A Novel Resonant Converter Topology for DC-to-DC Power Supply

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A class-E dc-to-dc converter with half-wave controlled current rectifier is proposed. Its output voltage is controlled by the conduction angle of the rectifier switch at constant switching frequency. Zero voltage switching for all the switches can be maintained from full load to no load. Its steady state characteristics are analyzed and the effects of the circuit parameters are studied. Some extensions of the proposed converter are also discussed. The analysis is verified by PSPICE simulation and an experimental prototype.

The size of electronic equipment is shrinking steadily. The size of its power supply has to be reduced as well. For the switching power supply, the effective way to reduce the size is to increase the switching frequency so that the size of the filter capacitor and inductor, as well as the transformer, can be reduced as they occupy a large portion of the overall size.

In the PWM (pulsewidth modulated) converter [1-4], the active switch is turned on and off at controlled instant and the switch voltage and current change almost as a step. Because of the finite switching time, large switch current and switch voltage are present at the same time during switching turn on and turn off interval. The switching loss is, therefore, induced for every switching action. At high switching frequency, the switching loss becomes very large. Furthermore, the existence of the layout inductance and junction capacitance of the semiconductor devices causes the electronic switch to inductively turn off and capacitively turn on. The consequence is that large voltage and current spikes are induced and extra power is lost. High switching loss reduces the efficiency of the switching power supply and also requires larger heat sink for the switches. Therefore, it is not suitable to reduce the size of the PWM switching converter by increasing the switching frequency further.

In the resonant converters [5-8], the current flowing through the switches are quasi-sinusoidal and the switching loss is small because the switches are turned on and/or off at zero voltage and/or zero current. Therefore, high switching frequency is achieved and the size of the switching power supply can be reduced. Among the resonant converters, the class-E converters [8-14], both operated at constant switching frequency and variable switching frequency, offer particular advantages in high frequency operation because of extremely low switching loss and simple topology.

In this work, various topologies based on the class-E converter are at first reviewed and their advantages and disadvantages are addressed in next section. In Section III, a class-E converter topology is proposed by employing the concept of the half-wave controlled current rectifier. This topology retains all the merits of the conventional class-E converter and at the same time, overcomes the disadvantages, namely, variable switching frequency control and limited load variation range for lossless operation. The steady state characteristics of the proposed converter are analyzed and the effects of the various circuit parameters are studied in Section IV. Some extensions of the proposed converter are discussed in Section V. In Section VI, PSPICE simulation is made and experimental prototype is built to verify the feasibility of the new converter and the validity of the analysis. Both the analysis and experiment show that the proposed...
topology can maintain the desirable characteristics of zero voltage switching over the entire operating range at constant switching frequency. Section VII is the conclusion.

II. REVIEW OF CLASS-E DC-TO-DC CONVERTERS

The basic class-E dc-to-dc converter [8] is shown in Fig. 1. This topology is suitable for high switching frequency operation because 1) the turn on loss of the metal-oxide-semiconductor field-effect transistor (MOSFET) is zero, 2) the turn off loss is minimized by the parallel capacitor $C_1$, 3) the body diode of the MOSFET can be utilized, and 4) the parasitic capacitor of the MOSFET can also be included as part of the external capacitor. In addition, the topology configuration is very simple. The dc output voltage is obtained by rectifying the resonant current $i$. When the load resistor changes, the switching frequency is changed accordingly to regulate the amplitude of the resonant current and hence to maintain the output voltage unchanged. Unfortunately, the switching frequency has to change over a wide range to accommodate the worst combination of the load resistor and supply voltage variation.

Another problem associated with the class-E converter (Fig. 1) is that at small output current, or at large load resistor, the resonant current is unable to discharge the capacitor $C_1$ completely and zero voltage turn on can not be maintained for $S_1$. It is shown in [8, 10] that zero voltage turn on can only be maintained for

$$0 < R < R_{\text{max}}$$

(1)

where $R$ is the load resistor and the value $R_{\text{max}}$ is dependent on the circuit variables. When load resistor $R$ is larger than $R_{\text{max}}$, zero voltage turn on for $S_1$ will no longer be present. The circuit cannot operate properly for no load conditions.

Several attempts have been made to solve these problems of the class-E dc-to-dc converter.

In [10], an inductor and a coupling capacitor are added at the input of the rectifier, as shown in Fig. 2. The inductor $L_2$ is used as impedance inverter so that zero voltage switching can be maintained for larger variation range of the load resistor. The variation range of the switching frequency is also reduced as compared with the conventional class-E converter (Fig. 1). The circuit can operate at no load condition. Unfortunately, there still exists some value of the load resistor at which zero voltage switching for $S_1$ is lost and the switching frequency has still to vary about 12% to keep the output voltage at a desirable value when load resistor changes.

One limitation of variable switching frequency control is that the filter inductor and filter capacitor have to be designed according to the lowest switching frequency. Another drawback is that the spectrum of the noise generated by the switching converter varies over a wide frequency range. In some circumstances, the switching frequency of the power supply must be kept at certain value to avoid interference with other parts of the electronic equipment. It is, therefore, desirable to keep the switching frequency constant.

Two techniques have been proposed to operate the class-E dc-to-dc converter at the fixed switching frequency [11, 12].

The output voltage of the class-E dc-to-dc converter is dependent on both the switching frequency and the resonant frequency. In [11], a controlled capacitor, called the "switch-controlled capacitor" (SCC), is used to change the equivalent resonant frequency of the resonant branch in the class-E dc-to-dc converter so as to regulate the output voltage at fixed switching frequency. Its circuit topology is given in Fig. 3. When switch $S_2$ conducts all the time, the capacitor $C_2$ is short circuited all the time and the resonant frequency is determined by $L$ and $C$ as $\omega_r = \sqrt{1/(L_rC_r)}$.

When the auxiliary switch $S_2$ does not conduct at all, capacitor $C_2$ acts as an element of the resonant circuit together with $L_r$ and $C_r$. The corresponding resonant frequency is $\omega_r = \sqrt{(1/L_rC_r) + (1/L_rC_2)}$. By changing the conduction angle of switch $S_2$, the equivalent resonant frequency is changed and the output voltage can also be changed. Therefore, the output voltage can be regulated at the constant switching frequency. Unfortunately, at light load, the characteristic of zero voltage switching for $S_1$ is lost. In addition, the circuit cannot operate at no load condition.

In order to control the class-E dc-to-dc converter at constant switching frequency and also keep zero voltage switching from no load to full load, two identical conventional class-E inverters are combined together with common input and output terminals [12], as shown in Fig. 4, where it is required that $C_1 = C_2$, $L_{r1} = L_{r2}$, $C_{r1} = C_{r2}$ and $L_{f1} = L_{f2}$. The output of
these two class-E inverters, i.e., the resonant current $i_1$ and $i_2$ are vector added together and then rectified to obtain the dc output. The output voltage is controlled by the phase difference between the drive signals for $S_1$ and $S_2$. When the drive signals for $S_1$ and $S_2$ are in phase, the current $i_1$ and $i_2$ are also in phase and with same amplitude because the two class-E inverters are identical. The output current of the combined converter $i$ is large and the output voltage is high. When the drive signals for $S_1$ and $S_2$ are out of phase, $i_1$ and $i_2$ are also out of phase and with same amplitude because of symmetry. The output current $i$ is equal to zero, so that the output voltage is zero. By changing the phase shift between the drive signals for $S_1$ and $S_2$, the phase angle between $i_1$ and $i_2$ and the amplitude of $i_1$ and $i_2$ are also changed so that the output voltage is regulated. Using this technique, the output voltage can be regulated at fixed switching frequency and the desirable zero voltage switching for both switches can be maintained from full load to no load. The problem of this scheme is twofold. One is that there are too many components, two input dc choke inductors, two resonant branches. The other is that two resonant branches $L_{r1} - C_{r1}$ and $L_{r2} - C_{r2}$ should be identical and the capacitor in parallel with the two switches $C_1$ and $C_2$ should also be identical to ensure symmetrical operation of the converter. This is very difficult in the practical circuit because it is very difficult to control the parasitic parameters which are utilized in the high frequency operation.

From the above analysis, it is, therefore, worthwhile to investigate new class-E dc-to-dc converter topologies that can keep the advantages of the conventional class-E dc-to-dc converter, mainly low switching loss, and at the same time, eliminate its drawbacks, i.e., variable switching frequency control and limited load variation range for lossless operation. The new topology should also be simple. This is really the objective of this work.

III. PROPOSED CLASS-E DC-TO-DC CONVERTER WITH HALF-WAVE CONTROLLED CURRENT RECTIFIER

Let us take a close look at the class-E dc-to-dc converter, as shown in Fig. 1. The output voltage is obtained by rectifying the resonant current $i$. For the positive half cycle of this current, diode $D_3$ conducts and the energy is transferred from the resonant branch to the load. For the negative half cycle, diode $D_2$ conducts and no energy is transferred to the load. When the switching frequency or the equivalent resonant frequency changes, the amplitude of the output current $i$ changes so that the energy delivered to the load changes and consequently, the output voltage changes.

There is another method to regulate the energy delivered to the load resistor. Instead of putting an uncontrolled current rectifier ($D_2$ and $D_3$) at the output of class-E inverter, a controlled current rectifier can be used to control the average current delivered to the load resistor, as shown in Fig. 5, where $S_2$, $D_2$, $C_2$, and $D_3$ consist of the half-wave controlled current rectifier. As compared with the conventional class-E dc-to-dc converter, switch $S_2$ is introduced to control the energy delivered to the load resistor and capacitor $C_2$ is used to ensure zero voltage switching of $S_2$. All the other parts of the converter are the same. In the practical circuit, $S_2$ and $D_2$ are composed of a MOSFET and its intrinsic diode, and $C_2$ is partly composed of its parasitic capacitor, $C_{ds}$.

The basic idea of this circuit was proposed in our earlier paper in 1992 [13], but the detailed operation principle and the steady state characteristics were not presented.

A similar topology was also presented later in 1993 in [14]. However, the paper did not discuss the various aspects of the topology and the analysis in that paper was not complete.

In the present paper, the proposed class-E dc-to-dc converter has been thoroughly investigated. The operating principle is explained and the mechanism of zero voltage switching for all the switches is outlined. The steady state characteristics are also analyzed and the effects of various parameters are discussed. Some extensions of the proposed circuit are also proposed. Computer simulation by PSPICE and experimental results are provided to verify the feasibility of the circuit and the validity of the analysis.

Assumptions

The operation of the class-E dc-to-dc converter with half-wave controlled current rectifier can be explained as follows. Assume the following.

1) The filter inductor $L_f$ and filter capacitor $C_f$ are large enough so that the input current $i_i$ and output voltage $V_0$ are pure dc.
2) The switches are ideal ones with no transient time and no loss.
3) The losses in the circuit are neglected.
4) The resonant branch, \( L, -C \), is a high \( Q \) series tuned network. The harmonics of the resonant current \( i \) is negligible.
5) The circuit operates at lossless mode, i.e., zero voltage switching for \( S_1 \) is maintained.

Notations

Typical waveforms are plotted in Fig. 6. The notations used in the figure are explained as follows:

- \( i_1 \) is the sum of current flowing through \( S_1, D_1 \), and \( C_1 \),
- \( e_1 \) is the voltage across the inverter switch \( S_1 \),
- \( \gamma \) is the off interval of \( S_1 \),
- \( \phi \) is the turn off instant of \( S_1 \),
- \( -\mu \) is the turn on instant of \( S_1 \),
- \( \beta \) is the conduction angle of the rectifier switch and is the control variable,
- \( e_2 \) is the voltage across the rectifier switch \( S_2 \),
- \( i_{C1} \) is the current flowing through \( C_1 \).

The inverter switch \( S_1 \) operates at 50% duty ratio. The gate drive signal of the rectifier switch \( S_2 \) is synchronized with the resonant current \( i \). \( S_2 \) is turned on when the resonant current changes polarity from negative to positive. Just before the resonant current changes direction, it flows through diode \( D_2 \). The equivalent circuit is shown in Fig. 7(a). The current direction shown in the figure denotes the actual one. The gate signal for \( S_2 \) can be supplied at this time. When the resonant current changes from negative to positive at \( \theta = \pi / 2 \), \( S_2 \) conducts and the current flows through \( S_2 \), as shown in Fig. 7(b). Therefore, zero voltage turn on for \( S_2 \) can always be achieved as the current always commutates from \( D_2 \) to \( S_2 \). \( S_2 \) is turned off after it conducts for a certain conduction angle, \( \beta \) (the control variable), defined as:

\[
\beta = \frac{T_{on}}{T_S} \cdot 360^\circ 
\]  

where \( T_S \) is the switching period, \( T_{on} \) is the on time of switch \( S_2 \). After \( S_2 \) is turned off at \( \theta = \pi / 2 + \beta \), the resonant current at first charges the capacitor \( C_2 \), as shown in Fig. 7(c), and \( V_{C2} \) rises slowly. Zero voltage turn off for \( S_2 \) is thus obtained. At \( \theta = \pi / 2 + \beta + \delta \), \( V_{C2} = V_0 \), diode \( D_3 \) is forward biased and the power is delivered from the resonant tank to the load. The equivalent circuit is shown in Fig. 7(d).

The regulation of the output voltage can be described as follows.

1) When \( \beta = 0 \), which means that the rectifier switch \( S_2 \) does not conduct at all, the diode \( D_3 \) conducts for the whole positive half cycle of \( i \). The circuit behaves equivalently as the conventional class-E converter. The output voltage is high.
2) When \( \beta = 180^\circ \), which means that the rectifier switch \( S_2 \) conducts for half switching cycle, all the positive half cycle of the resonant current \( i \) flows through \( S_2 \). The negative portion of \( i \) will flow through diode \( D_3 \). The output of the inverter stage is equivalently short circuited. The diode \( D_3 \) will never conduct and the output voltage is zero.
3) When the conduction angle \( \beta \) is between 0 and \( 180^\circ \), part of the positive resonant current \( i \)
flows through $S_2$ and part of $i$ flows through $D_3$. The averaged current through diode $D_3$ is somewhere between zero and the value corresponding to $\beta = 0$.

When the conduction angle $\beta$ is varied, the average current through diode $D_3$, i.e., the output current, is changed and the output voltage will also be changed. Therefore, the output voltage of the converter can be regulated at a fixed switching frequency by modulating the conduction angle $\beta$ of the rectifier switch $S_2$.

Another advantage of the class-E converter with controlled current rectifier, given in Fig. 5, is that zero voltage switching for all the switches can be maintained from full load to no load. It is already shown that zero voltage switching for the rectifier switch $S_2$ can be maintained for entire load range. The following analysis shows that zero voltage switching for the inverter switch $S_1$ can also be maintained from no load to full load. In the analysis, it is assumed that the steady state output voltage remains constant when load resistor changes.

When the load resistor is small, the conduction angle $\beta$ should also be kept small so that the conduction angle of diode $D_3$ is large to provide higher output current. In this case, the equivalent load to the class-E inverter satisfies the lossless operation condition, i.e., see (1). When the load resistor increases, the output current reduces because the output voltage is kept constant. The diode $D_3$ conducts for shorter period of time and the conduction angle for $S_2$ is thus increased. The equivalent load appearing at the output of the inverter stage will not increase, but it actually be reduced. In the extreme case, when the output is open circuit and the load current is zero, $S_2$ conducts for the whole positive half cycle of $i$. The inverter output is actually short circuited. Therefore, when the load resistor changes from its minimum to maximum (open circuit), the equivalent resistor appeared at the inverter output reduces from its maximum to zero and (1) is always satisfied. Therefore, zero voltage switching can always be maintained for the inverter switch $S_1$.

It is shown from the above analysis that the proposed class-E dc-to-dc converter with half-wave controlled current rectifier can keep the switching frequency constant and at the same time keep the desirable zero voltage switching characteristics from no load to full load.

IV. STEADY STATE ANALYSIS

In order to investigate the characteristics of the proposed class-E converter with half-wave controlled current rectifier, the steady state characteristics is analyzed in this section. For simplicity, the Fourier Series Expansion method is used and only the dc and fundamental components are considered. From the waveforms of Fig. 6, the notation and assumptions described in the previous section, the resonant current $i$ can be expressed as:

$$i = -I_p \cos \theta$$  \tag{3}

where $I_p$ is the peak value of the resonant current.

From Kirchhoff's current law and from Fig. 5 and (3), we have

$$i_1 = I_1 - i = I_i + I_p \cos \theta.$$  \tag{4}

The voltage across the inverter switch $S_1$ can, therefore, be calculated as:

$$e_1 = \left\{ \begin{array}{ll}
\frac{1}{\omega C_1} \int_{\phi}^{\theta} i_1 d\theta = \frac{1}{\omega C_1} [I_i (\theta - \phi) + I_p (\sin \theta - \sin \phi)]
& \text{for } \phi \leq \theta \leq \phi + \gamma \\
0 & \text{elsewhere.}
\end{array} \right.$$  \tag{5}

For lossless operation we have

$$e_1(\theta = \phi + \gamma) = 0.$$  \tag{6}

From (5) and (6), the following equation is obtained:

$$\gamma I_i = I_p [\sin \phi - \sin (\phi + \gamma)].$$  \tag{7}

Expanding the voltage $e_1$ into Fourier series and only considering the dc and fundamental components, the following equation is obtained:

$$e_1 = E_1 + E_{11} \cos \theta + E_{12} \sin \theta$$  \tag{8}

where

$$E_1 = \frac{I_p}{2 \pi \omega C_1} \left[ \frac{I_i \gamma^2}{I_p^2} - \cos (\phi + \gamma) + \cos \phi - \gamma \sin \phi \right]$$  \tag{9}

$$E_{11} = \frac{I_p}{\pi \omega C_1} \left\{ \frac{I_i}{I_p} \left[ \gamma \sin (\phi + \gamma) + \cos (\phi + \gamma) - \cos \phi \right.ight.$$  
$$- \frac{1}{4} \cos 2(\phi + \gamma) - \sin \phi \sin (\phi + \gamma)$$  
$$- \frac{1}{4} \cos 2\phi + \frac{1}{2} \left. \right] \right\}$$  \tag{10}

$$E_{12} = \frac{I_p}{\pi \omega C_1} \left\{ \frac{I_i}{I_p} \left[ -\gamma \cos (\phi + \gamma) + \sin (\phi + \gamma) - \sin \phi \right.ight.$$  
$$- \frac{1}{4} \sin 2(\phi + \gamma) + \sin \phi \cos (\phi + \gamma)$$  
$$- \frac{1}{4} \sin 2\phi + \frac{\gamma}{2} \left. \right] \right\}.$$  \tag{11}

Because the voltage across the input inductor has no dc component, the average voltage across $S_1$ is the same as the supply voltage, i.e.,

$$V_s = E_1 = \frac{I_p}{2 \pi \omega C_1} \left[ \frac{I_i \gamma^2}{I_p^2} - \cos (\phi + \gamma) + \cos \phi - \gamma \sin \phi \right].$$  \tag{12}
According to the waveforms shown in Fig. 6, at $\theta = \pi/2 + \beta$, the rectifier switch $S_2$ is turned off. The capacitor $C_2$ is charged by the resonant current $i$ and the voltage $e_2$ rises. At $\theta = \pi/2 + \beta + \delta$, $e_2$ reaches $V_0$ and $D_3$ is turned on. The voltage $e_2$ is clamped by $D_3$, as shown in Fig. 6. At $\theta = 3\pi/2 + \delta$, $e_2$ falls to zero. The expression for $e_2$ can be found as:

$$
e_2 = \begin{cases} 
\frac{I_p}{\omega_s C_2} (\cos \beta - \sin \theta) & \frac{\pi}{2} + \beta \leq \theta \leq \frac{\pi}{2} + \beta + \delta \\
V_0 & \frac{\pi}{2} + \beta + \delta \leq \theta \leq \frac{3\pi}{2} \\
V_0 - \frac{I_p}{\omega_s C_2} (1 + \sin \theta) & 3\pi/2 \leq \theta \leq 3\pi/2 + \alpha \\
0 & \text{elsewhere}
\end{cases}
$$

where $\delta$ and $\alpha$ are expressed as:

$$
\delta = \cos^{-1} \left( \cos \beta - \frac{V_0 \omega_s C_2}{I_p} \right) - \beta, \\
\alpha = \cos^{-1} \left( 1 - \frac{V_0 \omega_s C_2}{I_p} \right).
$$

The Fourier expansion of $e_2$ can be expressed as:

$$
e_2 = E_{21} \cos \theta + E_{22} \sin \theta
$$

where the high-order harmonics are neglected. $E_{21}$ and $E_{22}$ are expressed as:

$$
E_{21} = \frac{I_p}{4\pi \omega_s C_2} \left[ 4 \cos \beta [\cos(\beta + \delta) - \cos \beta] + \cos 2\beta - \cos 2(\beta + \delta) - (1 - \cos 2\alpha) \right] \\
- \frac{1}{\pi} V_0 [1 + \cos(\beta + \delta)] + \frac{1}{\pi} \left( V_0 - \frac{I_p}{\omega_s C_1} \right) (1 - \cos \alpha)
$$

$$
E_{22} = \frac{I_p}{4\pi \omega_s C_2} \left[ 4 \cos \beta [\sin(\beta + \delta) - \sin \beta] - \sin 2(\beta + \delta) + \sin 2\beta - \sin 2\alpha - 2(\delta + \alpha) \right] \\
- \frac{1}{\pi} V_0 \sin(\beta + \delta) - \frac{1}{\pi} \left( V_0 - \frac{I_p}{\omega_s C_2} \right) \sin \alpha.
$$

The voltage across the resonant branch, $L_r - C_r$, is found as:

$$
e = E + I_p \left( \omega_s L_r - \frac{1}{\omega_s C_r} \right) \sin \theta.
$$

From Kirchhoff's voltage law, the following equation for the fundamental component is derived:

$$
E_{11} \cos \theta + E_{12} \sin \theta = I_p \left( \omega_s L_r - \frac{1}{\omega_s C_r} \right) \sin \theta + E_{21} \cos \theta + E_{22} \sin \theta.
$$

Two equations can be derived from (20) as:

$$
E_{11} = E_{21},
$$

$$
E_{12} = I_p \left( \omega_s L_r - \frac{1}{\omega_s C_r} \right) + E_{22}
$$

where the expressions for $E_{11}$, $E_{12}$ and $E_{21}$, $E_{22}$ can be found in (10), (11), (16) and (17). In the analysis, the load resistor $R$, supply voltage $V_s$, and the desired output voltage $V_0$ are known. Therefore, the input current $I_i$ can be calculated from the power balance as:

$$
I_i = \frac{V_0^2}{RV_s}.
$$

There are four unknowns, $I_p$, $\beta$, $\phi$, and $\gamma$ in four equations (7), (12), (20) and (21). Given the supply voltage $V_s$, load resistor $R$ and the desired output voltage $V_0$, as well as the circuit variables, $L_r$, $C_r$, $C_1$, $C_2$, the required conduction angle $\beta$, $I_p$, $\phi$, and $\gamma$ can all be obtained.

The voltage stress for inverter switch $S_1$ can be obtained from (23) by setting $d e_1/d \theta = 0$ and calculated as:

$$
e_{1\text{max}} = \frac{1}{\omega_s C_1} [I_i (\theta_m - \phi) + I_p (\sin \theta_m - \sin \phi)].
$$

where $\theta_m$ is expressed as: $\theta_m = \cos^{-1}(I_i/I_p)$.

The peak voltage across the rectifier switch is the output voltage, i.e.,

$$
e_{2\text{max}} = V_0.
$$

The rms value of the current flowing through $S_1$ is calculated as follows:

$$
I_{1\text{rms}} = \sqrt{\frac{1}{2\pi} \int_{-\pi}^{\pi} (I_i + I_p \cos \theta)^2 d\theta}
$$

$$
= \sqrt{\frac{1}{2\pi} \left[ I_i^2 (\phi + \mu) + 2 I_i I_p (\sin \phi \sin \mu) + I_p^2 \left( \frac{\phi + \mu}{2} - \frac{\sin 2\phi - \sin 2\mu}{4} \right) \right]}
$$

where $\mu$ is calculated from $\mu = \cos^{-1}(-I_i/I_p)$. 

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The rms value of the current flowing through \( S_2 \) is

\[
I_{2\text{max}} = \sqrt{\frac{1}{2\pi} \int_{\pi/2}^{\pi/2 + \beta} (-I_p \cos \theta)^2 d\theta} = I_p \sqrt{\frac{1}{2\pi} \left( \frac{\beta}{2} - \frac{\sin 2\beta}{4} \right)}. \tag{25}
\]

Based on the above equations, the characteristics of the class-E dc-to-dc converter with half-wave controlled current rectifier can be analyzed. In the analysis, the parameters for the resonant branch, \( L_r \) and \( C_r \), are defined as the resonant frequency \( f_r \) (\( f_r = 1/(2\pi L_r C_r) \)) and the branch impedance \( X(X = \omega_f L_r - (1/\omega_f C_r)) \). The \( L_r \) and \( C_r \) can be calculated easily from \( f_r \) and \( X \) as:

\[
L_r = \frac{X}{2\pi f_r \left[ 1 - \frac{(2\pi f_r)^2}{(2\pi f_s)^2} \right]} \tag{27}
\]

\[
C_r = \frac{1}{(2\pi f_s)^2 L_r}.
\]

At first the effect of the desired output voltage is examined. Fig. 8(a) gives the relationship between the output current and the control input (the conduction angle \( \beta \)) for different output voltage, \( V_o = 10 \text{ V}, 20 \text{ V}, 40 \text{ V}, \) and \( 60 \text{ V} \). The circuit parameters are: \( V_s = 30 \text{ V}, C_1 = 5 \text{ nF}, C_2 = 2.5 \text{ nF}, X = 10 \Omega, f_s = 0.9 \text{ MHz}, f_r = 1 \text{ MHz} \). It shows that the output voltage can be kept at the desired value by changing the conduction angle \( \beta \) when the load current \( I_0 \) changes. The effect of the output voltage on the output current is very large, e.g., when the output voltage changes from \( 10 \text{ V} \) to \( 60 \text{ V} \) at \( \beta = 50^\circ \), the corresponding output current changes only from about 1.35 A to about 1.55 A. It is also noticed that the curves do not go down to \( \beta = 0 \). This is because the output of the class-E inverter can only be considered as weak current source. It cannot provide high enough current to keep the output voltage constant (as the output voltage is assumed constant in the above analysis) when the load resistor becomes too small. In the analysis, this fact is shown as no real solution for smaller \( \beta \). It is also noted that the conduction angle \( \beta \) will never reach 180° because of the charging time for \( C_2 \). The largest possible \( \beta \) depends on the value of \( C_2 \) and amplitude of \( i \). It also depends on the value of the output voltage \( V_o \). When \( V_o \) is high, the time used to charge \( C_2 \) is longer and the largest possible \( \beta \) is less and vice versa.

The effect of the output voltage on the voltage stress of the inverter switch \( S_1 \), \( e_{1\text{max}} \), is also analyzed, as shown in Fig. 8(b). It shows that when the desired output voltage becomes higher, the voltage stress also becomes higher. But again, the influence is not very significant, i.e., when the desired output voltage increases from \( 10 \text{ V} \) to \( 60 \text{ V} \), the maximum voltage across \( S_1 \) increases from about 115 V to about 130 V. The other observation is that for a certain output voltage, the voltage stress will increase a little bit (about 14% for \( V_o = 60 \text{ V} \) and about 2% for \( V_o = 10 \text{ V} \) when the conduction angle \( \beta \) increases. It is noted that when \( C_2 \) is charged up, the resonant frequency is changed due to the presence of \( C_2 \). Therefore, \( e_{1\text{max}} \) is different which is similar to the case of conventional class-E converter at different switching frequency. At high output voltage, the time needed to charge \( C_2 \) is longer. Therefore, the effect of \( C_2 \) on the resonant frequency is larger and \( e_{1\text{max}} \) varies more as compared with a low output voltage case, as can be seen from Fig. 8(b).

The effect of the impedance of the resonant branch \( X \) is also studied. The circuit parameters in the analysis are: \( V_s = 30 \text{ V}, V_o = 30 \text{ V}, C_1 = 5 \text{ nF}, C_2 = 2.5 \text{ nF}, f_s = 0.9 \text{ MHz}, f_r = 1 \text{ MHz} \). The effect of \( X \) on the output current is given in Fig. 9(a). It shows that the output current increases when \( X \) reduces. This phenomenon is understandable because for smaller \( X \), the amplitude of the resonant current increases and therefore, the output current also increases. The voltage stress for \( S_1 \) also increases when the impedance of the resonant branch reduces, as shown in Fig. 9(b).
Fig. 9. Effect of resonant impedance on (a) output current and (b) voltage stress of $S_1$ at $V_i = 30$ V, $V_o = 30$ V, $C_1 = 5$ nF, $C_2 = 2.5$ nF, $f_r = 0.9$ MHz, $f_s = 1$ MHz.

Fig. 10. Effect of parallel capacitor $C_1$ on (a) output current and (b) voltage stress of $S_1$ at $V_i = 30$ V, $V_o = 30$ V, $C_2 = 2.5$ nF, $X = 10$ Ω, $f_r = 0.9$ MHz, $f_s = 1$ MHz.

The effect of the capacitor in parallel with the inverter switch $S_1$ is also analyzed with the circuit parameter: $V_i = 30$ V, $V_o = 30$ V, $C_2 = 2.5$ nF, $X = 10$ Ω, $f_r = 0.9$ MHz, $f_s = 1$ MHz. The results are plotted in Fig. 10. It is shown that when $C_1$ increases, the voltage stress on $S_1$ is reduced significantly, as given in Fig. 10(b). On the other hand, the effect of $C_1$ on the output current is not significant, as shown in Fig. 10(a). This is a good feature, which illustrates that in the class-E dc-to-dc converter with half-wave controlled current rectifier, the voltage stress of $S_1$ can be reduced by a larger $C_1$ while the output current is not sacrificed.

Similarly, the effect of other circuit parameters can also be analyzed. The effect of the capacitor in parallel with the rectifier switch $C_2$ is not significant and the resonant frequency $f_r$ has no effect on the output current, but $f_s$ will affect the voltage across $C_r$ and $L_r$. The details are not presented here.

V. SOME EXTENSIONS OF THE PROPOSED CIRCUIT

When the half-wave controlled current rectifier used in Fig. 5 is replaced by a full-wave controlled current rectifier, the class-E dc-to-dc converter with full-wave controlled current rectifier is obtained, as shown in Fig. 11, where the inverter stage is the same as that in Fig. 5. The full-wave controlled current rectifier is composed of $S_2 D_2 C_2$, $S_3 D_3 C_3$, and $D_4, D_5$.

The control signal is arranged as follows. The gate drives for $S_2$ and $S_3$ are synchronized with the resonant current $i$. $S_2$ is turned on when $i$ changes polarity from negative to positive and conducts for conduction angle $\beta$. $S_3$ is turned on when $i$ changes from positive to negative and also conducts for the conduction angle $\beta$. By changing the conduction angle $\beta$, the output voltage can be regulated in the similar manner as that of half-wave controlled current rectifier. For example, when $S_2$ and $S_3$ does not conduct at all, the circuit
behaves like the conventional class-E converter with full-wave rectifier and the output voltage is high. When \( S_2 \) and \( S_3 \) conduct as long as they can, the output voltage is zero.

Zero voltage switching for \( S_2 \) and \( S_3 \) can always be maintained from no load to full load as the parallel capacitors \( C_2 \) and \( C_3 \) can always be discharged by the resonant current \( i \) and this is independent of the load current. The switching condition for the inverter switch \( S_1 \) is similar to the half-wave controlled current rectifier and zero voltage switching can also be maintained from no load to full load.

When the isolation between the input side and the output side is necessary or large voltage gain is required, an isolation transformer can be added for both the half-wave and full-wave controlled current rectifier, as shown in Fig. 12. Fig. 12(a) is the half-wave version and Fig. 12(b) is the full-wave one. Their operating principle is similar to that of nonisolated ones and are not explained here. The capacitors at the secondary side of the transformers in Fig. 12(a) and Fig. 12(b) are used to block the dc component to avoid saturation of the transformer.

VI. COMPUTER SIMULATION AND EXPERIMENTAL VERIFICATION

The operation of the class-E dc-to-dc converter with half-wave controlled current rectifier (Fig. 5)
is simulated by PSPICE to show the feasibility of the proposed circuit. The circuit parameters used in the simulation are: \( L_r = 4.5 \, \mu \text{H}, \, C_r = 9 \, \text{nF}, \, C_1 = 5 \, \text{nF}, \, C_2 = 1.5 \, \text{nF}, \, \text{and} \, L_f = 100 \, \mu \text{H}. \) The switching frequency is selected as 1 MHz. The supply voltage \( V_s = 30 \, \text{V} \) and the output is modeled by a constant voltage source with \( V_0 = 15 \, \text{V} \).

Figs. 13 and 14 give the switching waveforms of \( S_1 \) and \( S_2 \) for the class-E converter with half-wave controlled current rectifier. Fig. 13 gives the voltage \( V_{ds1} \) and current \( I_{ds1} \) associated with the inverter switch \( S_1 \) at different conduction angle, \( \beta \). (a) \( \beta = 0 \), i.e., \( S_2 \) does not conduct at all, which is equivalent to the conventional class-E converter, (b) \( \beta = 90^\circ \) when \( S_2 \) conducts for half of the positive cycle of the resonant current \( i \), and (c) \( \beta = 144^\circ \) when the output current is very small. It is shown that when the current commutates from the antiparallel diode to \( S_1 \) (at \( t_1 \) in Fig. 13), the voltage across \( S_1 \) is zero. Zero voltage turn on is achieved. When \( S_1 \) is turned off at \( t_2 \), as shown in Fig. 13, the voltage \( V_{ds1} \) rises slowly. Zero voltage switching for \( S_1 \) can be observed for all these operating conditions. Fig. 14 gives the switching waveforms of the rectifier switch \( S_2 \) when (a) \( \beta = 90^\circ \) and (b) \( \beta = 144^\circ \), respectively. It is also illustrated that when \( S_2 \) is turned on (at time \( t_1 \)), the voltage across it is zero and when \( S_2 \) is turned off, the voltage \( V_{ds2} \) rises slowly (at time \( t_2 \) in Fig. 14). Obviously, zero voltage switching is achieved.

An experimental prototype of class-E dc-to-dc converter with half-wave controlled current rectifier is also breadboarded in order to show the feasibility of the proposed circuit and the validity of the analysis. The experimental circuit is the same as that shown in Fig. 5. The circuit parameters used in the experiment are as follows: \( V_s = 30 \, \text{V}, \, L_r = 4.4 \, \mu \text{H}, \, C_r = 9 \, \text{nF}, \, C_1 = 5.5 \, \text{nF}, \, C_2 = 2.3 \, \text{nF}, \, L_f = 106 \, \mu \text{H}, \, C = 52 \, \mu \text{F}. \) The switching frequency is fixed at 1 MHz and the output voltage is regulated at 15 V. The load resistor changes from 10 \( \Omega \) to open circuit.

Fig. 15 gives the gate drive signal and device voltage of the inverter switch \( S_1 \) at different conduction angle \( \beta \). Fig. 15(a) gives the waveforms when \( \beta = 44^\circ \) and the output current is measured as 1.46 A. Fig. 15(b) is the waveforms when \( \beta = 82^\circ \) and the output current is at 0.9 A. Fig. 15(c) shows the waveforms when the output current is zero and \( \beta = 144^\circ \). It is observed from these oscillograms that when the gate signal \( V_{GS1} \) rises to 15 V (device turn on), the voltage across \( S_1 \), \( V_{DS1} \), is zero and when the gate signal \( V_{GS1} \) falls to zero (device turn off), the voltage \( V_{DS1} \) rises slowly. Therefore, zero voltage switching for the inverter switch \( S_1 \) can be maintained for the entire output current range.

Fig. 16 gives the gate signal and the device voltage of the rectifier switch \( S_2 \) for different conduction angle \( \beta \). Fig. 16(a) and Fig. 16(b) give the waveform when \( \beta = 60^\circ \) and \( \beta = 144^\circ \), respectively. When \( \beta = 144^\circ \), the output current is zero. It can also be observed that when the gate signal for \( S_2 \) (i.e., \( V_{GS2} \)), rises to 15 V, the voltage across \( S_2 \) (i.e., \( V_{DS2} \)), is zero and when the gate signal \( V_{GS2} \) falls to zero, the voltage \( V_{DS2} \) rises from zero value. This shows that zero voltage switching is achieved for the rectifier switch \( S_2 \).

The relation between the steady state output current versus the conduction angle \( \beta \) is also measured when the output voltage is regulated at 15 V, as shown in Fig. 17. The calculated value is also plotted at the same graph. These two curves are very close. The small difference is caused by the nonideality of the prototype. It demonstrates that by changing the conduction angle of the rectifier switch, the output voltage can be kept at the desired value when the output current changes.

VII. CONCLUSION

In this paper, the present techniques of the class-E dc-to-dc converters are reviewed and their drawbacks
are addressed. A new class-E dc-to-dc converter topology with half-wave controlled current rectifier is proposed. The mechanism to maintain the zero voltage switching for all the switches is explained. The output voltage can be regulated by changing the conduction angle of the rectifier switch. The salient advantages of the new converter topologies are: 1) the switching frequency is constant, 2) the circuits can operate at no load condition, and 3) zero voltage switching for all the switches can be maintained from no load to full load.

The steady state characteristics of the proposed class-E dc-to-dc converter with half-wave controlled current rectifier is analyzed. The effects of the circuit parameters on the output current are also studied. The analysis shows that the output current can be regulated by the conduction angle of the rectifier switch to keep the output voltage constant. Some extensions of the proposed circuit, i.e., class-E dc-to-dc converter with full-wave controlled current rectifier, the isolated
class-E converters with controlled current rectifier, are also presented.

The operation of the class-E dc-to-dc converter with half-wave controlled current rectifier is simulated by PSPICE to show the feasibility of the proposed circuit. An experimental prototype is also breadboarded to demonstrate the feasibility of the proposed class-E converter and to verify the analysis. Zero voltage switching is shown clearly in the simulation and the experimental oscillograms. The measured relation between output current and conduction angle is very close to the calculated one.

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