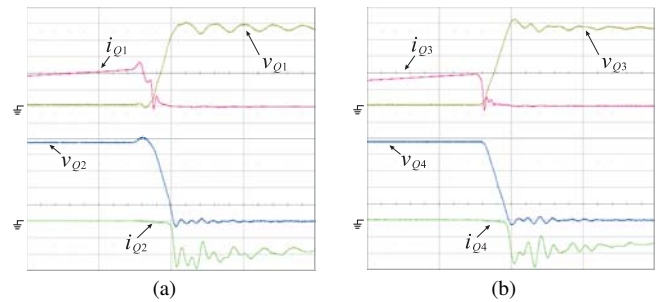


Fig. 20. Enlarged switching waveforms in the proposed HF-R inverter at  $P_o = 0 \text{ W}$ : (a) fixed-phase switches  $Q_1$  &  $Q_2$ , (b) controlled-phase switches  $Q_3$  &  $Q_4$  (50 V/div, 5A/div, 500 ns/div).

conduction power losses of the active switches in the proposed HF-R inverter. The maximum efficiency 96.3 % is obtained at  $P_o = 1 \text{ kW}$ . In the case of implementing the PS angle control only, the conversion efficiency declines gradually and lowers than 90 % in the power range below  $P_o = 400 \text{ W}$  with  $\phi = 90 \text{ deg}$ . This is due to occurrence of the relatively large peak currents which circulate through the resonant link inductors for the low output power settings as indicated in Fig. 16 (c) and Fig. 18 (b). Those cross currents cause the

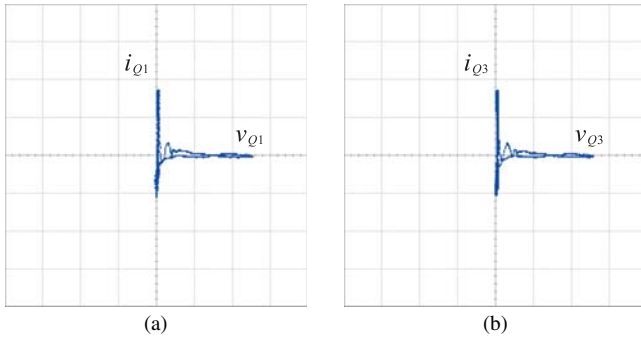


Fig. 21. Experimental current-voltage traces for switching transitions in proposed HF-R inverter at  $P_o = 1$  kW,  $\phi = 0$  deg: (a) fixed-phase  $Q_1$ , (b) controlled-phase  $Q_3$  (100 V/div, 5 A/div).

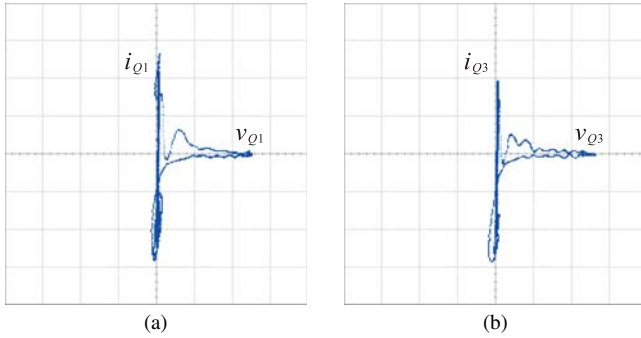


Fig. 22. Experimental current-voltage traces for switching transitions in proposed HF-R inverter at  $P_o = 0$  W,  $\phi = 180$  deg: (a) fixed-phase  $Q_1$ , (b) controlled-phase  $Q_3$  (100V/div, 5A/div).

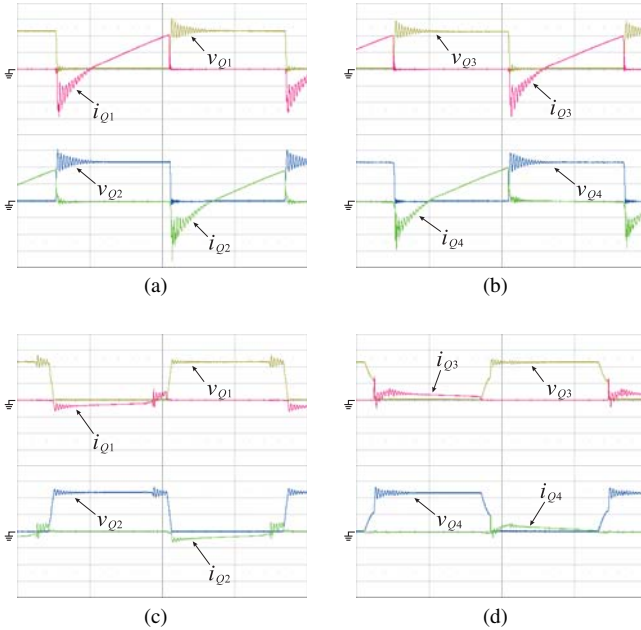


Fig. 23. Switching waveforms of ZVS PS-PWM FB HF inverter: (a) fixed-phase  $Q_1$  &  $Q_2$  ( $P_o = 1$  kW,  $\phi = 0$  deg), (b) controlled-phase  $Q_3$  &  $Q_4$  ( $P_o = 1$  kW,  $\phi = 0$  deg), (c) fixed-phase  $Q_1$  &  $Q_2$  ( $P_o = 60$  W,  $\phi = 144$  deg), (d) controlled-phase  $Q_3$  &  $Q_4$  ( $P_o = 60$  W,  $\phi = 144$  deg) (100V/div, 10A/div, 5 $\mu$ s/div).

efficiency deterioration for the low output power conditions while they work effectively for attaining ZVS commutations in  $Q_1 - Q_4$ .

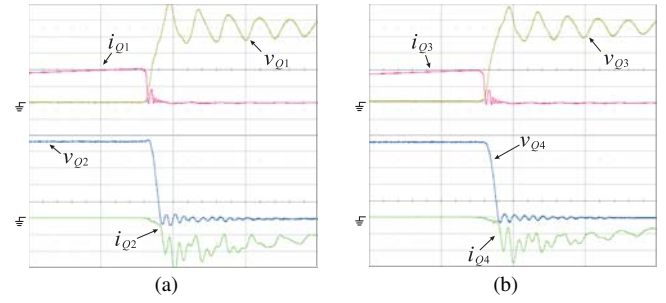


Fig. 24. Enlarged switching waveforms in the ZVS PS-PWM FB HF inverter at  $P_o = 1$  kW: (a) fixed-phase switches  $Q_1$  &  $Q_2$ , (b) controlled-phase switches  $Q_3$  &  $Q_4$  (50 V/div, 10A/div, 500 ns/div).

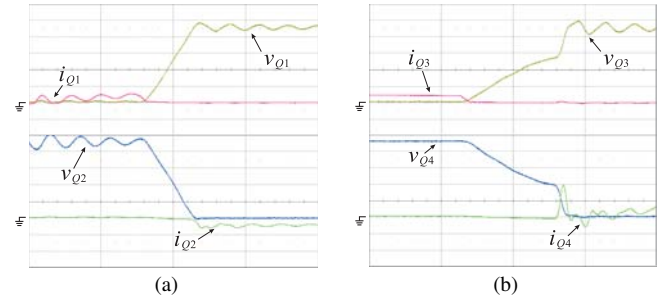


Fig. 25. Enlarged switching waveforms in the ZVS PS-PWM FB HF inverter at  $P_o = 0$  W: (a) fixed-phase switches  $Q_1$  &  $Q_2$ , (b) controlled-phase switches  $Q_3$  &  $Q_4$  (50 V/div, 10A/div, 500 ns/div).

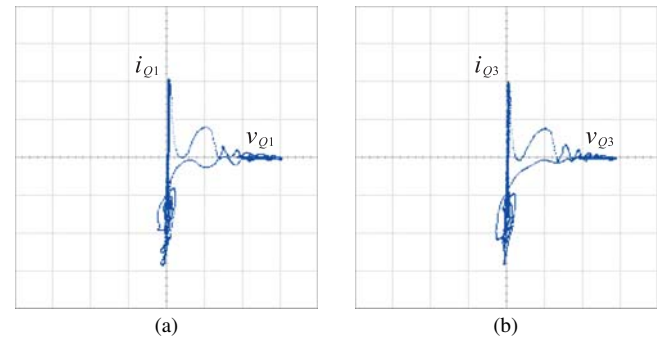


Fig. 26. Experimental current-voltage traces for switching transitions in ZVS PS-PWM FB HF inverter at  $P_o = 1$  kW,  $\phi = 0$  deg: (a) fixed-phase  $Q_1$ , (b) controlled-phase  $Q_3$  (100V/div, 10A/div).

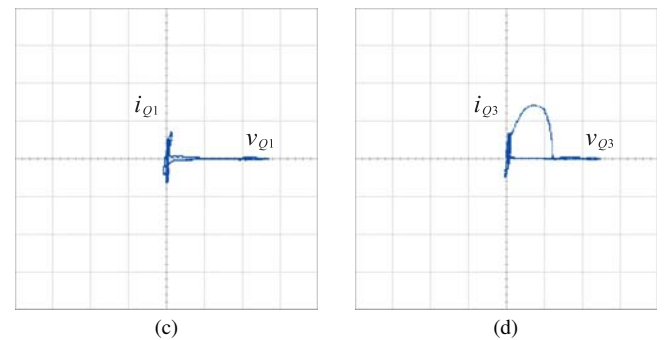


Fig. 27. Experimental current-voltage traces for switching transitions in ZVS PS-PWM FB HF inverter at  $P_o = 60$  W,  $\phi = 144$  deg: (a) fixed-phase  $Q_1$ , (b) controlled-phase  $Q_3$  (100V/div, 10A/div).

Fig.30 shows that the state phase planes of the resonant



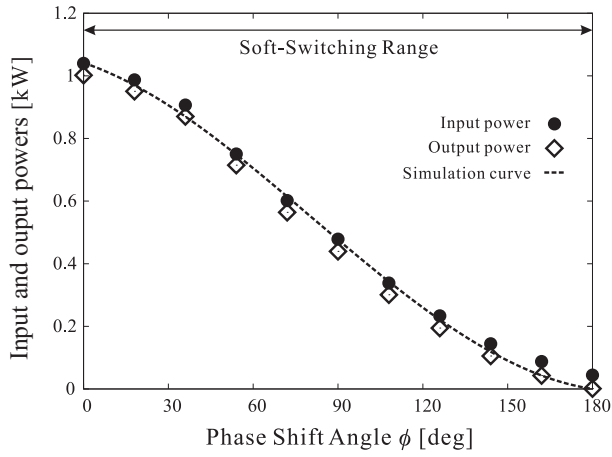


Fig. 28. Experimental characteristics of input and output powers vs. phase-shift angle.

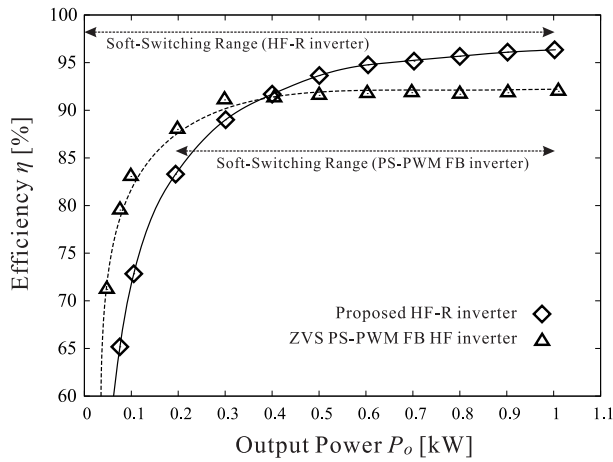


Fig. 29. Comparison of actual efficiencies vs. output power curves.

voltages and currents in the IH load, the fixed-phased inverter unit  $U_1$  and controlled-phase inverter unit  $U_2$  under the load conditions of 100 % and 10 %, respectively. It can be observed therein that the resonant tank-stored energy in  $U_1$  swells out in accordance with the decrease of IH load power, while that of  $U_1$  shrinks gradually in the area of  $0 < \phi < 90$  deg and expands out in the area of  $90 < \phi < 180$  deg, respectively.

#### D. Efficiency Improvements for Light Loads

As one of the strategies for improving the efficiency under the condition of low output power, only the single inverter unit operates in the proposed HF-R inverter by the asymmetrical PWM scheme. It is obvious that the cross currents inherent to the two-inverter unit configuration can be eliminated completely in the single inverter. The gate pulse pattern of the asymmetrical PWM is illustrated in Fig. 31, where the duty cycle  $D$  is defined by

$$D = \frac{t_{on1}}{T}. \quad (30)$$

The simulated and experimental operating waveforms of the active switches and IH load with the asymmetrical PWM are depicted in Fig. 32. The enlarged switching voltage and current

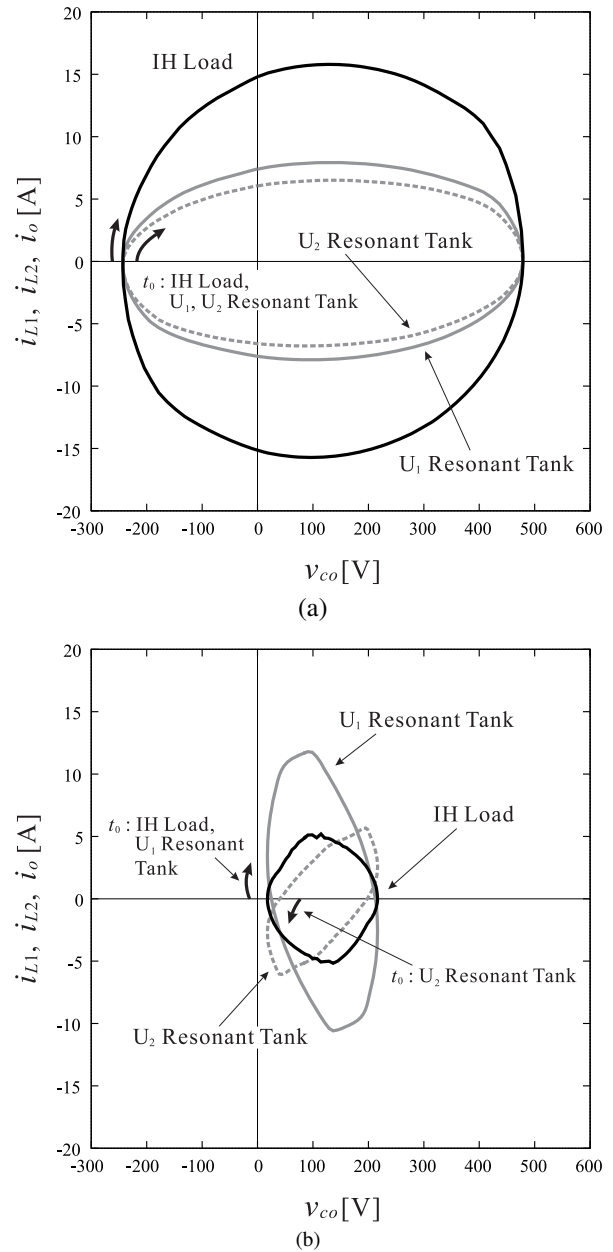


Fig. 30. Experimental state phase planes: (a)  $P_o = 1$  kW,  $\phi = 0$  deg, (b)  $P_o = 100$  W,  $\phi = 145$  deg.

waveforms are also provided in Fig. 33, where ZVS operations can be observed in  $Q_1$  and  $Q_2$ . The output power vs. duty cycle characteristics with the asymmetrical PWM are shown in Fig. 34. It can be observed that the proposed HF-R inverter operates from the middle to low output power area (from 50 % to 10 %) by the asymmetrical PWM scheme.

The actual efficiencies of the proposed HF-R inverter with the PS angle control and asymmetrical PWM are compared with that of the ZVS PS-PWM FB HR inverter in Fig. 35 for the output power range less than 400 W. It can be confirmed that 90 % can be achieved at  $P_o = 100$  % by asymmetrical PWM while 73 % is measured by the PS angle control at the same output power, although the soft switching range decreases. Thus, the effectiveness of the asymmetrical PWM-



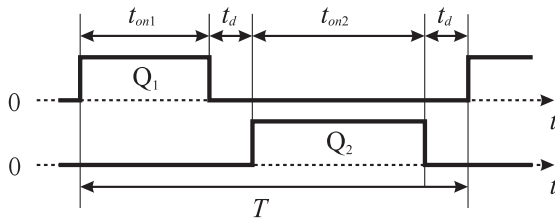


Fig. 31. Asymmetrical PWM pulse timing sequence.

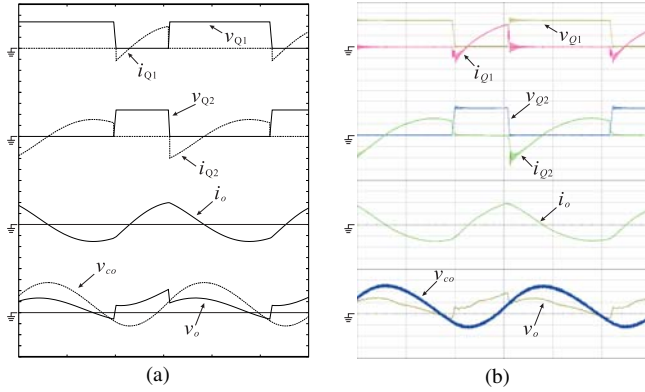


Fig. 32. Switching operations and IH load current waveforms under the condition of  $P_o = 300$  W,  $D = 0.32$  (100V/div, 5A/div, 5μs/div): (a) simulated waveforms, (b) experimental waveforms.

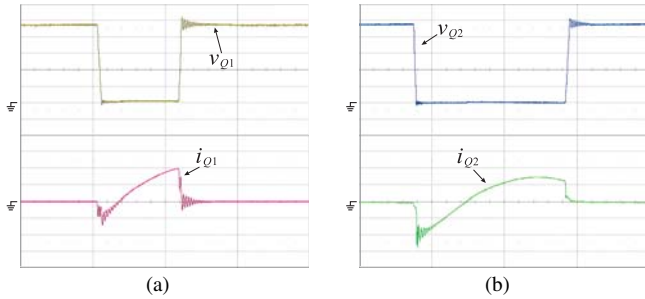


Fig. 33. Enlarged switching waveforms in the asymmetrical-PWM HF-R inverter at  $P_o = 330$  W: (a) fixed-phase switches  $Q_1$ & $Q_2$ , (b) controlled-phase switches  $Q_3$ & $Q_4$  (50 V/div, 5A/div, 5 μs/div).

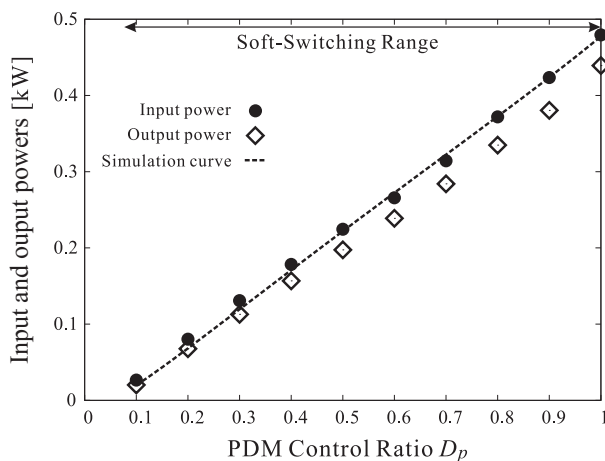


Fig. 34. Experimental characteristics of input and output powers vs. duty cycle curves in asymmetrical PWM.

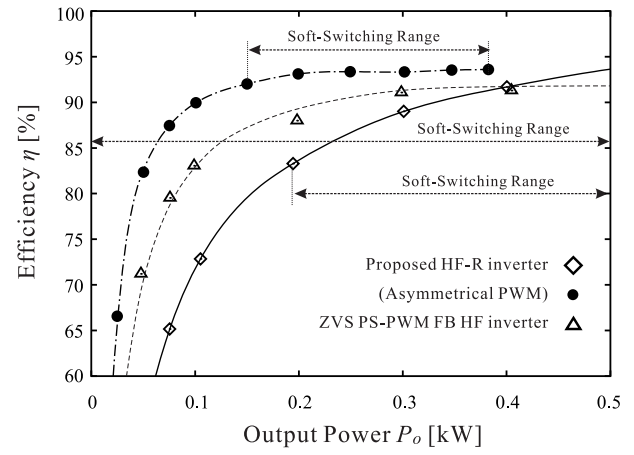


Fig. 35. Actual efficiency vs. output power characteristics in the middle-low output power area.

controlled one unit operation in the proposed HF-R inverter is actually verified for the low output power settings.

## V. CONCLUSIONS

The novel current phasor-controlled HF-R inverter for the IH applications has been newly proposed in this paper. The operation principle of the proposed HF-R inverter has been explained and the output power regulation performed by the current phasor control has been described by the frequency-domain equivalent circuit-based theoretical analysis. The performances on the soft-switching transitional commutations and the output power regulation in the proposed HF-R inverter have been originally evaluated in an experiment with its laboratory prototype, and its beneficial properties such as a wide-range soft switching operation and high-efficiency power conversion have been actually verified.

The efficiency deteriorates under the conditions of low output powers because the cross current and its relevant reactive power increase between the two half-bridge inverter units when the output power regulation is carried out by only the PS angle control. In order to prevent the efficiency drops, the unit number exchanging from two to one with an asymmetrical PWM has been additionally incorporated into the proposed HF-R inverter prototype operating for the low output power setting. It has been actually demonstrated from the relevant experimental results that more than 20 % efficiency improvement can be achieved for the low output power settings by adopting the PS & asymmetrical PWM dual-mode control scheme.

The comparative investigation between the PS & asymmetrical PWM scheme evaluated herein and another dual mode strategy employing pulse-density-modulation (PDM) [19]–[21] for improving efficiency under the condition of low output power will be one of the next research subjects for the proposed HF-R inverter.

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