New Class-E DC–DC Converter Topologies with Constant Switching Frequency

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Abstract— **The Class-E dc–dc converters with half-wave and full-wave controlled current rectifier are proposed. Zero voltage switching for all the switches can be maintained from full load to no load. The switching frequency is constant. When the load current changes from maximum to zero, the output voltage can be kept constant by regulating the conducting angle of the rectifier switch. The operation of the new converters is analyzed and the zero voltage switching characteristic has been demonstrated. The current gain of the circuit versus the conducting angle has been presented.**

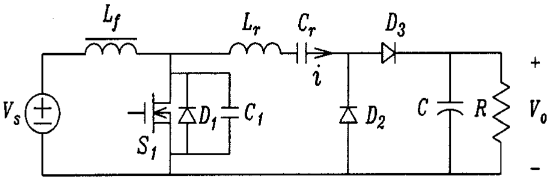

Fig. 1. Class-E dc–dc converter.

I. Introduction

THE SIZE of the 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 reactive component, 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 pulsewidth-modulated (PWM) 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 intolerable. High switching loss reduces the efficiency of the switching power supply and also requires larger heat sink for the switches. Therefore, it is difficult to reduce the size of the PWM switching converter by increasing the switching frequency further.

In the resonant converters [5]–[7], 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 converter offers particular advantages in high frequency operation because of extremely low switching loss and simple topology [8]–[11]. In this paper, various topologies based on the Class-E converter are at first reviewed and their disadvantages are addressed at first in next section. In Section III, a new kind of Class-E converter topology is proposed that can keep all the merits of the conventional Class-E converter and, at the same time, eliminate the disadvantages. Some extension of the new converter topologies are discussed in Section IV. The feasibility of the new converters are illustrated by the PSPICE simulation in Section V. Both the simulation and analysis show that the proposed topologies can maintain the desirable characteristics of zero voltage switching over the entire operating range at constant switching frequency. Section VI is the conclusion.

II. Review of Class-E Technique

The basic Class-E dc–dc converter [5] is shown in Fig. 1. It is particularly suited for high-frequency operation because the turn on loss of the MOSFET is zero, the body diode of the MOSFET can be utilized, and the parasitic capacitor of the MOSFET can also be used as part of the external capacitor. Therefore, the topology configuration is very simple.

Unfortunately, its output voltage is regulated by frequency modulation. The consequence is that the switching frequency has to be changed over a wide range to accommodate the worst combination of the load current and supply voltage variation. Another problem associated with it is that at small output current, or large load resistor, zero voltage switching can not be maintained and the circuit can not operate properly at no load. Although the latter problem can be solved by using an inductor L_2 and a capacitor C_2 at the input of the rectifier [10], as shown in Fig. 2, the switching frequency has still to be changed to regulate the output voltage.

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

In [11], a controlled capacitor, called the "switch controlled capacitor," is used to change the equivalent resonant frequency

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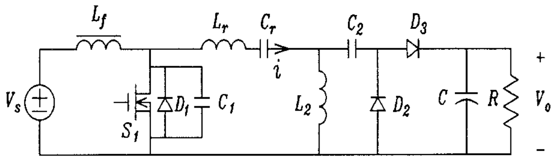


Fig. 2. Class-E dc–dc converter with inductive impedance inverter.

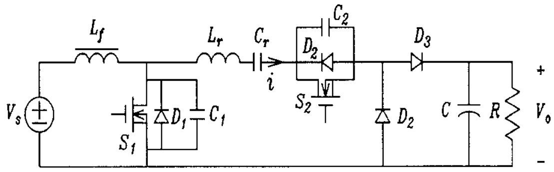


Fig. 3. Class-E dc–dc converter with switch controlled capacitor.

so as to regulate the output voltage at fixed switching frequency as the output voltage of the Class-E dc–dc converter is dependent on both the switching frequency and resonant frequency. Its circuit topology is given in Fig. 3. When the 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 $f_r = \sqrt{1/(LC)}/(2\pi)$. When the auxiliary switch S_2 does not conduct at all, the equivalent resonant capacitor is $C\|C_2$ and the corresponding resonant frequency is $f_r = \sqrt{(1/LC) + (1/LC_2)}/(2\pi)$. By changing the conducting angle of the switch S_2, the equivalent resonant frequency is changed and the output voltage is also changed. Therefore, the output voltage can be regulated at the constant switching frequency. Unfortunately, at light load, the output voltage can not be controlled and the characteristic of zero voltage switching is lost.

In order to control the Class-E dc–dc converter at constant switching frequency and also for wide load variation range, two identical conventional Class-E inverters are combined together with common input and output terminals [12], as shown in Fig. 4. 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.

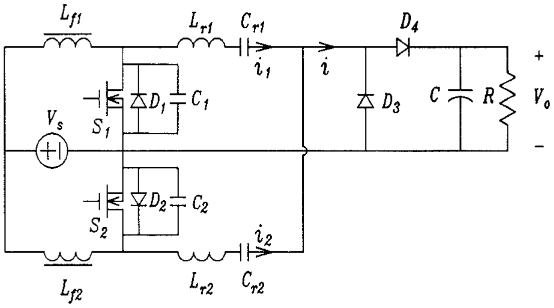


Fig. 4. Class-E dc–dc combined converter.

From the above analysis, it is, therefore, worthwhile to investigate new Class-E dc–dc converter topologies that can keep the advantages of the conventional Class-E dc–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. The new topology should also be simple. This is really the objective of this paper.

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

Let us take a close look at the Class-E dc–dc converter, as shown in Fig. 1. The current flowing through the resonant branch L-C, i, can be assumed as sinusoidal because the resonant branch is a sharply tuned series resonant circuit. 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 changes, the amplitude of the output current i changes so that the energy delivered to the load changes and 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–dc converter, switch S_2 and capacitor C_2 are introduced to control the energy delivered to the load resistor and 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 C_2 is partly composed of its parasitic capacitor, C_{ds}.

Assume that all the parasitic parameters are neglected, the filter inductor and capacitor are large enough so that the input current and the output voltage can be considered as pure dc

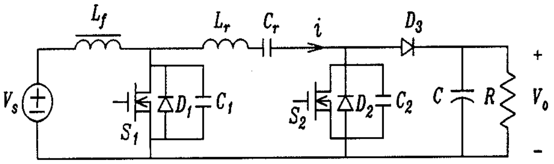


Fig. 5. Class-E dc–dc converter with half-wave controlled current rectifier.

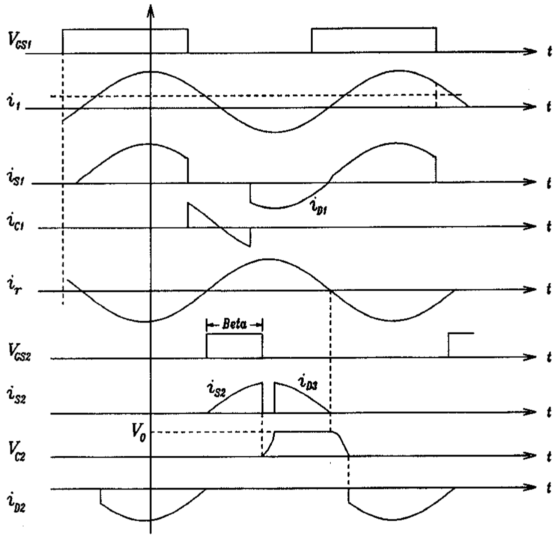
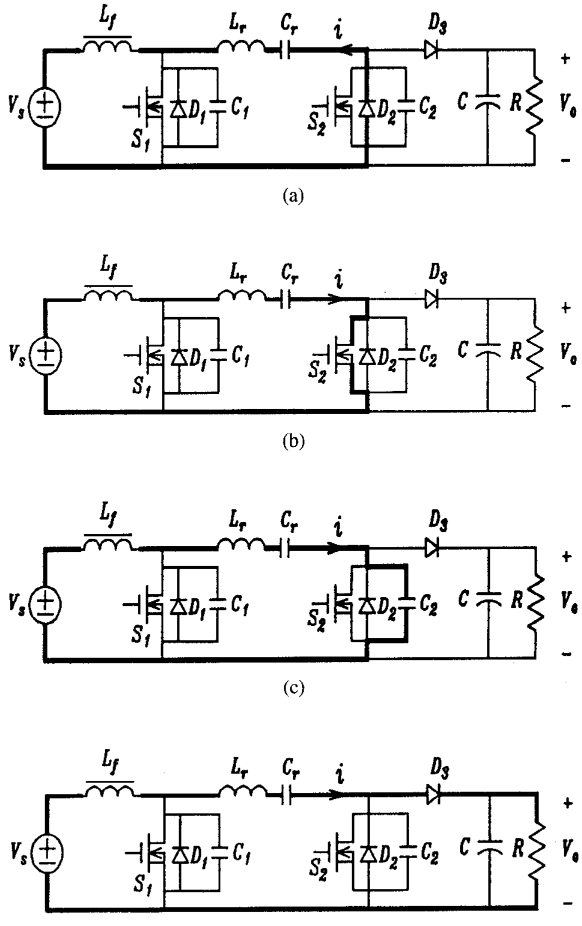


Fig. 6. Typical waveform of Fig. 5.

component and the resonant branch, L_r-C_r, is a high Q series tuned network so that the high order harmonics of the resonant current, i, is negligible. Typical waveforms are plotted in Fig. 6, where i_1 is the sum of current flowing through S_1, D_1 and C_1, i_{S1} is the current flowing through MOSFET S_1 and Beta denotes the conduction angle β. 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, as 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, S_2 conducts and the current flows through S_2, as shown in Fig. 7(b), so that 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 conducting angle, β (the control variable) defined as

$$\beta = \frac{T_{\text{on}}}{T_s} \cdot 360^\circ \tag{1}$$

where T_s is the switching period, T_{on} is the on time of switch S_2. After S_2 is turned off, 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 obtained. When $V_{C2} = V_o$, diode D_3 is forward biased and the power is delivered from the resonant tank to the load, as shown in Fig. 7(d).



(a)

(b)

(c)

(d)

Fig. 7. Path of the current flow for Class-E converter with half-wave controlled current rectifier. (a) D_2 conducts. (b) S_2 conducts. (c) C_2 charged. (d) D_3 conducts.

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_2. 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 conducting angle β is between 0 and 180°, 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 its maximum value corresponding to $\beta = 0$.

When the conducting angle β 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 conducting angle β.

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 has already been shown that zero voltage switching for the rectifier switch S_2 is maintained for entire load range. It can also be shown that zero voltage switching for the inverter switch S_1 can also be maintained from no load to full load.

It is known, [10], that for the conventional Class-E dc–dc converter, the lossless operation for S_1 can only be achieved for

$$0 < R < R_{\max} \tag{2}$$

where R is the load resistor and the value of $R_{\max}$ is dependent on the circuit variable. Zero voltage switching will be no longer present when $R > R_{\max}$. The circuit cannot operate at no load condition. These are other drawbacks of the conventional Class-E dc–dc converter.

For the Class-E converter with controlled current rectifier shown in Fig. 5, the switching condition for the inverter switch S_1 can be analyzed as follows assuming that the steady-state output voltage does not change when the load resistor changes.

When the load resistor is small, the conducting angle β should also be kept small so that the conducting angle of diode D_3 is large to provide higher output current. In this case, the equivalent load to the Class-E inverter satisfies (2). When the load resistor increases, the conducting angle of D_3 is reduced and the conducting angle for S_2 is increased. The equivalent load appeared at the output of the inverter stage will not increase, but will 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. Therefore, (2) is always satisfied and 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–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.

It is noted that in the conventional class-E converter [8], [9], the voltage stress for S_1 is dependent heavily on capacitor C_1. The larger the C_1 value, the smaller the peak voltage. However, the large the C_1, more difficult for S_1 to achieve zero voltage switching. This is the compromise we must make when selecting the component value. In the proposed class-E converter with controlled current rectifier, zero voltage switching for S_1 is guaranteed so we can use relatively large C_1 to reduce the voltage stress of S_1. The voltage stress for S_2 is the output voltage. The current stress for S_2 will be in the same order of the output current. It is expected that the current rating for S_1 will be same as that of conventional class-E converter.

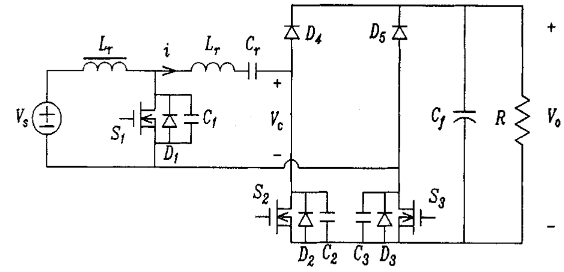


Fig. 8. Class-E dc–dc converter with full-wave controlled current rectifier.

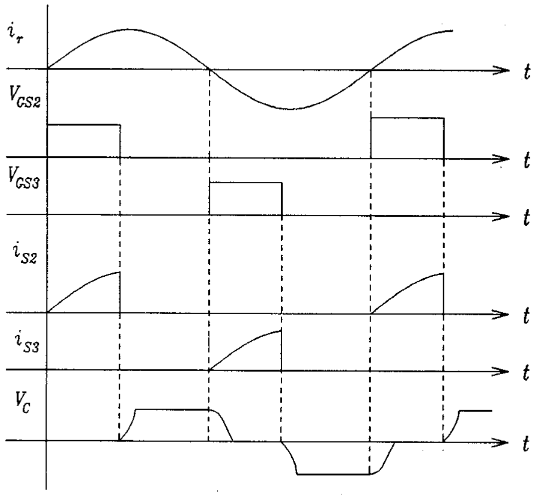


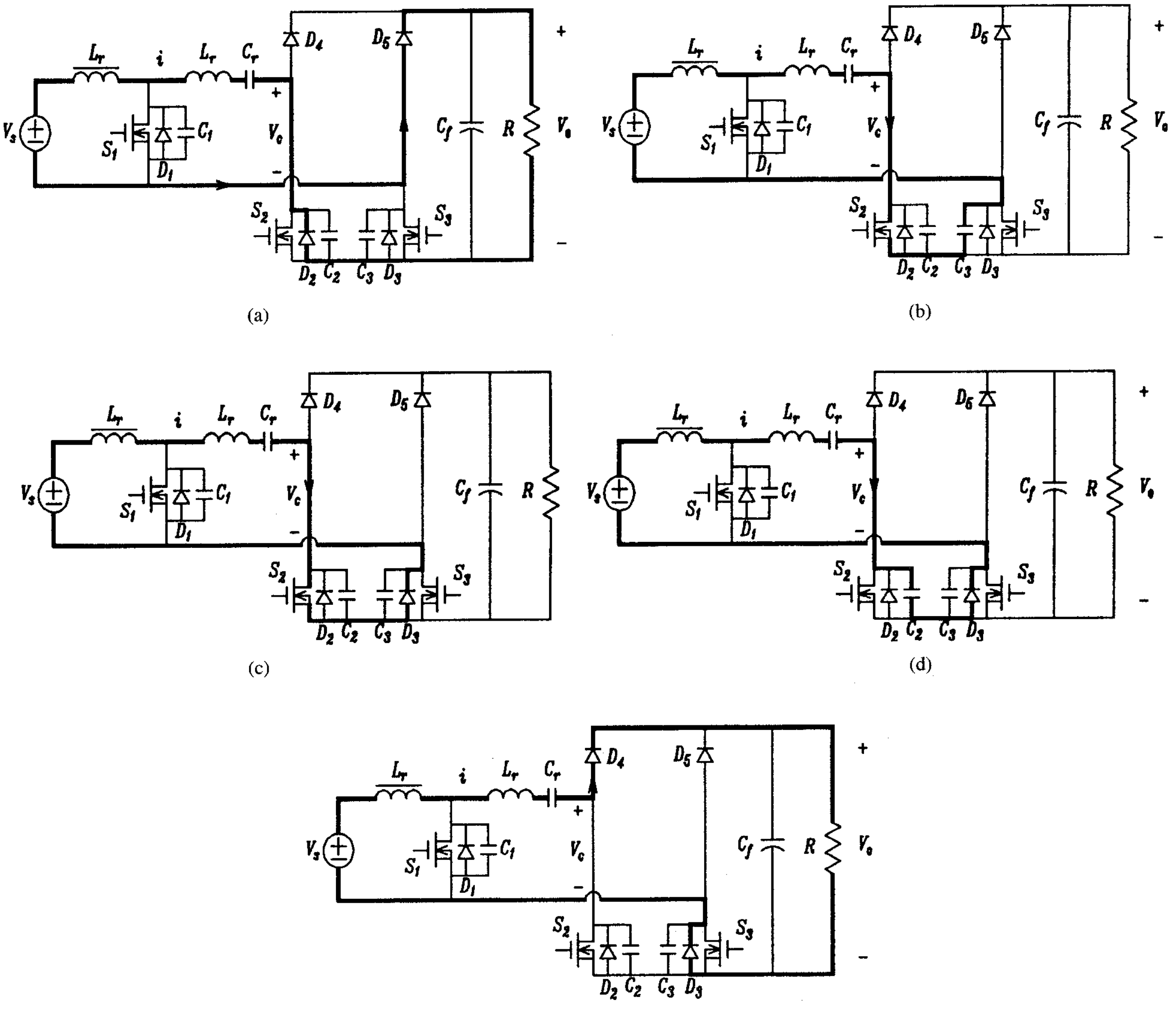
Fig. 9. Typical rectifier waveforms of Fig. 8.

It should also be noticed that at light load, the conduction angle for S_2 is large and the resonant branch sees small impedance. There is relatively high circulating current which is used to achieve zero voltage switching for S_1. The light load efficiency is thus suffered.

IV. Proposed Class-E DC–DC Converters with Full Wave and Isolated Controlled Current Rectifier

When the half-wave controlled current rectifier in Fig. 5 is replaced by a full-wave controlled current rectifier, the Class-E converter with full-wave controlled current rectifier is obtained, as shown in Fig. 8.

In Fig. 8, the inverter stage is the same as that in Fig. 5. The full-wave controlled current rectifier is composed of D_4, D_5, S_2, D_2 C_2 and S_3 D_3 C_3. The typical waveforms associated with the rectifier are shown in Fig. 9. The other waveforms are similar to those of half-wave. 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 S_3 is turned on when i changes from positive to negative. Just before the resonant current changes polarity from negative to positive, it flows through diode D_5 and D_2 and delivers the energy to the load resistor, as shown in Fig. 10(a). The current direction in the figure denotes the actual direction. The gate signal for S_2 should be applied at this time. When the resonant current rises to positive, S_2 is turned on at zero voltage. The current flows



(a)

(b)

(c)

(d)

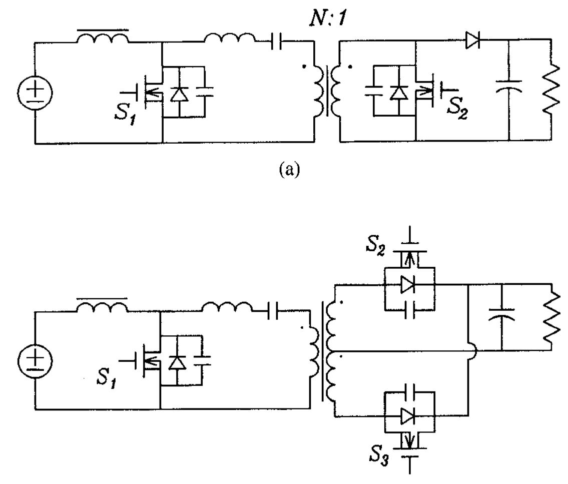
(e)

Fig. 10. Path of the current flow for Class-E converter with full-wave controlled current rectifier. (a) Through D_2 and D_5. (b) Through S_2 and C_3. (c) Through S_2 and D_3. (d) Through C_2 and D_3. (e) Through D_3 and D_4.

through S_2 and C_3 to discharge C_3, as shown in Fig. 10(b). When the capacitor C_3 is discharged completely, diode D_3 conducts, as shown in Fig. 10(c). When S_2 is turned off, the resonant current first charges C_2 and the voltage across S_2 rises slowly and zero voltage turn off for S_2 is thus achieved, as shown in Fig. 10(d). When the voltage across C_2 rises to the output voltage, diode D_4 conducts and the energy is delivered to the load through D_4 and D_3, as shown in Fig. 10(e). The operation of the next half cycle is similar and is not discussed here in detail.

From the above analysis, it is obvious that zero voltage switching for S_2 and S_3 can be maintained from no load to full load as they are 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.

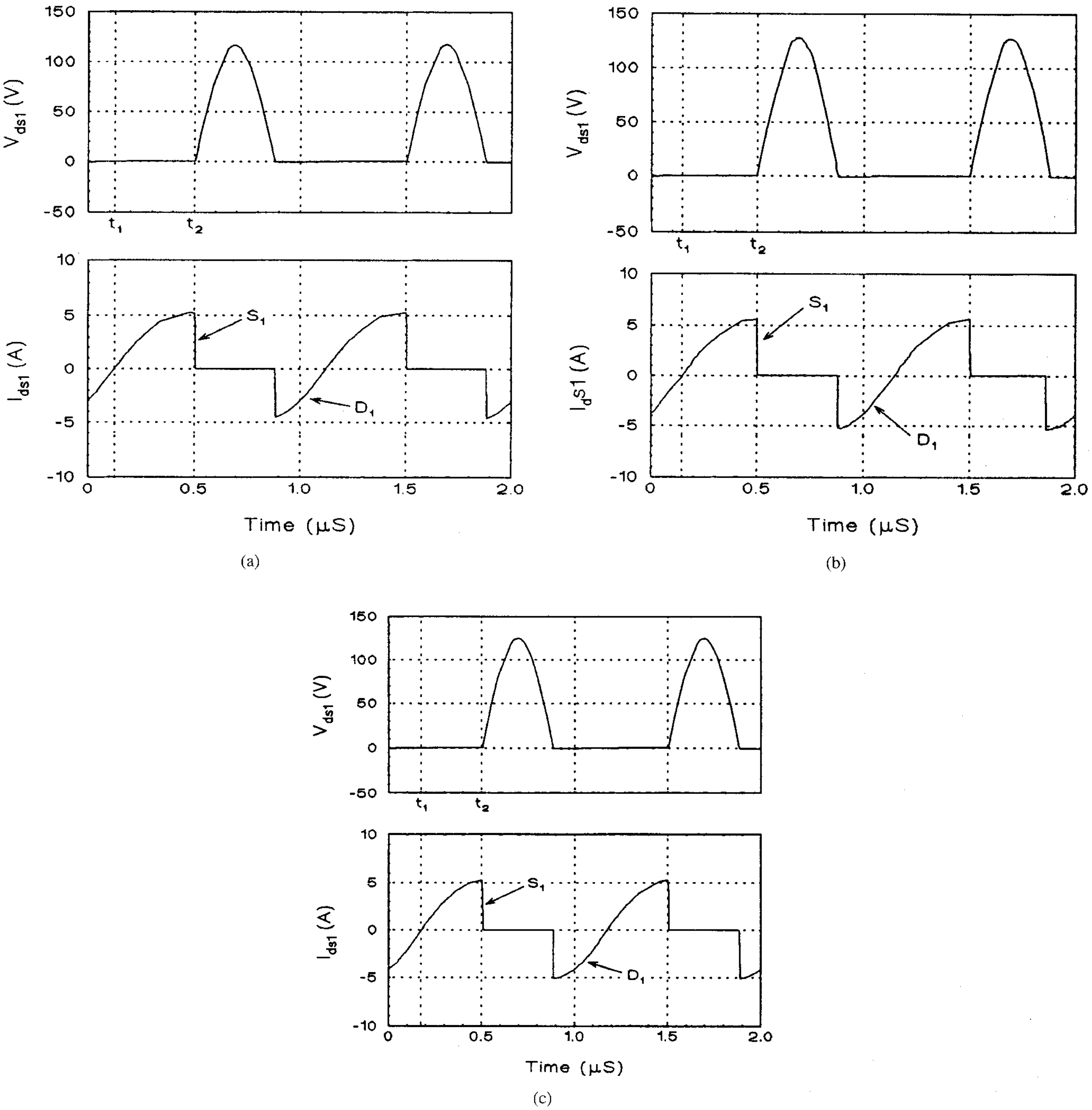
The output voltage can be regulated by changing the conducting angle of S_2 and S_3 in the similar manner as that of half-wave controlled current rectifier. For example, when S_2 and S_3 does not conduct at all, the circuit behaves like the



(a)

(b)

Fig. 11. Isolated version of Class-E dc–dc converters with controlled current retifier. (a) Half-wave. (b) Full-wave.



(a)

(b)

(c)

Fig. 12. Switching waveforms of S_1 at different conduction angle β. (a) $\beta = 0°$. (b) $\beta = 90°$. (c) $\beta = 144°$.

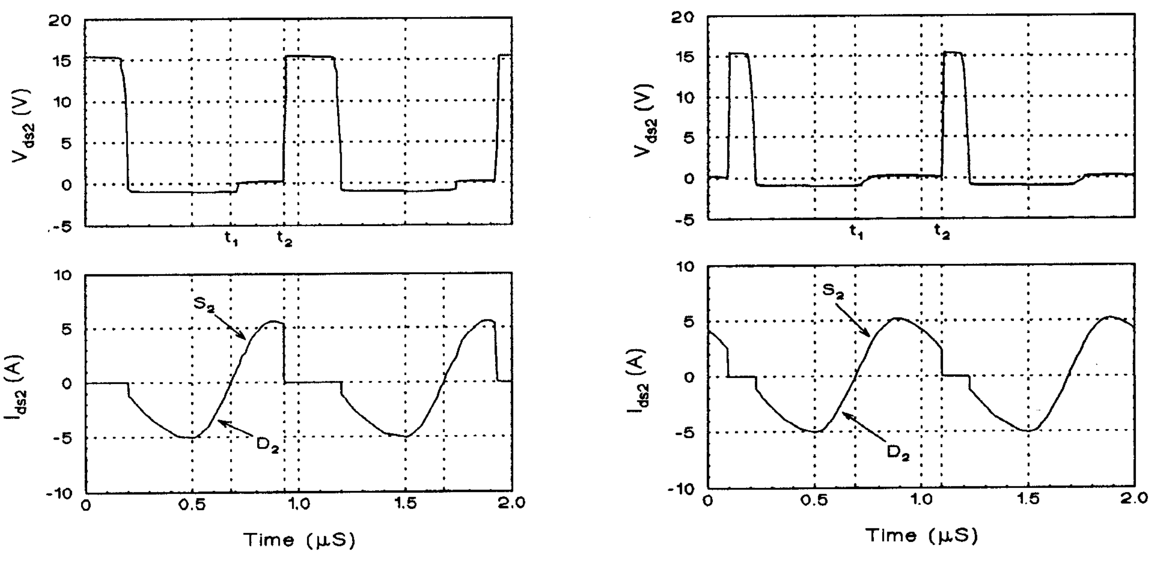
conventional Class-E converter with full-wave rectifier and the output voltage is high. When S_2 and S_3 conduct as long as they can, the output voltage is zero.

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. 11. Fig. 11(a) is the half-wave version and Fig. 11(b) is the full-wave one. Their operating principle is similar to that of nonisolated ones and not explained here. One thing that should be noticed here is that in the full-wave version, as shown in Fig. 11(b), there is no dc component in the transformer so that its implementation is simplified.

V. COMPUTER SIMULATION AND EXPERIMENTAL RESULTS

The operation of the Class-E dc–dc converter with half-wave and full-wave controlled current rectifier (Fig. 5 and Fig. 8) is simulated by PSPICE to show the feasibility of the proposed circuit. The circuit parameters used in the simulation are $L_r = 12$ μH, $C_r = 2.3$ μF, $C_1 = 2$ nF, $C_2 = 1$ nF, $C_3 = 1$ nF (for full-wave controlled rectifier) and $L_f = 100$ μH. The switching frequency is selected as 1 MHz. The supply voltage $V_s = 30$ V and the output is modeled by a constant voltage source with $V_o = 15$ V.

Figs. 12 and 13 give the switching waveforms of S_1 and S_2 for the Class-E converter with half-wave controlled current



(a)

(b)

Fig. 13. Switching waveforms of S_2 at different conduction angle β.

rectifier. Fig. 12 gives the voltage V_{ds1} and current I_{ds1} associated with the inverter switch S_1 at different conducting angle, β, (a) $\beta = 0$, i.e., S_2 does not conduct at all, which is equivalent to the conventional Class-E converter, (b) $\beta = 90°$ when S_2 conducts for half of the positive cycle of i and (c) $\beta = 144°$ when the output current is very small. It is clear that the zero voltage switching for S_1 can be maintained from full load to no load. Fig. 13 gives the switching waveforms of the rectifier switch S_2 when (a) $\beta = 90°$ and (b) $\beta = 144°$, respectively. Obviously, zero voltage switching is achieved.

For the Class-E dc–dc converter with full-wave controlled rectifier (Fig. 8), zero voltage switching for all the switches, S_1, S_2, S_3, can also be maintained for different β and the simulation results are not presented here.

By controlling the conducting angle, β, the output current can be regulated at constant switching frequency. Fig. 14 gives the simulated output current I_o versus the conducting angle β for the Class-E dc–dc converter with half-wave and full-wave controlled current rectifier. It can be observed that: 1) the output current can be effectively controlled by changing the conducting angle β to keep the output voltage constant, and 2) the current gain of the full-wave version is about twice as high as that of the half-wave version for the same circuit parameters.

Another phenomenon that can be observed is that given the parameters used in the parameters, when β increases from zero, the current gain also increases a little bit and then reduces. The reason for this is that the output of the class-E inverter is not an ideal current source. Its amplitude actually increases a little bit first when β increases from zero and then reduces. The minimum β is determined by the impedance Z_r of the resonant tank (L_r and C_r). The large the Z_r compared with the output resistance, the closer to an ideal current source for the output of the class-E inverter and, therefore, the smaller the minimum β. In the actual circuit, the minimum β should be limited to obtain one-to-one control to output characteristics.

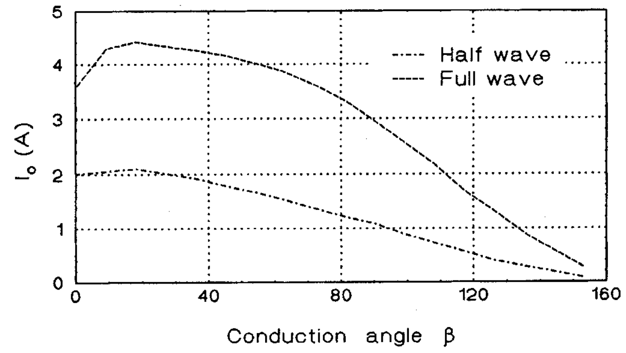
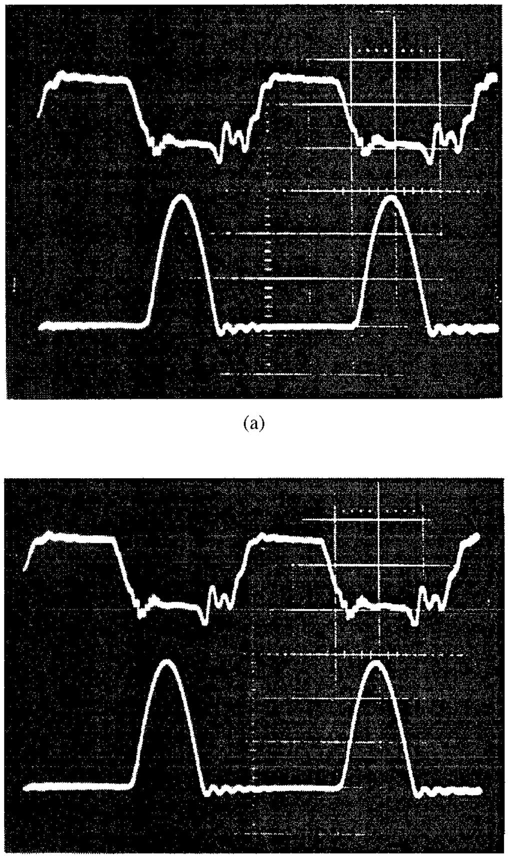


Fig. 14. Simulated output current for half-wave and full-wave controlled current rectifier.

An experimental prototype of class-E dc–dc converter with half-wave controlled current rectifier is also breadboarded in order to show the feasibility of the proposed circuit. The experimental circuit is the same as that shown in Fig. 5. The input voltage is 30 V ad the output voltage is regulated at 15 V. The load resistor changes from 10 Ω to open circuit.

Fig. 15 gives the gate drive signal and device voltage of the inverter switch S_1 at different conduction angle β. Fig. 15(a) gives the waveforms when $\beta = 60°$ and the output current is measured as 0.9 A. Fig. 15(b) is the waveforms when $\beta = 144°$ and the output current is at 0 A. 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.

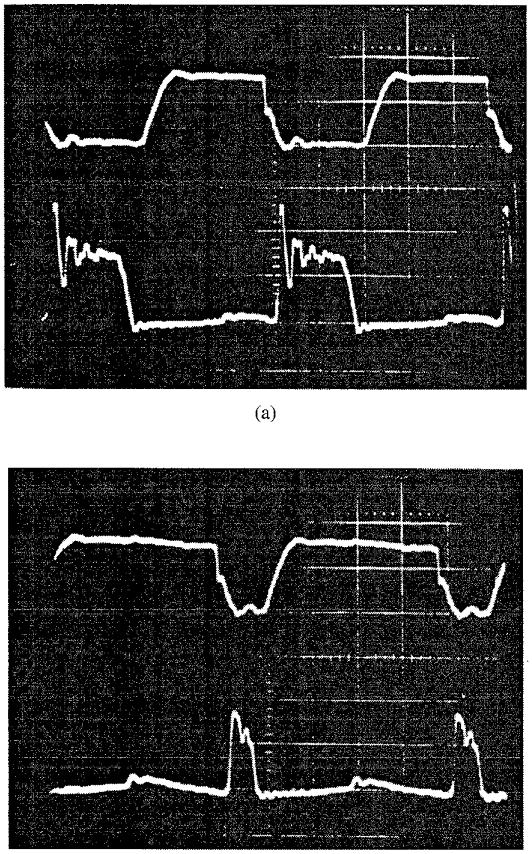
Fig. 16 gives the gate signal and the device voltage of the rectifier switch S_2 for different conduction angle β. Fig. 16(a) and (b) gives the waveform when $\beta = 60°$ and $\beta = 144°$,



(a)

(b)

Fig. 15. Switching waveforms of the inverter switch S_1 at (a) $\beta = 60°$ and (b) $\beta = 144°$. Upper trace: V_{GS1}, 10 V/div, lower trace: V_{DS1}, 50 V/div, Horizontal: 0.2 μS/div. (a) $\beta = 60°$. (b) $\beta = 144°$.



(a)

(b)

Fig. 16. Switching waveforms of the rectifier switch S_2 at (a) $\beta = 60°$ and (b) $3 = 144°$. Upper trace: V_{GS2}, 10 V/div, lower trace: V_{DS2}, 10 V/div, Horizontal: 0.2 μS/div. (a) $\beta = 60°$. (b) $\beta = 144°$.

respectively. When $\beta = 144°$, 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.

VI. CONCLUSION

In this paper, the present techniques of the Class-E dc–dc converters are reviewed and their drawbacks are addressed. New Class-E dc–dc converter topologies with half-wave and full-wave controlled current rectifier are proposed. The mechanism to maintain the zero voltage switching for all the switches is explained. The output voltage can be regulated by changing the conducting angle of the rectifier switch. The salient advantages of the new converter topologies are:

1) The switching frequency is constant.
2) Zero voltage switching for all the switches can be maintained from no load to full load.
3) The circuits can operate at no load condition.

The operation of the Class-E dc–dc converters with half-wave and full-wave controlled current rectifier is simulated by PSPICE and an experimental prototype is also built to show the feasibility of the proposed circuits. Zero voltage switching is shown clearly in the simulation and experimental waveforms. The output current versus the control signal, i.e., the conducting angle β, is also obtained through the simulation and it shows that output voltage can be kept at constant when the load current changes by regulating the conducting angle β. The switching frequency is not changed.

REFERENCES

[1] P. C. Sen, *Thyristor DC Drives*. New York: Wiley, 1981.

[2] ——, *Principles of Electric Machines and Power Electronics*. New York: Wiley, 1988.

[3] N. Mohan, T. M. Undeland, and W. P. Robbins, *Power Electronics, Converter, Applications and Designs*. New York: Wiley, 1989.

[4] M. H. Rashid, *Power Electronics, Circuits Devices and Applications*, Englewood Cliffs, NJ: Prentice-Hall, 1988.

[5] R. L. Steigerwald, "High-frequency resonant transistor DC-DC converters," *IEEE Trans. Ind. Electron.*, vol. 31, pp. 181–191, May, 1984.

[6] V. T. Ranganathan, P. D. Ziogas and V. R. Stefanovic, "A regulated DC-DC voltage source converter using a high-frequency link," *IEEE Trans. Ind. Applicat.* vol. 18, pp. 279–287, May/June 1982.

[7] R. L Steigerwald, "A comparison of half-bridge resonant converter topologies," *IEEE Trans. Power Electron.*, vol. 3, pp. 174–182, Apr. 1988.

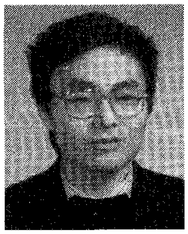
[8] R. Redl, B. Molnar, and N. O. Sokal, "Class-E resonant DC/DC power converters: analysis of operations, and experimental results at 1.5MHz," *IEEE Trans. Power Electron.*, vol. 1, pp. 111–119, Apr. 1986.

[9] G. Lütteke and H. C. Raets, "220-V mains 500-KHz Class-E converter using a BIMOS," *IEEE Trans. Power Electron.*, vol. 2, pp. 186–193, July 1987.

[10] M. K. Kazimierczuk and X. T. Bui, "Class-E DC/DC converters with an inductive impedance inverter," *IEEE Trans. Power Electron.*, vol. 4, pp. 124–135, Jan. 1989.

[11] W. J. Gu and K. Harada, "A new method to regulate resonant converters," *IEEE Trans. Power Electron.*, vol. 3, pp. 430–439, 1988.

[12] C. Q. Hu, X. Z. Zhang, and S. P. Huang, "Class-E combined converter by phase-shift control," in *IEEE PESC 1989*, pp. 229–234.



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