Abstract—The design of a resonant inverter for high-frequency ac (HFAC) power distribution systems is complicated by the following three factors: 1) A number of electronic loads located in different locations are connected to the resonant inverter, the impedance, and the power factor of the equivalent load of which varies over a wider range than a system with a certain load; 2) the resonant inverter is subject to an input-line voltage varying over a wide range; and 3) the characteristics of the resonant inverter depend on the load impedance. It is mandatory to operate the resonant inverter with zero-voltage switching under various load conditions of different power factors and over wide input variations. It is further desirable that multiple resonant inverters can be paralleled with simple current-sharing control (CSC). A phase-shift-modulation (PSM)-controlled full-bridge series-parallel resonant inverter is proposed for the HFAC power distribution architectures. A new PSM method is proposed with which the phase angle of the inverter output voltage is independent of the modulation signal of the phase-shift modulator. Such a feature allows multiple resonant inverters to operate in parallel with a magnitude CSC. The resonant inverter is analyzed with a general nonresistive load model, and the design curves are developed. A prototype resonant inverter system is designed and implemented with an operation frequency of 1 MHz, a rated output power of 150 W, and a sinusoidal output voltage of 1 MHz, 28-V rms. The proposed resonant inverter has the advantages of high efficiency over wide input/output line variations, high waveform quality of the output voltage, and phase-angle independence of the voltage-feedback/feed-forward control and CSC.

Index Terms—Full bridge, phase-shift modulation (PSM), power distribution architecture, resonant inverter, soft switching.

I. INTRODUCTION

In the past few decades, the semiconductor industry has witnessed the successful prediction of Moore’s law that the scale of integration of transistors in cutting-edge integrated circuits doubles every 18 months or more. According to the international technology roadmap for semiconductors [1], such advancement will continue for another 15 years or more. The high-frequency alternative current [high-frequency ac (HFAC)] power distribution architecture, which was proposed by NASA decades ago for space power supply applications [5], has recently attracted a lot of attention from both industry and academia [6]–[12]. A conceptual schematic diagram of the HFAC distributed power system is shown in Fig. 1(b), consisting of one or a number of high-frequency dc/ac resonant inverter(s) in parallel; HFAC bus and a number of ac voltage regulation modules (ACVRM) located at points-of-use for local power management. Power is delivered to the ACVRM through HVLC tracks, reducing the power delivery losses and footprints. While at the points-of-use, the power is delivered to the load through a very short low-voltage high-current tracks, reducing the adverse effects associated with the parasitic parameters. Compared with the dc power distribution architecture in Fig. 1(a), the ac/dc conversion in the front stage and the dc/ac in the VRM stage are eliminated in the HFAC power distribution architecture. Hence, the overall efficiency should be higher than the dc power distribution architecture. In addition, there are other advantages of the HFAC power distribution system [6]–[12], such as simpler structure, lower cost, smaller component count, effective ground noise isolation, and regenerative energy steering back from the load.

The design of the resonant inverter is challenging because of the following factors: 1) The HFAC bus with a cluster of
distributed loads is subject to more dynamic changes than a single load. 2) The power factor of the equivalent load of the HFAC bus changes over a wide range. The characteristics of the resonant tank change with the load impedance. 3) The total harmonic distortion (THD) of the HFAC bus voltage must be kept low. 4) The input-line voltage varies over wide range. The soft switching should be maintained for all the power switches with a consideration of all these factors.

Resonant inverters consisting of a switching network and a high-order resonant tank are used to convert the dc voltage into high-frequency sinusoidal ac voltage. Both half- and full-bridge topologies can be used to generate the semisquare waveform voltages, the width of which can be controlled through phase-shift modulation (PSM) [6], PWM, or asymmetrical PWM (APWM) [8]. The resonant tank not only filters out the harmonics in the semisquare voltage but also offers soft switching for the power switches. Normally, the resonant tank can have two, three, or four energy storage components, and the topologies can be very complicated [13]. A resonant tank with two energy storage components is usually used in the dc/dc conversion topologies, for example, the parallel resonant converter and the series resonant converter [14]. A resonant tank of three energy storage components, such as an LCC tank, has been used for resonant inverters for high-intensity-discharge lamp ballasts, where a high-ignition voltage is required [15], [16]. However, sinusoidal voltage waveform with a very small THD is desirable in an HFAC system to avoid electromagnetic interference (EMI) in the HFAC bus. The series-parallel resonant tank with four energy storage components has been explored for applications in the dc/dc power conversions [17]–[19]. It gives more design freedom in the soft-switching design. Such resonant tank also has better filtering performance than an LCC or LC tank. Therefore, it is a rational choice for the HFAC applications.

However, the output-voltage waveform quality depends not only on the resonant tank but also the pulse modulation methods. Any APWM [8] will generate a certain amount of even order harmonics in the output voltage. In this case, an extra filter is needed. Hence, it is preferable that the symmetrical PWM such as PSM be used.

Another very important issue about the HFAC power distribution system is that the paralleling of multiple resonant inverters is difficult. Multiple module operation brings numerous advantages such as flexible maintenance, distributed power-loss dissipation, low expense expansion of the power capacity, and off-the-shelf modular design. However, the paralleling of multiple resonant inverters is challenging. The circulating current in a multiple module system can be caused by the phase and/or magnitude discrepancy of the output voltages of individual inverters. A current-sharing control (CSC) is necessary to ensure an evenly current sharing among the inverter modules. A phasor CSC has been proposed in [20] for accurate current sharing and circulating current minimization in multiple inverter systems consisting of a number of two-stage resonant inverters.

One alternative and simpler method for the full-bridge resonant inverters is an average magnitude CSC, which is shown in Fig. 2(a) for a two-inverter system. The resonant inverters are connected to the HFAC bus through the small connection inductors $Z_1$ and $Z_2$. In order to keep the voltage drop and power loss small over the inductors, the impedance, particularly the resistance, must be small and of similar parameters, i.e.,

$$X_{L_{K1}} \approx X_{L_{K2}}$$

$$X_{L_{K1}} \gg R_{k1}; \quad X_{L_{K2}} \gg R_{k2}.$$  

The control stage consists of CSC unit, current controller $K_C(s)$, voltage controller $K_V(s)$, one-cycle controller, and
the PSM. The current magnitudes of individual inverters are averaged by the CSC and used as the reference-current signal $I_{\text{ref}}$ for all the inverters. The current error signal is the difference between the actual current of each of the inverters $I_{1,2}$ and the reference current $I_{\text{ref}}$, as given in (3). The voltage error signal is the difference between the reference $\nu_{r}$ and the HFAC bus voltage $\nu_{o}$, as given in (4). The current error signal is sent to the current controller $K_{C}(s)$. The output is then combined with the output of the voltage-feedback control $K_{V}(s)$ and the rectified bridge voltage, as the modulation...
signal for the phase-shift modulator, which is given in (5). That is,

\[
i_e = [I_{ref} - I_1I_{ref} - I_2]
\]

\[
n_v = \nu_t - \nu_o
\]

\[
\nu_{m,n} = -||\nu_n|| + K_c i_{e,n} + K_p \nu_{e,n}
\]

\[
= \left( -\frac{\delta_n}{\pi} V_{in,n} + K_c (\nu_t - \nu_o) \right) + K_{e,n} \left( \frac{1}{N} \sum_{N} I_n - I_n \right), \quad n = 1, 2. \tag{5}
\]

The magnitudes of individual inverters can be controlled by changing the phase-shift angle \(\alpha\) between the two legs of the full-bridge inverter according to the modulation signal \(\nu_{in}\). Suppose that module 1 takes less current than module 2. The current error is less than that of module 2. The corresponding modulation signal sent to the PSM is larger than module 2. This increases the output voltage of module 1. Since the HFAC bus voltage is regulated, the current in module 1 will be increased. Therefore, the CSC is achieved by the control of the magnitudes of the output voltages of the inverters. Such a method is simple to implement.

However, the conventional PSM [21] has the inherent problem that it will cause further phase-angle difference of the output voltages, because the phase angle of the output voltage is dependent of the modulation signal of the PSM. The mechanism of the conventional PSM is explained with the block diagram shown in Fig. 2(b) and the operation waveforms shown in Fig. 2(c). From Fig. 2(c), with the geometry relationship, it is easy to find the phase-shift angle between the two sets of PWM pulses for Legs A and B, which is given in (6), where \(V_{pp}\) is the peak-peak value of the carrier signal, and \(V_m\) is the value of the modulation signal. We have

\[
\alpha = \frac{\pi}{V_{pp}} (V_{pp} - V_m). \tag{6}
\]

However, it is noted that the switching function of the full bridge shifts with respect to the different modulation levels because, in such scheme, the falling edge of the pulse for Leg A is fixed, while the falling edge of the pulse for Leg B shifts for phase \(\alpha\). Therefore, the fundamental component of the bridge voltage shifts phase \(\alpha/2\). The output voltage of the inversion stage is given in (7) in terms of the modulation level, where \(Z_p\) and \(Z_s\) are the impedance of the parallel and series resonant tank. Note that the phase angle of the output voltage is dependent on the modulation level. That is,

\[
V_p = V_o \frac{Z_p}{Z_p + Z_p} = \frac{Z_s}{Z_s + Z_p} \frac{4}{Z_s + Z_p} \sin \left( \frac{\pi}{2} - \frac{\pi}{2} \frac{(V_{pp} - V_m)}{2V_{pp}} \right) \times \sin \left( \omega_c t + \frac{\pi}{2} \frac{(V_{pp} - V_m)}{2V_{pp}} \right). \tag{7}
\]

This will introduce phase discrepancy in the terminal voltages for the different inverter modules. Obviously, according to (5), the phase discrepancy can be a result of CSC, different input-line voltages, or component tolerances. Therefore, it cannot achieve CSC with the conventional PSM.

A full-bridge resonant inverter with a novel PSM is proposed in this paper for the HFAC power distribution applications. The block diagram of the HFAC power distribution architecture is shown in Fig. 3(a), which consists of the full-bridge switch network, a series-parallel resonant tank, an HFAC bus, a number of ACVRMs, and the control circuits. Compared with the half-bridge topology [8], the full-bridge topology can handle more power and higher input voltage. A high-order resonant tank consisting of the series and parallel LC resonant tanks is adopted for the best filtering performance. The soft switching of the power switches can be achieved by tuning the resonant frequency of the series resonant tank lower than the operation frequency of the resonant inverter. To reduce the circulating current, the parallel resonant tank is tuned approximately to the resonance. The high-frequency transformer not only provides isolation between the input and output but also eases the soft-switching design with the consideration of the wide range of input voltage. A new PSM method is proposed with which the phase angle of the output voltage of the full-bridge resonant inverter is independent of the voltage-feedback control. Such a feature allows multiple resonant inverters operated in parallel with a magnitude CSC.

The typical waveforms for the PSM control are shown in Fig. 3(b), where \(\alpha\) is the phase-shift angle between the two legs, which are Legs A and B. \(\delta\) is the effective pulsewidth for the PSM. \(\varphi\) is the phase angle between the fundamental component of the bridge voltage \(\nu_n\) and the resonant current \(i_n\). This angle is dependent on both the resonant tank and the load impedance, which makes the analysis and design complicated. The principal mechanism of the full-bridge series-parallel resonant inverter has been described in detail in [12].
is noted that $\varphi > 0$ is not a sufficient condition for the zero-voltage switching (ZVS) of all the switches. But the switches $S_1$ and $S_4$ always have soft switching. The phase angle between the phases of the triggering of the $S_3$ and the zero crossing of the resonant current $i_s$ is denoted as $\theta$ and is related to $\varphi$ and $\alpha$ by

$$\theta = \varphi - \frac{\alpha}{2}. \quad (8)$$

Theoretically, if the angle $\theta$ is larger than zero, the switches $S_2$ and $S_3$ have soft switching. The rest of this paper is organized as follows. In Section II, ac analysis with a general load model will be performed, and the design curves will be presented. In Section III, the voltage and current stresses and the turning-off currents will be analyzed. A new PSM will be described in Section IV, with which the phase angle of the output voltage of the inverter systems. To reduce the voltage drop and power loss over the connection inductance, both $L_k$ and $R_k$ are very small. The conduction resistance of the switches and the parasite resistance of the series resonant branch are lumped into one parasite resistance $R_e$. The parasitic resistance of the parallel resonant branch consists of the equivalent resistance and reactance for the equivalent resistance $Z_e$ of the series and parallel resonant tanks are determined by the series and parallel inductors and capacitors as follows:

$$\omega_e = \frac{1}{\sqrt{L_e C_e}}; \quad \omega_p = \frac{1}{\sqrt{L_p C_p}}. \quad (12)$$

The quality factor of the series resonant network and parallel resonant tank is defined in (13) in terms of the reactance of the operation frequency $\omega_e$, respectively. That is,

$$Q_e = \frac{X_{L_e}}{R_e}; \quad Q_p = \frac{R_p}{X_{L_p}} \quad (13)$$

$$k_s = \frac{\omega_s}{\omega_e}; \quad k_p = \frac{\omega_p}{\omega_e}. \quad (14)$$

The voltage across the parallel network component $L_p$ and $C_p$, or the primary-winding voltage of the transformer, is given in (15) for $h$th harmonics, where $h$ stands for the $h$th harmonics. The impedances of the series and parallel resonant tanks are given in (16) and (17), respectively. That is,

$$V_{p,h} = V_{a,h} \frac{Z_{p,h}}{Z_{s,h} + Z_{p,h}} \quad (15)$$

$$Z_{a,h} = R_s + j \left(h \frac{X_{L_s} - \frac{X_c}{h}}{h} \right) \quad (16)$$

$$Z_{p,h}^{-1} = \frac{R_p}{R_p^2 + X_{p,h}^2} + j \left(\frac{h}{X_{C_p} - \frac{1}{X_{L_p} h} - \frac{X_{e,h}}{h^2}} \right) \quad (17)$$

where

$$X_{L_s} = \frac{1}{k_e} \sqrt{\frac{L_s}{C_s}}; \quad X_{C_s} = k_s \sqrt{\frac{L_s}{C_s}} \quad (18)$$

$$X_{L_p} = \frac{1}{k_p} \sqrt{\frac{L_p}{C_p}}; \quad X_{C_p} = k_p \sqrt{\frac{L_p}{C_p}} \quad (19)$$

The impedance of the parallel resonant branch is therefore rewritten in (20). That is,

$$Z_{p,h} = R_{p,h} + j X_{p,h} \quad (18)$$

$$R_{p,h} = \left(\frac{R_p}{R_p^2 + X_{p,h}^2} \right)^2 + \left(\frac{h}{k_p} - \frac{h}{k_p} \right) \sqrt{\frac{L_p}{C_p} - \frac{X_{e,h}}{R_p^2 + X_{p,h}^2}} \quad (19)$$

$$X_{p,h} = \left(\frac{k_p}{h} - \frac{k_p}{h} \right) \sqrt{\frac{L_p}{C_p} + \frac{X_{e,h}}{R_p^2 + X_{p,h}^2}} \quad (20)$$
Hence, the impedance viewed from the input port of the resonant network is given in (21), from which we can derive the resonant current and the impedance phase angle \( \varphi \) in (23). That is,

\[
Z_{i,h} = Z_{s,h} + Z_{p,h} \\
= (R_a + R_{p,h}) + j(X_{s,h} + X_{p,h})
\]

(21)

\[
X_{s,h} = hX_{L_s} - \frac{X_{C_s}}{h} = \left( \frac{h}{k_a} - \frac{k_a}{h} \right) \sqrt{\frac{L_s}{C_s}}
\]

(22)

\[
\varphi_h = \cos \left( \frac{R_s + R_{p,h}}{\sqrt{(R_s + R_{p,h})^2 + (X_{s,h} + X_{p,h})^2}} \right)
\]

(23)

The primary-winding voltage of the transformer and the ac gain of the resonant network can be derived in (24) and (25) (shown at the bottom of the page), respectively.

The phase angle \( \gamma_h \) between the primary-winding voltage \( V_{p,h} \) and the bridge voltage \( V_{a,h} \) is given as follows:

\[
\gamma_h = \tan \left( \frac{X_{p,h}(R_a + R_{p,h} - R_{p,h}(X_{s,h} + X_{p,h}))}{R_{p,h}(R_a + R_{p,h}) + X_{p,h}(X_{s,h} + X_{p,h})} \right)
\]

(26)

A. Resistive Load

Assume that the equivalent load of the HFAC bus is resistive. The ac gain is a function of both the pulsewidth \( \delta \) and the circuit components and can be simplified as in (27) and shown in Fig. 5(a), where \( A_1 \) is given in (24). We have

\[
G_I = \frac{V_o}{V_{in}} = A_1 \frac{4 \sin \left( \frac{\delta}{2} \right)}{\pi}
\]

(27)

The impedance angle in (23) is the characteristics of the resonant tank and the equivalent load of the HFAC bus. Therefore, it is independent of \( \delta \). However, for different \( k_s \) and load conditions, the phase angles are quite different, as shown in Fig. 5(b). Obviously, both \( \delta \) and \( \varphi \) are dependent on the load conditions, the parameters of the resonant tank, and the transformer turns ratio. If \( \varphi < \alpha/2 \), then the inverter loses ZVS.

The pulsewidth \( \delta \) for different circuit parameters are plotted in Fig. 6(a). The pulsewidth increases as \( k_s \) decreases, indicating the reduction of the effective gain of the inverter. In order to achieve enough gain, a high \( k_s \) is expected but cannot be larger than 1. In practice, it can be much smaller for the consideration of ZVS range. The \( \delta \) for different transformer turns ratios is also plotted in Fig. 6(b). If the transformer turns ratio is too small, then the inverter cannot reach 1 per unit power capacity or rated voltage, because the pulsewidth cannot go beyond 3.14. However, the ZVS range can be narrowed for large turns ratio.

B. Nonresistive Load

The phase angle \( \varphi \), pulsewidth \( \delta \), and minimum voltage are influenced by the power factor of the equivalent load of the HFAC bus, which is paralleled with the parallel resonant tank. In addition, nonresistive load causes more circulating current to flow in the switches and the resonant components. Therefore, the conduction losses are higher, and the efficiency is lower for the nonresistive load conditions, comparing to the resistive load conditions.

As shown in Fig. 7(a), the pulsewidth \( \delta \) depends on both the output power and the power factor of the equivalent load of the HFAC bus. Because of the reduced ac gain of the inverter, the input voltage range for ZVS is also narrowed. The minimum
input voltage increases as the power factor decreases, as shown in Fig. 7(b). The phase angle \( \phi \) is plotted in Fig. 7(c) for different load conditions, with an input voltage of 60 V and the transformer turns ratio of 0.9. It can be observed that the phase angle \( \phi \) is increased because of the reactive impedance of the equivalent load of the HFAC bus, which is strongly indicating that the nonresistive load condition results in more circulating current in the resonant inverter and lower system efficiency.

III. VOLTAGE AND CURRENT STRESSES ANALYSIS

The bridge voltage to the resonant network \( v_b(t) \) can be expressed in terms of Fourier series in (28). For the \( h \)th harmonics, the resonant current \( i_{s,h}(t) \) is represented by (29), where \( \varphi_h \) is the phase angle of the \( Z_{i,h} \). That is,

\[
\nu_a(t) = \sum_{h=1,3,5,...}^{\infty} \frac{4V_{in}}{h\pi} \sin \frac{h\delta}{2} \sin \frac{h\pi}{2} \sin(h\omega_0 t) \tag{28}
\]

\[
i_{s,h}(t) = \frac{V_{in}}{Z_{i,h}} \angle \varphi_h = \frac{4V_{in}}{h\pi Z_{i,h}} \sin \frac{h\delta}{2} \sin \frac{h\pi}{2} \sin(h\omega_0 t - \varphi_h). \tag{29}
\]

The voltage stresses over the two series resonant components \( L_a \) and \( C_a \) are derived from the resonant current and are given as

\[
\nu_{L_a}(t) = \sum_{h} \frac{4V_{in}X_{L_a}}{h\pi} \sin \frac{h\delta}{2} \sin \frac{h\pi}{2} \cos(h\omega_0 t - \varphi_h). \tag{30}
\]

\[
\nu_{C_a}(t) = \sum_{h} \frac{4V_{in}X_{C_a}}{h^2\pi Z_{i,h}} \sin \frac{h\delta}{2} \sin \frac{h\pi}{2} \cos(h\omega_0 t - \varphi_h). \tag{31}
\]

The instantaneous voltage of the primary winding of the transformer can be expressed in (32), where the phase angle \( \gamma_h \) is the phase angle between the \( h \)th harmonics of the bridge voltage and the transformer primary-winding voltage, which is given in (26)

\[
\nu_{p,h}(t) = \sum_{h} \frac{4V_{in}X_{L_a}}{h\pi} \sin \frac{h\delta}{2} \sin \frac{h\pi}{2} \cos(h\omega_0 t - \gamma_h). \tag{32}
\]
The currents in the parallel resonant components $L_p$ and $C_p$ can be derived from the voltage in (32) and are written as

$$i_{C_p}(t) = \sum_h A_h \frac{4V_{in}}{X_{C_p}\pi} \sin \frac{h\delta}{2} \sin \frac{h\pi}{2} \cos(h\omega_c t - \gamma_h)$$

$$i_{L_p}(t) = -\sum_h A_h \frac{4V_{in}}{X_{L_p}\pi h^2} \sin \frac{h\delta}{2} \sin \frac{h\pi}{2} \cos(h\omega_c t - \gamma_h).$$

(33)

(34)

The following base values are defined for the normalization of the voltage and current stresses for the visual presentation in the figures that followed:

1 per unit Power = $P_0$, 1 per unit Volt = $V_{in}$

1 per unit Ampere = $V_{in}/R_e$.

With (30) and (31)–(34), the voltage stresses and current stressed across the resonant components for different operation conditions can be studied. According to (30) and (31), the voltage stresses over the series components are more related to the load conditions. In Fig. 8, the voltage stresses are almost in proportion to the load, while, for various $k_s$ and $k_p$, they do not change much. The current stresses in the parallel inductor $L_p$ and capacitor $C_p$ are relatively independent of the $k_s$. However, they are reversely proportional to the load conditions and are affected by the tuning factor $k_p$ of the parallel resonant tank.

The resonant current of the inverter is rewritten in (35). The turning-off/turning-on currents of the four switches in the full-bridge resonant inverter can be derived from the resonant current at the instants when the switches are turned off/turned on. Therefore, the turning-off/turning-on currents of each of the switches can be determined and expressed in terms of Fourier series, supposing that the switching instants are known.

$$i_s(t) = \sum_{h=1,3,...} \frac{4V_{in}}{h\pi|Z_{i,h}|} \sin \frac{h\delta}{2} \sin \frac{h\pi}{2} \sin(h\omega_c t - \varphi_h).$$

(35)

For instance, the turning-on current of switch $S_1$ is the turn-off current of switch $S_4$ and is given in (36), assuming that the dead time is ignored. Similarly, the turning-on current of switch $S_3$ is the turn-off current of switch $S_2$ and is given in (37)

$$i_s(\omega t_1) = -\frac{\pi - \delta}{2}$$

$$i_s(\omega t_2) = \frac{\alpha}{2}$$

(36)

(37)

The root mean square (rms) current conducted by $S_1$ and $S_3$ can be derived from (38) and (39), respectively. If we consider the fundamental component only, we can get the approximate evaluation of the rms currents in (40) and (41), respectively, for the two switches $S_1$ and $S_3$. That is,

$$I_{1\text{rms}} = \sqrt{\frac{\pi - \alpha/2}{2\pi} \sum_h \frac{4V_{in}}{h\pi|Z_{i,h}|} \sin \frac{h\delta}{2} \sin \frac{h\pi}{2} \sin(h\omega t - \varphi_h)^2}$$

$$I_{3\text{rms}} = \sqrt{\frac{\pi + \alpha/2}{2\pi} \sum_h \frac{4V_{in}}{h\pi|Z_{i,h}|} \sin \frac{h\delta}{2} \sin \frac{h\pi}{2} \sin(h\omega t - \varphi_h)^2}$$

(38)

(39)

$$I_{1\text{rms}} = \frac{4V_{in}}{\pi|Z_{i,1}|} \sin \delta$$

$$I_{3\text{rms}} = \frac{2\sqrt{2}V_{in}}{\pi|Z_{i,1}|} \sin \delta$$

(40)
Fig. 9. Turning-on currents versus output power for switch $S_3$ and $S_1$ ($V_{in} = 50$ V, $k_s = 0.95$, and $N = 1.1$).

$$I_{3rms} = \frac{4V_{in}}{\pi |Z_{i,1}|} \sin \frac{\delta}{2} \sqrt{\frac{1}{2\pi} \left(\sin(\omega t - \varphi_1)\right)^2 d\omega t}$$

$$= \frac{2\sqrt{2}V_{in}}{\pi |Z_{i,1}|} \sin \frac{\delta}{2}.$$  

(41)

Figs. 9 and 10 show the turning-on currents of switches $S_1$ and $S_3$ that are calculated from (36) and (37) for various output powers and input voltages, respectively. In Fig. 9, the turning-on currents of $S_1$ and $S_3$ versus the load conditions are plotted for different tuning factors of the parallel network, with a fixed tuning factor of the series network $k_s = 0.95$. The input voltage is assumed as 50 V; the transformer turns ratio is 1.1. Evidently, a larger $k_p$ is favorable for more turning-on current and correspondingly wider load range. However, this is also in price of a large circulating current and increased conduction losses. In Fig. 10, the turning-on currents of switches $S_1$ and $S_3$ versus the input voltage are plotted. It is noted that, for high input line, the ZVS of $S_3$ can be lost because less current is available to discharge the snubbers of the switches.

IV. NEW PSM

The logic block diagram of the new phase-shift modulator is shown in Fig. 11(a). It consists of one D Flip-Flop (FF) and one RS FF, fast speed comparators, and a number of logic gates. The mechanism of this modulator is explained with the operation waveforms shown in Fig. 11(b).

In Fig. 11(b), the error (modulation) signal $\nu_m$ from the controller is compared with the triangle waveform signal $\nu_c$ of the peak-to-peak voltage $V_{p-p}$ and frequency $2f_o$. The leading edge modulation determines the rising edge of the pulse for switch $S_1$, while the falling edge modulation determines the rising edge of the pulse for switch $S_3$. For each of the pulse, the width is half of the switching period ignoring the duty ratio loss because of the dead time and other considerations. The D FF generates a square waveform of duty ratio of 0.5 and is exactly in phase with the triangle waveform $\nu_c$. The clock signal resets the RS FF at the beginning of each cycle but is set to 1 as the PWM flips high. The leading edge modulation triggers the switch $S_1$. The pulse is generated with the following logic operation:

$$S_1 = \text{XOR} \left\{ Q(RS), \overline{Q(D)} \right\}.$$  

(42)

Similarly, the falling edge modulation is used for $S_3$, which is generated with (43), where $Q'(RS)$ is given by (44)

$$S_3 = \text{XOR} \left\{ Q'(RS), \overline{Q(D)} \right\}$$

(43)

$$Q'(RS) = \text{AND} \left\{ Q(RS), \overline{\text{NOT}(PWM)} \right\}.$$  

(44)
Here, the PWM width is equal to the phase shift $\alpha$ of the two pulses for the two legs, which are Legs A and B. For a change on the error signal $\Delta \alpha$, the dual-edge modulation will push the $S_1$ and $S_3$ simultaneously in the opposite direction for $\Delta \alpha/2$. Therefore, the change of the modulation signal from the voltage mode controller will not cause the shift of the switching function of the full bridge in Fig. 3. In other words, the phase angle of the bridge voltage is independent of the modulation signal. Therefore, the phase angle of the output voltage of the resonant inverter is independent of the change of the modulation signal $\nu_{m}$, which dynamically changes to regulate the magnitude of the output voltage against the variation of the input line, load, and CSC as in a multiple inverter system. The theoretical range of the control angle for the PSM is from 0 to $\pi$.

The simulation waveforms of the inverter systems with the new PSM and the conventional PSM are shown in Fig. 12 for two different modulation levels [12]. The bridge voltages of the inverter with the new PSM are shown in Fig. 12(a) for two different control efforts. The corresponding output voltages are shown in Fig. 12(b). The control effort is the output of the voltage-feedback control together with the CSC in a multiple inverter system. It is time varying and depends on load/line variations. In a multiple inverter module system, the current distribution error also causes the difference of the control efforts for the individual inverter modules. Evidently, the phase angle of the output voltage is fixed with the new PSM, regardless of the phase-shift control. In contrast, for the conventional PSM, the bridge voltage and output voltage phases shift for the different control efforts, as shown in Fig. 12(c) and (d), respectively. Such a PSM method is not fit for the synchronization of the control and output voltages. Neither it is suitable for the synchronization of multiple inverter module applications where the multiple inverter modules are operated in parallel for the same HFAC bus.
TABLE I
CIRCUIT PARAMETERS OF THE FULL-BRIDGE RESONANT INVERTER

<table>
<thead>
<tr>
<th>Specifications &amp; Component Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Inputs</td>
<td>Minimum input 35V, Maximum input 75V, Rated input 48V</td>
</tr>
<tr>
<td>Outputs</td>
<td>Output 28V (RMS), Power 150W</td>
</tr>
<tr>
<td>Frequency</td>
<td>1MHz</td>
</tr>
<tr>
<td>Resonant Tank</td>
<td>Parallel resonant tank: ( L_s = 0.39 , \mu H ) (Magentics 55381-A2); ( C_s = 43 , \text{nF} ) (Winma FKP)</td>
</tr>
<tr>
<td></td>
<td>Series resonant tank: ( L_r = 3.9 , \mu H ) (Magentics 55381-A2); ( C_r = 7.0 , \text{nF} ) (Winma FKP)</td>
</tr>
<tr>
<td></td>
<td>Snubbers: ( C_s = 110 , \text{pF} ) (TDK X7R)</td>
</tr>
<tr>
<td>Transformer</td>
<td>8.7, TDK PC44LP22/13</td>
</tr>
<tr>
<td>Switches</td>
<td>MOSFET, IRF540</td>
</tr>
</tbody>
</table>

V. EXPERIMENTAL RESULTS
A prototype resonant inverter of 1 MHz and 150 W with the same topology, as shown in Fig. 3(a), was implemented and controlled with the new phase-shift modulator. The output voltage was regulated to 28 V (rms) with an input voltage that ranges from 35 to 75 V. The rated output power is 150 W. The quality factors of both the series and parallel resonant tanks are designed to approximately 2 for the rated load conditions. The series resonant tank is tuned to 0.95 of the operation frequency, while the parallel resonant tank is tuned to about 1.2 of the operation frequency. The specifications and the power stage parameters of the inverter are listed in Table I.

The output-voltage waveform quality is almost sinusoidal with very low THD, because of the following design considerations. 1) The quality factor is designed to be higher than 2. 2) High-order resonant tank is used. 3) The modified PSM is used to control the full-bridge conversion circuit. The adverse EMI effect is avoided because the power is delivered in the HFAC bus with low-current high-voltage sinusoidal waveform voltages with very low THD.
Fig. 13. Key waveforms with two different input voltages. (a) Input voltage of 50 V. (b) Input voltage of 35 V.

In Fig. 13(a), the bridge voltage and the resonant current waveforms are shown for an input line of 50 V. The inverter feeds an equivalent load of 5 Ω or 150 W at 28-V rms. In Fig. 13(b), the waveforms for an input voltage of 35 V are presented. As shown from these figures, all the switches turn on under zero voltage. With small snubber capacitors, all the switches achieve near zero-voltage turning off.

The output voltage $\nu_o$ and the load current $i_o$ for the nonresistive load condition are shown in Fig. 14. A spectrum analysis of the output voltage shows that the main component of the harmonic is the third harmonics. The THD increases as the input voltage increases. Over all the input-line range, the THD is less than 5%. The system efficiency is shown in Fig. 15 for the resistive load conditions. For the input voltage from 35 to 65 V, the efficiency of the proposed full-bridge resonant inverter is around 94.5%, where the power loss of the gate driver and the control stage is not counted.

In Fig. 16, the circuit schematic for the new PSM is shown. It is implemented with the D FF and a number of logic gates. In Fig. 17, the experiment waveforms of the PSM signals are presented for the two modulation levels. As can be observed, the center of the two pulses for switches $S_2$ and $S_4$ keeps a constant interval (129.4 ns) from the front edge of the clock signal, regardless of the modulation level $\nu_m$. In Fig. 18, the features of the modified PSM are demonstrated. For two different input voltages, the modulation index $\nu_m$ is different, which means that the phase-shift angle between Legs A and B is different. However, the phase angle $\xi$ of the output voltage of the resonant inverter keeps the same. The intervals between the zero crossing of the output voltage and the front edge of the clock are the same for the two different input voltages (or, more generally, the modulation level $\nu_m$).

The advantage of such modulation is that the phase angle of the output voltage of individual inverter modules is independent of the voltage-feedback control. In the HFAC power distribution architectures, by using the same clock signal to synchronize all the inverter modules in parallel, the phase angles of the inverter outputs can be synchronized, while the output voltage magnitudes of these inverters can be regulated with the CSC to achieve an evenly current distribution among the multiple inverter modules.

VI. CONCLUSION

The analysis and design of a high-frequency full-bridge resonant inverter for the HFAC power distribution architecture
Fig. 16. Circuit schematics of the modified PSM.

Fig. 17. PSM signals for two different modulation signals. (a) $V_m = 0.1$. (b) $V_m = 0.7$.

Fig. 18. Waveforms showing constant phase angle of the output voltage regardless of the modulation level. (a) $V_m = 0.6$. (b) $V_m = 0.9$.

was presented with a new PSM. The resonant inverter in the ac power distribution system is complicated by the factors that the load impedance is unknown and the input-line voltage varies for a wide range. Because it is mandatory to take into consideration the load of different power factors as well as the wide input range, the soft-switching design of the resonant inverter is challenging. A full-bridge series-parallel resonant
inverter that is controlled with a novel PSM was proposed in this paper for the HFAC power distribution architectures. Analysis and design curves for the resonant inverter with a general load model were developed. A new PSM method was proposed with which the phase angle of the output voltage of the resonant inverter is independent of the voltage-feedback control. Experiment verifications on one prototype inverter of 1 MHz and output power of 150 W at 28-V rms output voltage were presented. The proposed new PSM-controlled series-parallel resonant inverter has the following advantages: 1) High efficiency is achieved over wide input line and load variations; 2) high waveform quality with very low THD is guaranteed because of the high order, high tuning factor, and high quality factor resonant tank, and the symmetrical PSM; and 3) The phase angle of the inverter output voltage is independent of the change of the modulation signals. Therefore, a magnitude current-sharing-control scheme can be used to achieve an evenly current sharing in a multiple inverter system.

REFERENCES


Paresh C. Sen (M’67–SM’74–F’89–LF’04) was born in Chittagong, Bangladesh. He received the B.Sc. degree in physics (with honors) and the M.Sc. (Tech.) degree in applied physics from the University of Calcutta, Calcutta, India, in 1958 and 1961, respectively, and the M.A.Sc. and Ph.D. degrees in electrical engineering from the University of Toronto, ON, Canada, in 1965 and 1967, respectively.

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Dr. Sen was the recipient of the IEEE Canada Outstanding Engineering Educator Award in 2006 for his outstanding contributions over four decades as an Author, Teacher, Supervisor, Researcher, and Consultant. He received the Prize Paper Award from the Industrial Drives Committee for his technical excellence at the IEEE Industry Applications Society Annual Meeting in 1986. He has served as an Associate Editor of the IEEE TRANSACTIONS ON INDUSTRIAL ELECTRONICS (1975–1982) and as the Chairman of the Technical Committees on Power Electronics (1979–1980) and Energy Systems (1980–1982) of the IEEE Industrial Electronics Society. He served as a Natural Science and Engineering Research Council of Canada Scientific Liaison Officer evaluating the university–industry coordinated projects (1994–1999). As an Emeritus Professor, he continues to be active in research and in several IEEE societies.
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