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Development and Optimization of a High-Current ZVS-PWM Converter for Industrial Application

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Abstract:

This paper introduces a novel auxiliary circuit designed for implementation in both DC-DC and AC-DC Zero Voltage Switching Pulse Width Modulation (ZVS-PWM) converters. The proposed auxiliary circuit enhances the performance of ZVS-PWM converters in applications requiring high-frequency operation and the capability to handle higher load currents than conventional ZVS-PWM converters.

The operation of the new ZVS-PWM converter is thoroughly described, providing insights into its working principles and the role of the auxiliary circuit in achieving ZVS across a wide range of operating conditions. A comprehensive steady-state analysis is performed to understand the circuit behavior, including voltage and current characteristics, switching loss reduction, and efficiency improvements.

Additionally, a systematic design procedure is developed to guide the implementation of the proposed converter. The design methodology is demonstrated through a detailed example, highlighting key considerations such as component selection, circuit optimization, and thermal management. To validate the proposed design, experimental results from a prototype converter are presented, showcasing its feasibility, efficiency, and performance under real-world operating conditions.

The findings of this research contribute to the advancement of high-performance ZVS-PWM converters, offering a practical solution for applications demanding high efficiency, high frequency, and the ability to handle elevated load currents.

Keywords: DC-DC converter, zero voltage switching, auxiliary circuit, pulse width modulation, high-frequency operation

1. Introduction

Numerous techniques employing active auxiliary circuits have been developed to assist the main switch of single-switch pulse-width modulated (PWM) converters in achieving zero-voltage switching (ZVS) [1–27]. These techniques are instrumental in reducing switching losses in the main power switch, minimizing reverse-recovery losses in the main power diode, and mitigating electromagnetic interference (EMI) within the converter. Typically, the auxiliary circuit is connected in parallel with the main switch, as depicted in Figure 1, and is activated shortly before the main converter switch is turned on. When activated, the auxiliary circuit progressively redirects current away from the main power diode, thereby eliminating the reverse recovery current. It subsequently discharges the capacitance across the main switch, enabling it to turn on under ZVS conditions. Once the main switch is successfully turned on, the auxiliary circuit is promptly deactivated, allowing the converter to operate as a conventional PWM converter for the remainder of the switching cycle. An advantage of this approach is that the auxiliary circuit components are designed with lower power ratings than those of the main power circuit. This is feasible because the auxiliary circuit is active only for a brief duration during each switching cycle. Furthermore, the auxiliary switch can be selected as a device with inherently lower switching losses compared to the main switch, further enhancing the overall efficiency and performance of the converter. Previously proposed ZVS-PWM converters suffer from several drawbacks that limit their efficiency and practicality. One common issue is that the auxiliary switch is often turned off while it is conducting current, resulting in switching losses and electromagnetic interference (EMI) that undermine the benefits of the auxiliary circuit. Additionally, many designs cause the main switch or boost diode to experience higher peak current stress and increased circulating current, leading to greater conduction losses and necessitating the use of higher current-rated components, which increases costs and reduces overall efficiency. Another significant drawback is the high peak voltage and current stresses imposed on auxiliary circuit components, with peak voltage stresses often exceeding twice the output voltage. This increases the demands on the auxiliary components, affecting reliability and scalability. Furthermore, some converters require energy from the output to be diverted into the auxiliary circuit to initiate a resonant process, contributing to additional circulating current and

energy losses. These challenges underscore the need for improved auxiliary circuit designs that address these limitations while maintaining the benefits of ZVS-PWM operation.

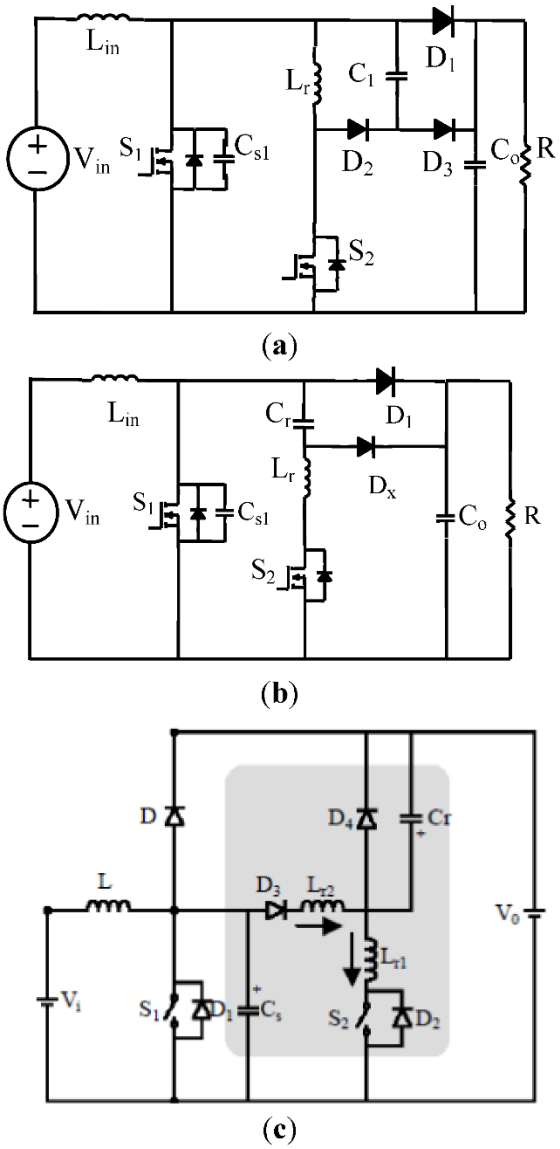

Figure 1. Zero-voltage switching (ZVS)-pulse-width modulated (PWM) boost converters with auxiliary circuits. (a) Non-resonant auxiliary circuit [1]; (b) Resonant auxiliary circuit [11]; (c) Dual auxiliary circuit [16].

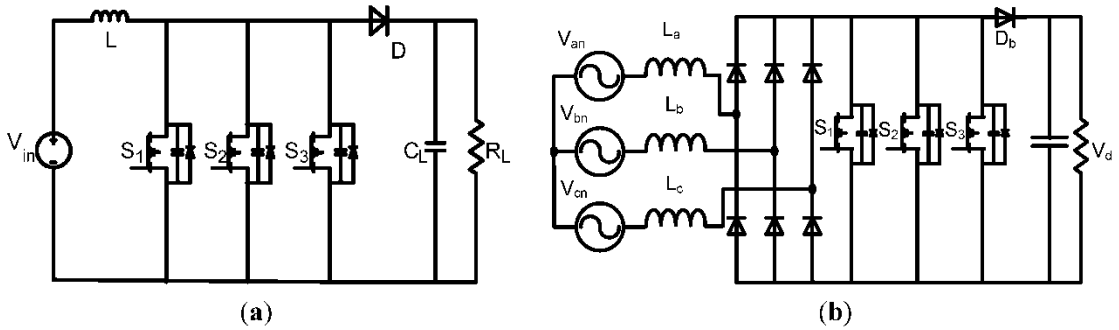

Figure 2. Converters with paralleled MOSFETs. (**a**) DC-DC boost converter; (**b**) Three-phase boost rectifier.

In industrial applications, it is standard practice to implement PWM converters with multiple MOSFETs connected in parallel to reduce the on-state resistance of the main power switch, thereby minimizing conduction losses. Examples of such configurations include the DC-DC PWM boost converter and the three-phase AC-DC converter, as illustrated in

Figures 2a and 2b, respectively. While a single IGBT can serve as the boost switch and may be more cost-effective, it cannot operate at switching frequencies as high as those achievable with MOSFETs. This limitation restricts the reduction of the size of magnetic and filtering components, which is critical for industrial applications like telecom power systems where compactness is essential due to space constraints in cabinets.

In converters designed for higher power and current, auxiliary circuits are often employed to reduce switching losses in paralleled MOSFET configurations. However, the drawbacks associated with such auxiliary circuits become more pronounced at higher power levels. For instance, in a non-resonant auxiliary circuit (as shown in Figure 1a), the turn-off losses of the auxiliary switch can be substantial. To address this, resonant approaches (Figure 1b and dual Figure 1c) may be employed to ensure the auxiliary switch turns off softly. However, these approaches introduce new challenges. The auxiliary inductor current waveform for the Figure 1b circuit generates a negative current that circulates through the main switches, increasing their peak stresses and conduction losses. Meanwhile, in the dual-resonant converter of Figure 1c, the current waveform of the auxiliary inductor (ILr1) exhibits extremely high peaks, often exceeding double the input current. This imposes stringent requirements on the auxiliary switch, making it challenging to select a suitable device capable of handling such high peak currents. These issues complicate the design and operation of auxiliary circuits in high-power converters, necessitating further refinement to address these challenges effectively.

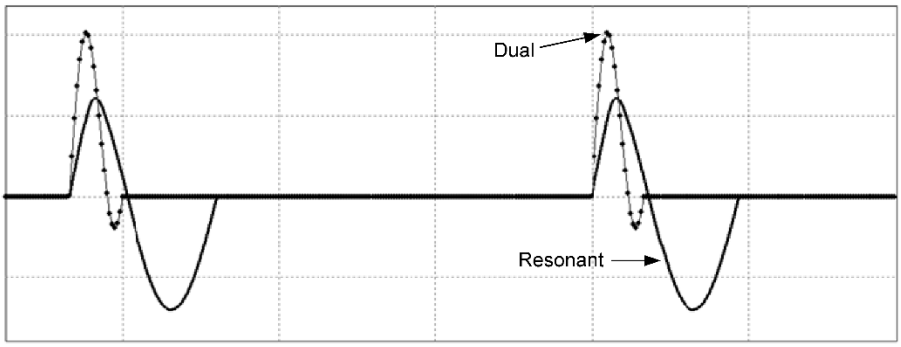

Figure 3. Typical auxiliary inductor current waveforms for ZVS-PWM boost converters operating with input voltage V_{in} = 100 V, output voltage V_o = 400 V, output power P_o = 2 kW, and switching frequency f_{sw} = 100 kHz. Scale: I = 20A/div., Time: t = 5 μs/div.

A new auxiliary circuit for ZVS-PWM converters that are implemented with paralleled MOSFETs for higher current applications is proposed in the paper. The circuit is shown in Figure 4. Although almost all previously proposed auxiliary circuits contain only a single active switch because of cost (it is difficult to justify a two-switch circuit in a converter with a single MOSFET as the power switch), the proposed auxiliary circuit can be justified on the following grounds:

Its performance is superior to all other single-switch auxiliary circuits for higher current applications because its switches can be turned off softly and it can operate with greater flexibility than single-switch resonant and dual auxiliary circuits. Resonant and dual auxiliary circuits have issues related to the timing of the operation of the auxiliary switch relative to that of the main power switch(es) as the time window of opportunity to turn the auxiliary switch softly varies considerably from light load to heavy load. In other words, ZVS-PWM converters with single-switch auxiliary circuits like the ones shown in Figure 1 are not suitable for higher current applications and should not be used for these applications.

Cost is less of an issue and performance is the key criterion in applications where multiple MOSFETs are used. If the cost of multiple MOSFETs to improve performance can be justified for the power switch, then it can be justified in the auxiliary circuit.

Two-switch auxiliary circuits for ZCS-PWM IGBT converters are commonly used in high current applications and there is a vast literature about them [19]. Most multi-switch auxiliary circuits for ZCS-PWM converters that have been proposed have been for three-phase buck-type rectifiers and three-phase current source inverters. In the case of a three-phase rectifier, as shown in Figure 5a, the auxiliary circuit can be placed either across the dc link inductor (Position A) or across the output of the bridge (Position B). Several multi-switch auxiliary circuits are shown in Figure 5b,c. Given that multi-switch auxiliary circuits are widely used in higher power ZCS-PWM applications to improve performance, the use of such circuits in higher power ZVS-PWM applications where paralleled MOSFETs are used can be justified for the same reason.

2. Modes of Operation

The proposed converter in Figure 4 has an auxiliary circuit that consists of two switches, S_{aux1} and S_{aux2}, three diodes, and a resonant tank made of capacitor C_r and inductor L_r. The basic operating principles of the proposed circuit are as follows:

Auxiliary switch S_{aux1} is turned on just before the main power switch S is to be turned on, thus diverting current away from the main power diode D. Once current has been completely diverted away from D, the output capacitances of the switch begin to discharge and the voltage across it eventually falls to zero. The main power switch can be turned on with ZVS as soon as the capacitance is fully discharged. Due to the C_r-L_r resonant tank, the current in the auxiliary circuit naturally falls to zero, thus allowing S_{aux1} to turn off with ZCS.

Sometime during the switching cycle, while the main power switch is conducting the input current, auxiliary switch S_{aux2} is turned on. This action results in the voltage across C_r flipping polarity so that it is negative instead of positive. When the main power switch is turned off, the input current completely discharges C_r so that there is no voltage across it when the auxiliary circuit is reactivated sometime during the next switching cycle. Equivalent circuit diagrams of the modes of operation that the converter goes through during a switching cycle are shown in Figure 6, and typical converter waveforms are shown in Figure 7. To save on space, switches S_1, S_2, and S_3 are shown in Figure 6 as a single switch, S_{123}.

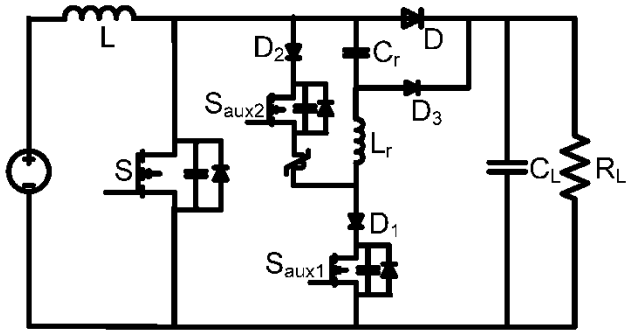

Figure 4. Proposed DC-DC boost converter.

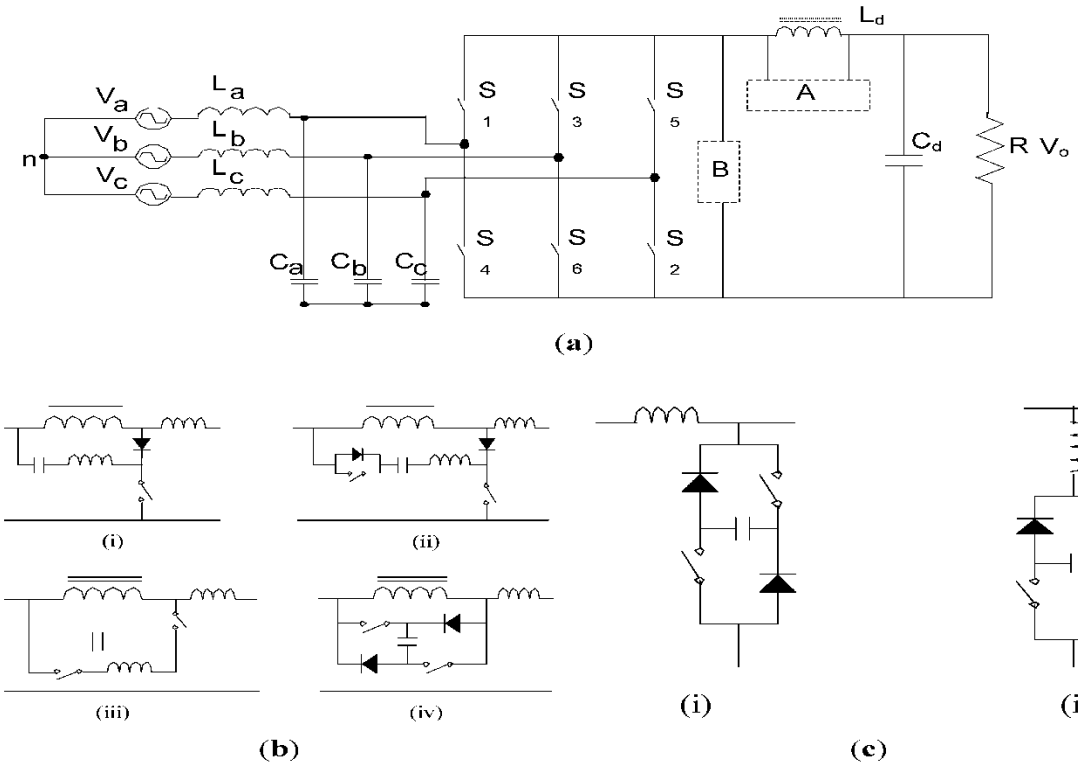

Figure 5. (**a**) Three-phase six-switch rectifier. (**b**) Various auxiliary circuit schemes for position A. (**c**) Various auxiliary circuit schemes for position B.

The converter's modes of operation are as follows:

Mode 0 ($t < t_0$): All converter switches are off during this mode and current is flowing through the main power diode D.

Mode 1 ($t_0 < t < t_1$): At $t = t_0$, switch S_{aux1} is turned on and current begins to be transferred away from diode D to the auxiliary circuit. This current transfer is gradual due to the presence of inductor L_r in the auxiliary circuit, so that charge

is removed at a sufficiently slow rate to allow diode D to recover; this helps minimize reverse recovery current. The equations that represent the auxiliary circuit inductor current I_{Lr} and the auxiliary circuit capacitor voltage V_{Cr} in this mode are:

$$I_{L_r}(t) = \frac{V_o}{Z_1}\sin(\omega_1(t-t_0))$$

$$V_{C_r}(t) = V_o - V_o\cos(\omega_1(t-t_0))$$

$$Z_1 = \sqrt{\frac{L_r}{C_r}}$$

$$\omega_1 = \frac{1}{\sqrt{L_r C_r}}$$

and the initial values of I_{Lr} and V_{cr} at the beginning of this mode are zero.

It should be noted that current can flow through the output capacitor of S_{aux2} after S_{aux1} is turned on. In order to minimize a sudden increase in current through this capacitor that can cause voltage spikes to appear, a saturable reactor or "spike-killer" inductor (L_s) should be placed in series with S_{aux2}.

Mode 2 ($t_1 < t < t_2$): At $t = t_1$, current stops flowing through the main power diode D and the net capacitance across S_{123} begins to be discharged through L_r and C_r. The current in the auxiliary circuit is the sum of the input current and the current due to the discharging of the capacitances across S_{123}. The equations that describe the auxiliary circuit inductor current I_{Lr}, the voltage across S_{123}, V_{Cs}, and the auxiliary circuit capacitor voltage V_{Cr} in this mode are

$$I_{Lr}(t) = \frac{V_1}{Z_2}\sin(\omega_2(t-t_1)) +$$

$$\mathrm{I}_1\cos(\omega_2(t-t_1)) - I_1 + I_{L_r}(t_1)$$

$$V_{Cs}(t) = \frac{C}{C_s}[V_1\cos(\omega_2(t-t_1)) - V_1 + V_o +$$

$$I_1 Z_2\sin(\omega_2(t-t_1))] + \frac{I_{Lm}}{C_r + C_s}(t-t_1)$$

$$V_{Cr}(t) = -\frac{C}{C_s}[V_1\cos(\omega_2(t-t_1)) + V_{Cr}(t_1)$$

$$-I_1 Z_2\sin(\omega_2(t-t_1)) - V_1] + \frac{I_{Lm}}{C_r + C_s}(t-t_1)$$

where

$$I_1 = I_{Lr}(t_1) - \frac{C}{C_s}I_{Lm}$$

$$V_1 = V_o - V_{Cr}(t_1)$$

$$C = \frac{C_r C_s}{C_s + C_r}$$

$$Z_2 = \sqrt{\frac{L_r}{C}}$$

$$\omega_2 = \sqrt{\frac{1}{L_r C}}$$

During this mode, the auxiliary circuit inductor current I_{Lr}, reaches its peak when $V_{Cs} - V_{Cr} = 0$ and it is equal to the peak current of S_{aux1} so that

$$I_{Lr,P} = \frac{\sqrt{V_1^2 + (I_1 Z_2)^2}}{Z_2} + \frac{C}{C_s}I_{Lm}$$

Mode 3 ($t_2 < t < t_3$): At $t = t_2$, the capacitance across the main power switches is completely discharged and current begins to flow through the body diodes of the devices; this allows the switches to be turned on with ZVS. The equations that describe the auxiliary circuit inductor current I_{Lr} and the auxiliary circuit capacitor voltage V_{Cr} in this mode are:

$$I_{L_r}(t) = I_2\cos(\omega_1(t-t_2)) - \frac{V_2}{Z_1}\sin(\omega_1(t-t_2))$$

$$V_{C_r}(t) = I_2 Z_1\sin(\omega_1(t-t_2)) + V_2\cos(\omega_1(t-t_2))$$

$$I_2 = \sqrt{\frac{L_r I_{L,P}^2 - C_r V_2^2}{L_r}}$$

$$V_2 = \frac{I_{Lm}}{C_r}(t_2 - t_1) + \frac{C_s}{C_r} V_o + V_{cr}(t_1)$$

Mode 4 ($t_3 < t < t_4$): At $t = t_3$, the current that was flowing in the body diodes of the main power switches in the previous mode reverses direction and begins to flow through the switches. The modal equations of this mode are the same as those of the previous mode except that the direction of the current through the main power switches is different.

Mode 5 ($t_4 < t < t_5$): Current stops flowing in the auxiliary circuit at $t = t_4$ due to the resonant interaction between L_r and C_r. Switch S_{aux1} can be turned off softly with zero-current switching (ZCS) sometime soon afterwards. The converter then operates like a standard PWM boost converter. The voltage across C_r remains fixed until S_{aux2} is turned on later in the switching cycle.

Mode 6 ($t_5 < t < t_6$): At $t = t_5$, auxiliary switch S_{aux2} is turned on, sometime before the main power switches are turned off. As a result, capacitor C_r begins to discharge through L_r, S_{aux2} and D_2, and the voltage that was across it at the start of the mode changes polarity. At the end of this mode, the current in C_r and L_r is zero so that S_{aux2} can be turned off with ZCS. The equations that define this mode are:

3. Steady-State Characteristics

The modal equations that are derived in the previous section of the paper can be used to generate steady-state characteristic curves that can be used to see the effect of certain key parameters on the operation of the auxiliary circuit. These key parameters include the values of auxiliary circuit components L_r and C_r and the net capacitance across the main power switches, C_S. Examples of such graphs are shown in Figure 8. Each graph has been generated by keeping certain parameters constant, then varying other parameters to see the effect of doing so.

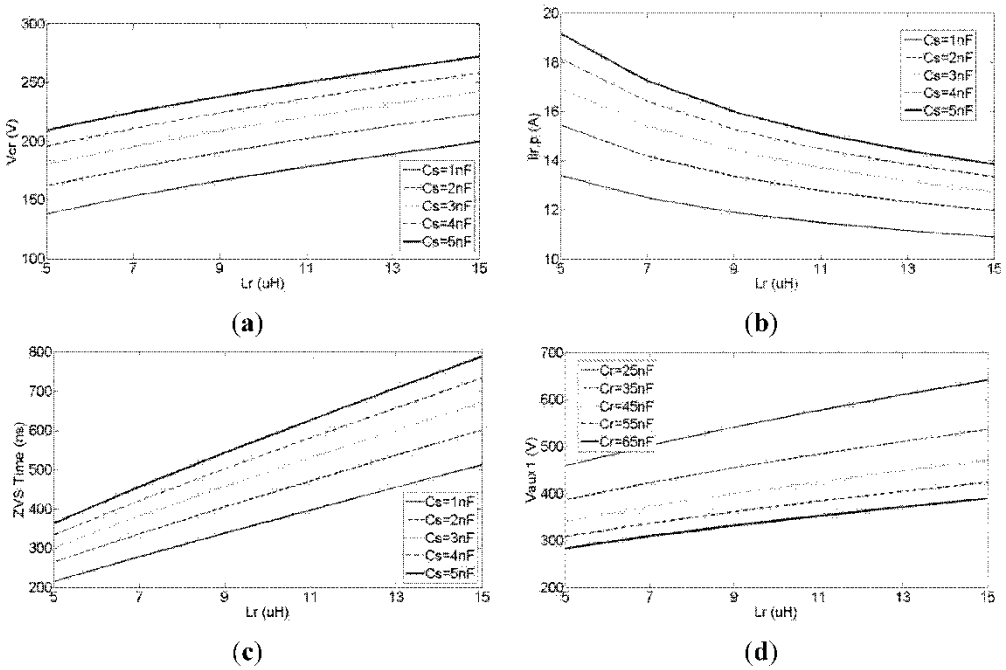

Figure 8. Characteristic curves. (**a**) Graph of characteristic curves of V_{cr} *vs.* L_r for different values of C_S with C_r = 50 nF; (**b**) Graph of characteristic curves of I_{Lr} *vs.* L_r for different values of C_S with C_r = 50 nF; (**c**) Graph of characteristic curves of I_{Lr} *vs.* L_r for different values of C_S with C_r = 50 nF; (**d**) Graph of characteristic curves of V_{aux1} *vs.* L_r for different values of C_r with C_S = 2 nF.

Figure 8a is a graph of V_{cr} *vs.* L_r for different values of C_S with C_r = 50 nF. This graph shows that V_{cr} increases as either C_S or L_r is increased. The first characteristic can be explained by noting that increasing C_S increases the amount of energy that is discharged into the auxiliary circuit and is stored in C_r after the main power diode stops conducting. More energy in C_r results in higher values of V_{cr}. On the other hand, according to (4) and (12), higher values of L_r increase the time duration between t_0 and t_2; therefore, more energy is transferred to C_r, which leads to higher values of V_{cr}.

Figure 8b is a graph of characteristic curves of I_{Lr} *vs.* L_r for different values of C_S with C_r = 50 nF. This graph shows that when C_S increases, more energy is stored in C_r, which results in higher peak values for I_{Lr}. Moreover, when L_r increases,

it extends the resonant cycle and reduces the peak value of I_{Lr}. The average value of the resonant current is related to C_S and load current and is independent of length of the resonant cycle and the peak of the resonant current.

Figure 8c shows a graph of characteristic curves of ZVS time values *vs.* L_r for different values of C_S with C_r = 50 nF. These time values are when the net capacitance across the main power switches is completely discharged after S_{aux1} is turned on and is measured from the turn-on instant of this switch. The graph shows that the ZVS times increase as L_r or C_r increases. Increasing L_r increases the time needed for current to be transferred away from the main power diode and it also increases the resonant cycle of the auxiliary circuit. On the other hand, by increasing C_S, the amount of stored energy in this capacitor increases and, therefore, it takes more time for it to be discharged.

Figure 8d shows a graph of characteristic curves of V_{aux1} *vs.* L_r for different values of C_r when C_S = 2 nF. It can be seen that increasing L_r increases the maximum voltage across S_{aux1}. Before D_1 goes off after Mode 7 and V_{aux1} becomes constant, V_{aux1} is

$$V_{aux1} = V_{cr} + L_r \frac{di_{Lr}}{dt}$$

Equation shows that V_{aux1} is increased by increasing L_r. Also, for the same amount of energy transferred to C_r, increasing C_r reduces the voltage across it and thus reduces V_{aux1} as well.

4. Design Procedure and Example

Steady-state characteristic curves, such as those illustrated in Figure 8, serve as a foundation for developing a systematic procedure to select key component values for the proposed ZVS-PWM converter. This section outlines the design procedure, focusing on the auxiliary circuit, while excluding the design of the main boost power circuit. The main boost power circuit design follows established principles of standard PWM boost converters, including the determination of the type and number of main power switches. Additionally, it is assumed that the combined output capacitance of the paralleled main power devices is sufficient to control the rise in voltage across them after turn-off, negating the need for extra external capacitance.

4.1. Selection of Auxiliary Circuit Inductor (Lr)

The auxiliary circuit inductor, Lr, plays a crucial role in ensuring smooth operation of the ZVS-PWM converter. Its minimum value is determined by its capability to limit the reverse recovery current of the main power boost diode. Reverse recovery current occurs when the diode transitions from conducting to blocking mode, which can lead to significant switching losses and electromagnetic interference (EMI) if not properly managed.

To mitigate these effects, the current transition from the main diode to the auxiliary circuit must be gradual. The rate of this transition is directly influenced by the inductance value of Lr. A larger inductance value slows down the current transfer, effectively reducing the peak reverse recovery current of the diode. This reduction minimizes switching losses and helps prevent potential damage to the diode caused by high peak currents.

Steps for Selecting Lr

1. **Establish Design Constraints:** Begin by identifying the diode's reverse recovery characteristics, such as its reverse recovery time (trr) and reverse recovery current (Irr). These parameters are typically provided in the diode's datasheet.
2. **Define Acceptable Current Transition Rate:** Determine the desired rate of current transition (di/dt) away from the diode to the auxiliary circuit. A lower di/dt value reduces reverse recovery current but increases the size of Lr, which must be balanced against physical and cost constraints.
3. **Validate Performance through Simulation:** After determining the minimum value of Lr, simulate the circuit operation to verify that the selected inductance sufficiently reduces reverse recovery current and does not introduce excessive delays in the switching operation.

Impact of Lr on System Performance

While increasing Lr effectively reduces reverse recovery current, it also has implications for the overall converter performance. A larger inductor size can increase the time required for the current to transfer completely to the auxiliary circuit, potentially affecting the efficiency of ZVS operation. Therefore, the selected value of Lr must strike a balance between reducing reverse recovery current and maintaining efficient switching transitions.

By following these guidelines, the inductor Lr can be properly sized to optimize the performance of the auxiliary circuit, ensuring smooth operation and high efficiency in the ZVS-PWM converter. Further adjustments can be made based on experimental validation and system-level requirements.

5. Experimental Results

An experimental proof-of-concept prototype of the proposed converter was built to confirm its feasibility. The converter was built according to the same specifications as in the design example with input voltage V_{in} = 70 V, output V_o = 375 V, maximum output power $P_{o,max}$ = 700 W and switching frequency f_{sw} = 100 kHz. The main power boost circuit was implemented as described in the design example. IRFP840 MOSFETs were used for the two auxiliary switches and 15ETX06 diodes for diodes D_1, D_2, D_3. The values of L_r and C_r were L_r = 8.2 µH and C_r = 44 nF.

Typical experimental waveforms are shown in Figure 11. Figure 11a,b shows the current waveform of L_r, I_{Lr}, and the gating signals of the two auxiliary switches. Since the positive part of I_{Lr} and the negative part of I_{Lr} represent the currents through S_{aux1} and S_{aux2} respectively, it can be seen that both switches can be turned off softly with ZCS. Figure 11c shows the gating signal and the drain source voltage of a main power switch. It can be seen that the switch turns on with ZVS, as the voltage across the switch is zero before it is turned on. Figure 11d shows the auxiliary inductor current and capacitor voltage waveforms. It can be seen that whatever energy is placed in C_r is removed before the auxiliary circuit is reactivated.

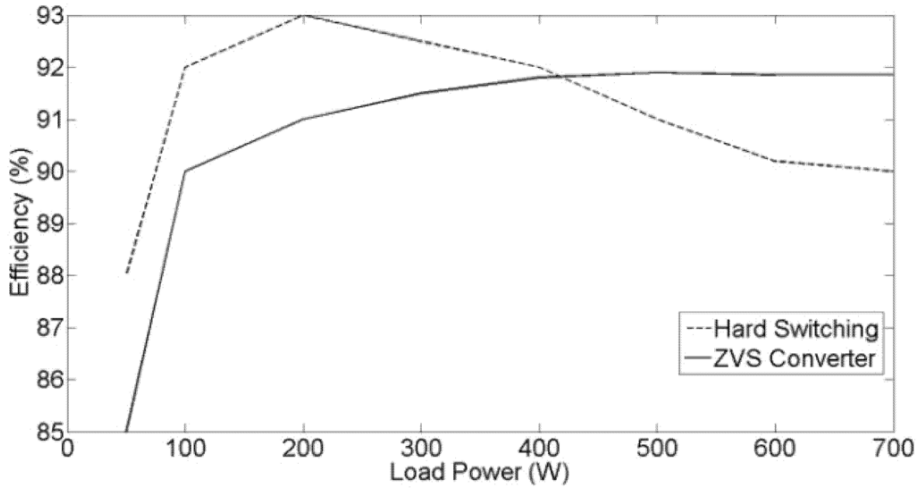

Figure 9. Efficiency *vs.* load power.

6. Conclusion

This paper presents a novel auxiliary circuit designed for ZVS-PWM converters, specifically targeted at applications involving paralleled MOSFETs for higher current handling. The proposed auxiliary circuit incorporates two auxiliary switches, addressing the limitations of conventional single-switch auxiliary circuits, which are typically constrained to lower current applications. By incorporating this dual-switch design, the circuit provides enhanced performance for high-current converters, reducing switching losses and improving overall efficiency.

In this study, the operation of a ZVS-PWM boost converter with the newly introduced auxiliary circuit was thoroughly described. The steady-state behavior of the converter was analyzed, and a systematic design procedure was developed and demonstrated through a practical example. The proposed design approach was validated through experimental results obtained from a prototype converter, confirming the feasibility and effectiveness of the auxiliary circuit in high-current applications.

It is important to emphasize that the proposed converter is specifically designed for high-current applications, a domain where traditional ZVS-PWM converters are typically not employed due to their limitations in handling elevated currents. For typical, lower-current ZVS-PWM converter applications, more straightforward, cost-effective, and conventional approaches are generally more suitable. However, the advancements introduced by this novel auxiliary circuit open new possibilities for ZVS-PWM converters in demanding industrial applications where high efficiency and reduced switching losses are paramount.

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