High-Frequency Soft-Switching DC-DC Converters for Voltage and Current DC Power Sources

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Abstract: The paper presents soft switching PWM DC-DC converters using power MOSFETs and IGBTs. The attention is focused mainly on the full-bridge converters suitable for high power applications. The properties of the PWM converters are described in comparison to other categories of soft switching converters. An overview of the switching techniques using in the DC-DC converters is included. Considerations are also given to the control methods. The principles of the switching and conduction losses reduction in the PWM converters are illustrated. Various types of snubber circuits are mentioned and their operation and limitations are discussed.

Keywords: DC-DC converter, PWM converter, soft switching, snubber

1 Introduction

One of the major trends in power electronics is increasing the switching frequencies. The advances in semiconductor fabrication technology have made it possible to significantly improve not only voltage – and current capabilities but also the switching speed. The faster semiconductors working at high frequencies result in the passive components of the converters – capacitors, inductors and transformers – becoming smaller thereby reducing the total size and weight of the equipment and hence to increase the power density. The dynamic performance is also improved [1], [2], [5], [20].

This frequency elevation is responsible for the growing importance of pulse-width modulation on the one hand and for the use of resonance on the other hand. Another important trend resides in reduction of voltage and current stresses on the semiconductors and limitation of the conducted and radiated noise generated by the converters due large di/dt and du/dt [1], [2], [5], [19], [21].

Both these requirements, size and noise, are minimised if each switch in a converter utilises soft switching technique to change its status. The converter topologies and the switching strategies, which result in soft switching, are discussed in this paper.

2 PWM DC-DC Converters

2.1 Power Stages of the PWM DC-DC Converters

The full-bridge and half-bridge converters shown in Fig. 1 and Fig. 2 are mostly used in high power applications.

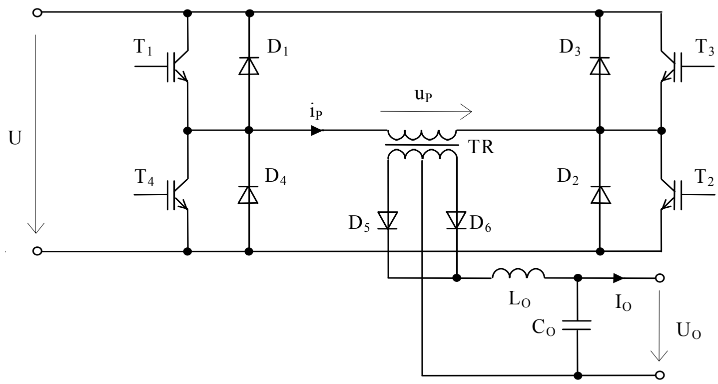

Figure 1
Full–bridge converter

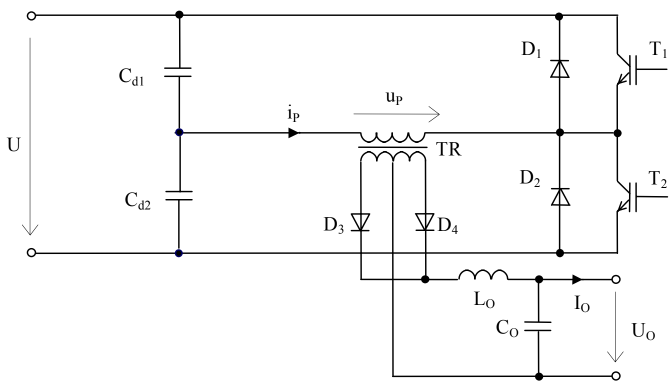

Figure 2
Half–bridge converter

Pulses of opposite polarity are produced on the primary and secondary windings of the transformer by switching of the transistor.

In a connection with the half–bridge inverter, the capacitors C_{d1} and C_{d2} establish a voltage midpoint between zero and the input dc voltage.

The input voltage is equally divided between the capacitors. The relationship between the input and output voltage for the half–bridge is

$$\frac{V_0}{V_d} = \frac{N_2}{N_1}.D \tag{1}$$

and for the full–bridge is

$$\frac{V_0}{V_d} = 2\frac{N_2}{N_1}.D \tag{2}$$

where duty cycle D= t_{on}/T and 0<D<0,5.

Comparison of the full–bridge (FB) converter with the half–bridge (HB) converter for identical input and output voltages and power ratings requires the following turn’s ratio:

$$\left(\frac{N_2}{N_1}\right)_{HB} = 2\left(\frac{N_2}{N_1}\right)_{FB} \tag{3}$$

Neglecting the ripple in the current through the filter inductor at the output and assuming the transformer magnetizing current to be negligible in both circuits, the transistor currents I_C are given by

$$(I_C)_{HB} = 2\,(I_C)_{FB} \tag{4}$$

In both converters, the input voltage U appears across the switching transistors. However, they are required to carry twice as much current in the half–bridge converter. Therefore, in high power applications, it may be advantageous to use a full bridge over a half bridge to reduce the number of paralleled transistors in the switch.

2.2 PWM Strategies for Full–Bridge Converter

The conventional control diagram used for hard driven converters is shown in Fig. 3. The transistors (T_1, T_2) and (T_3, T_4) are switched as pairs alternatively at the selected switching frequency, which alternately places the transformer primary across the input supply U for same interval t_{on}. The maximum duty cycle is 50% (D=0.5).

A disadvantage of this switching mode is that when all four switches are turned off, the energy stored in the leakage inductance of the power transformer causes severe ringing with junction capacitance of switching devices.

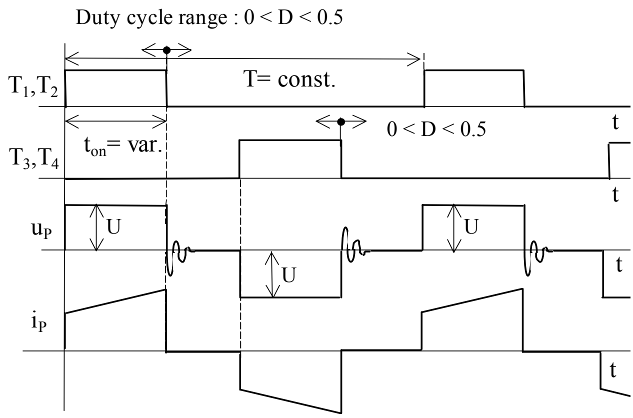

Figure 3

Waveforms of hard–switching converter with conventional PWM

The control scheme in Fig. 4 is almost the same as previous except that the duty cycle in one leg (transistors T_1, T_4) is constant (D=0.5) and in second leg (transistors T_2, T_3) is variable in a range between zero and 50%.

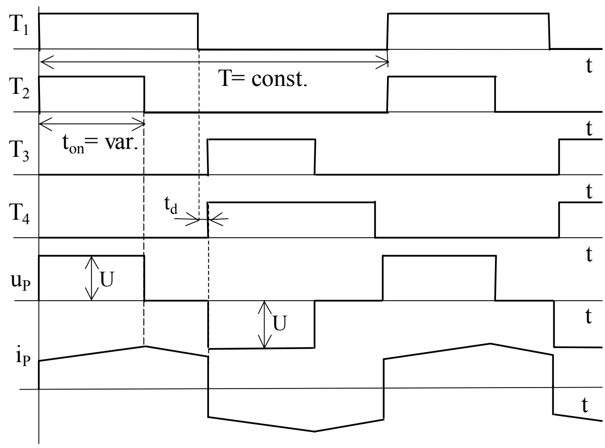

Figure 4

Waveforms of converter with modified PWM control

The output voltage can be controlled also via phase control as shown in Fig. 5, [8], [9], [13], [17].

Both legs (transistors T_1, T_4 and T_2, T_3) of the bridge operate with a 50% duty cycle, and the phase shift between the legs is controlled. When the two legs operate in phase, the differential voltage applied to the transformer is zero, and zero DC output voltage is obtained. When the two legs of the bridge are in opposite phase, the differential voltage applied to the transformer, and also the output voltage is maximal.

There are also other derivations of the switching strategies mentioned above [2], [10], [12], [15], [16], [21].

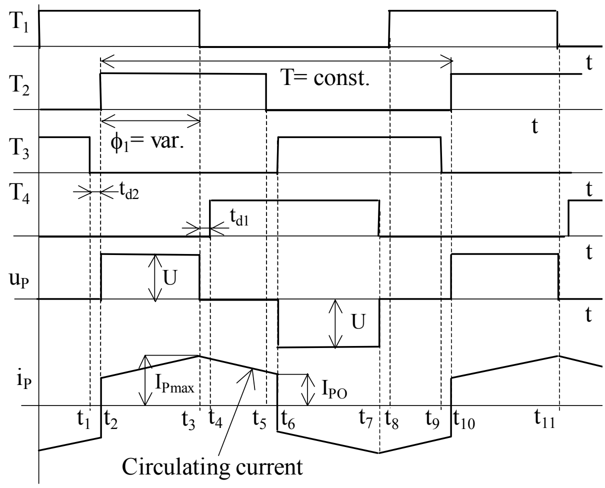

Figure 5

Waveforms of converter with phase–shifted PWM control

The phase shift pulse width modulation (PS-PWM) leads to asymmetrical switching waveforms. The **leading leg** consists of transistors T_1, T_4 and the **lagging leg** consists of transistors T_2, T_3. The transistor currents in these legs are not symmetrical. The PS-PWM control strategy leads to zero-voltage turn-on of the transistors in both of legs, as it is evident from oscillograms in Figs. 6 and 7.

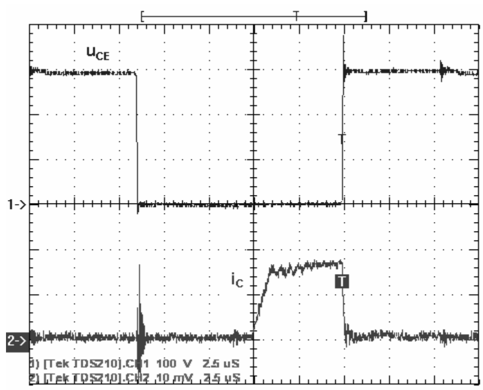

Figure 6

Transistor voltage u_{CE1} and current i_{C1} in the leading leg at turn–on and turn–off

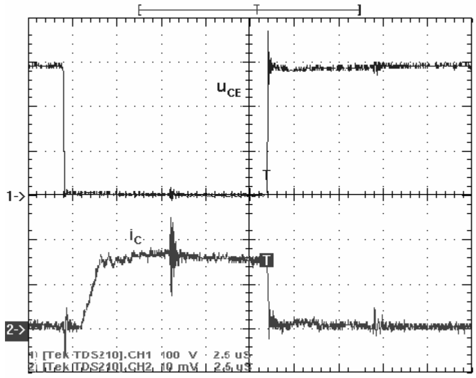

Figure 7

Transistor voltage u_{CE2} and current i_{C2} in the lagging leg at turn–on and turn–off

The turn-off losses occur as a result of hard turn-off of the transistors in both legs. The circulating current appears after turn-off of the transistors in the leading leg during freewheeling interval and consequently the conduction losses are increased in the lagging leg transistors.

3 Soft Switching PWM Converters

The soft switching PWM converter is defined here as the combination of converter topologies and switching strategies that result in zero–voltage and/or zero–current switching. This type of soft switching converter has been referred to as different names in the literature [1], [2], [3], [6], [9], [11], [14], [15], [18]. They are called also pseudo–resonant, quasi-resonant, resonant transition, clamped voltage topologies and other. In these converters the resonant transition is employed only during a short switching interval. The output voltage is usually controlled by PWM with constant switching frequency.

Soft switching PWM converters can be classified as follows:

1 ZVS PWM converters

2 ZCS PWM converters

3 ZVS ZCS PWM converters

This classification is explained further.

3.1 ZVS PWM Converters

The simplest ZVS PWM full bridge converter is shown in Fig. 8. The converter snubbers consist of capacitances C_1–C_4 and inductance L_R, which are represented by transistor and diode output capacitances and transformer leakage inductance respectively.

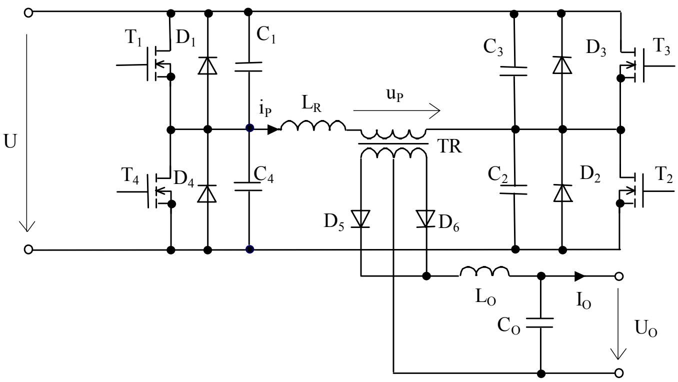

Figure 8
Full–bridge ZVS PWM converter

The converter is controlled by the phase–shifted PWM technique, which is shown in Fig. 5.

The transistors (MOSFETs or IGBTs) in leading or lagging leg are turned–on while their respective anti-parallel diodes conduct. Since the transistor voltage is zero during the entire turn-on transition, switching loss does not occur at turn–on (Figs. 9 and 10).

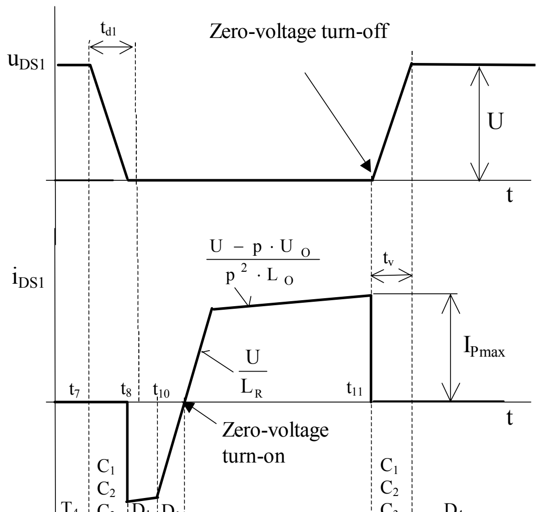

Figure 9
Switch (transistor MOSFET T_1 and its body diode D_1) voltage u_{DS1} and current i_{DS1} during turn–on and turn–off (leading leg)

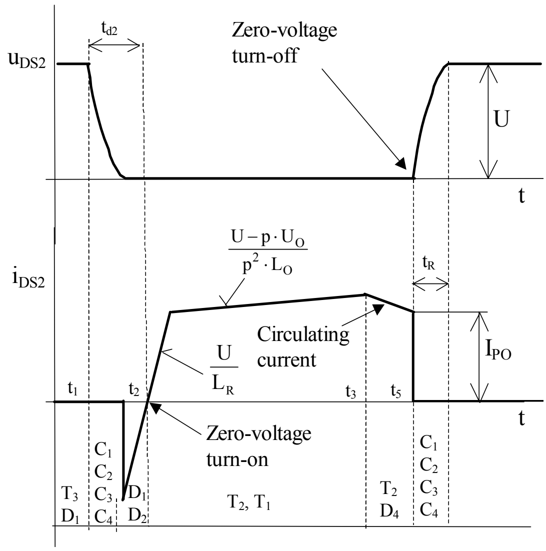

Figure 10

Switch (transistor MOSFET T_2 and its body diode D_2) voltage u_{DS2} and current i_{DS2} during turn–on and turn–off (lagging leg)

By utilising small snubber capacitors C_1 – C_4 the turn–off losses are sufficiently reduced. If the transistor turn–off time is sufficiently fast, then the transistor is switched fully off before the collector voltage rises significantly above zero, and thus negligible turn–off switching loss is incurred (Figs. 9 and 10).

The detail of the turn-off process of the IGBT transistor in the leading leg of the converter is shown in Fig. 11. Using snubber capacitors in parallel to transistors, the turn-off losses are remarkably decreased.

The ZVS converter exhibits low primary–side switching loss and generated EMI.

However, conduction losses are increased with respect to an ideal hard switching PWM full bridge topology.

At light load, the leakage inductance energy is not sufficient to ensure zero–voltage switching in the lagging leg of the converter. This critical load condition is also a function of the line condition. The worst case is high input voltage when more capacitive energy is required.

Another consideration is the delay time from the turn–off T_4 until the turn–on of T_1 and visa versa. If the delay time t_d is too short, then the device capacitance may not be fully discharged. However, if the delay time t_{d2} (Fig. 5, Fig. 10) is too long, the capacitor voltage will peak, continue to resonate and drop. Fortunately, the time of peak charge is relative independent of the input voltage and load condition and is equal to one quarter of L_RC time constant [13].

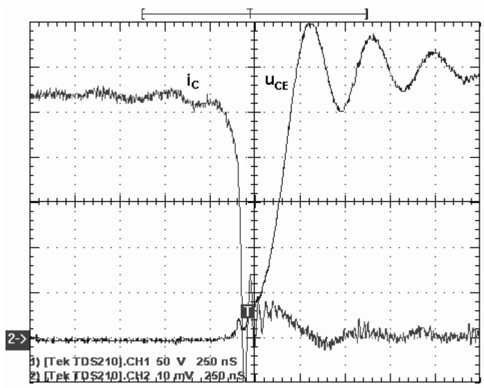

Figure 11
Transistor voltage u_{CE1} and current i_{C1} at turn–off –detail

The secondary–side diodes switch at zero current. This leads to switching losses and ringing as a result of interaction of diodes capacitance with the leakage inductance of the transformer. Additional snubber circuitry is usually required, for prevention of excessive diode voltage stress.

To remove the above-mentioned disadvantages a lot of derivations of the ZVS PWM converters were developed. The penalty for the improvement is usually higher complexity of the converter topology.

3.2 ZCS PWM Converters

The ZCS PWM converters can be derived from the ZVS PWM converters by applying the duality principle [14].

The scheme of the full–bridge ZCS PWM (FB–ZCS–PWM) converter is shown in Fig. 12, where L_R is the resonant inductor and C_R is resonant capacitor.

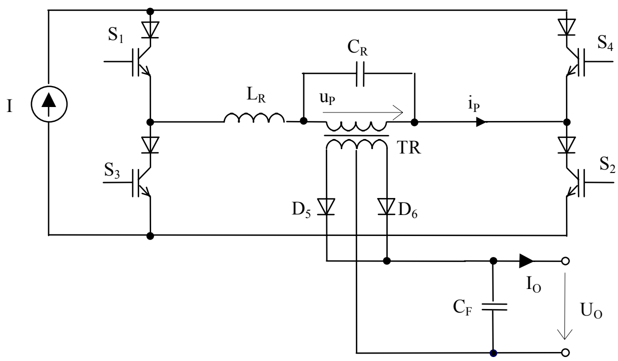

Figure 12
Basic circuit diagram of the FB ZCS-PWM converter

The transformer leakage inductance, the rectifier's junction capacitances, and the transformer winding capacitances can be utilised in this circuit.

Similar to the FB–ZVS–PWM converter, the FB–ZCS–PWM converter also uses phase–shift control at constant switching frequency to achieve required converter operation (Fig. 13).

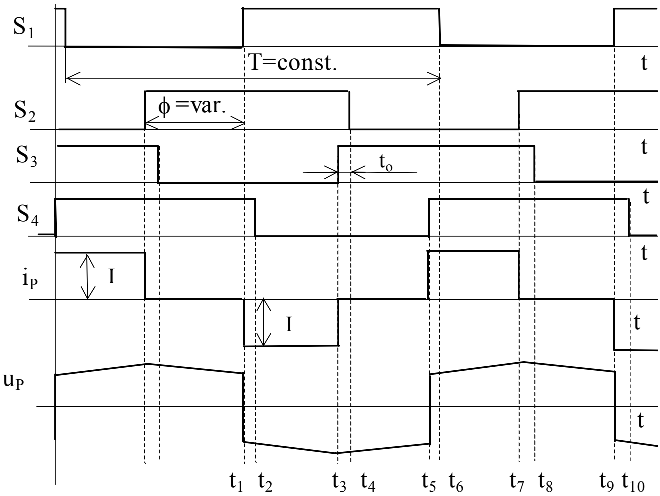

Figure 13
Idealised waveforms of the FB ZCS PWM converter

The switches must have reverse-voltage blocking capability. The switch can be implemented by an IGBT or a MOSFET in series with a reverse blocking diode, an IGBT with reverse–voltage blocking capability, a MCT, or a GTO. An important advantage of the circuit is that the rectifier diodes do not suffer from reverse recovery problem since they commutate with zero–voltage switching.

This feature makes the converter attractive for applications with high output voltage e. g. power factor correction circuits, where the rectifiers suffer from severe reverse–recovery problems when conventional PWM, ZVS–QRC, or ZVS–PWM converter techniques are used [14].

The efficiency of the converter drops significantly at low line and heavy load since the switches begin to lose zero current switching.

The turn-on and turn-off process of the switches are shown in Figs. 13 and 14 respectively.

Some dual characteristics of the FB ZVS PWM converter and FB ZCS PWM converter are summarised in Table 2.

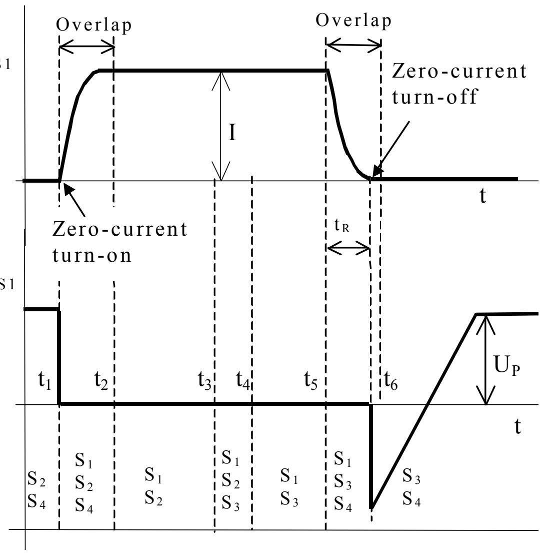

Figure 14

Switch S_1 voltage and current during turn-on and turn-off

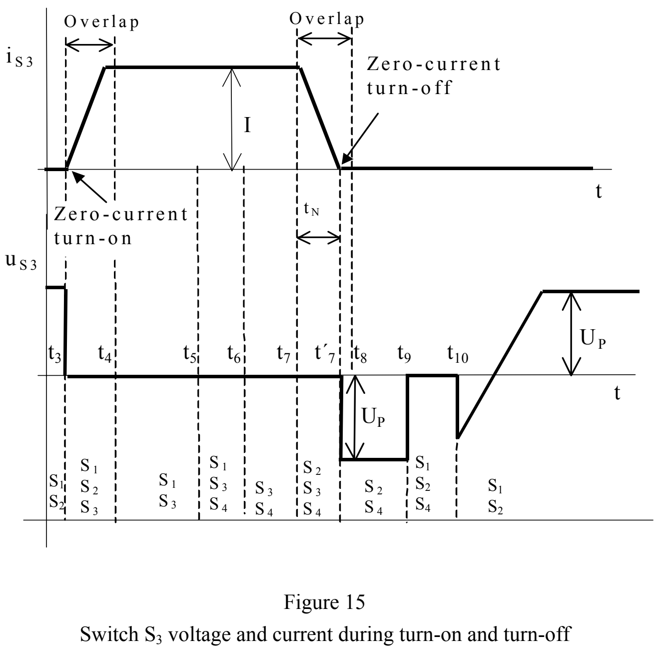

Figure 15

Switch S_3 voltage and current during turn-on and turn-off

Table 2
Some dual characteristics of the FB ZVS PWM converter and FB ZCS PWM converter

	FB-ZVS-PWM	**FB-ZCS-PWM**
Topology type	Buck type	Boost type
Switching conditions for active switches	Zero-voltage switching	Zero current switching
Switching conditions for rectifiers	Zero current switching	Zero-voltage switching
Soft switching easy to achieve at	Heavy load	Light load
Implementation of active switches	Diode in parallel with transistor	Diode in series with transistor

3.3 Zero-Voltage Zero-Current Switching PWM Converters

This type of converter is very attractive for high voltage, high power (>10 kW) applications where IGBTs are predominantly used as a power switches [1], [3], [4], [10], [12], [22].

The operating frequency of IGBTs is normally limited to 20-30 kHz because of their current tailing problem. To operate IGBTs at higher switching frequencies, it is required to reduce the turn-off switching losses. ZVS with substantial external capacitor or ZCS can be a solution. The ZCS, however, is deemed more effective since the minority carriers are swept out before turning off.

The zero-voltage zero-current switching (ZVZCS) PWM converters are derived from the full-bridge phase-shifted zero-voltage (FB-PS-ZVS) PWM converters. The PS-ZVS PWM converter is often used in many applications because this topology permits all switching devices to operate under zero-voltage switching by using circuit parasitics such a transformer leakage inductance and devices junction capacitance.

However, because of phase-shifted PWM control, the converter has a disadvantage that circulating current flows through a transformer and switching devices during freewheeling intervals (Fig. 4, Fig. 7).

The circulating current is a sum of the reflected output current and transformer primary magnetizing current. Due to circulating current, RMS current stresses of the transformer and switching devices are still high compared with that of the conventional hard-switching PWM full-bridge converter (Fig. 1). To decrease the circulating current to zero and thus to achieve zero-current switching for lagging leg, various snubbers and/or clamps connected mostly at secondary side of transformer are applied.

The principle by using all of the snubbbers and/or clamps is to secure disconnection of the transformer secondary side, as it is very simplified shown in

Fig. 16. It is usually realised by application of the reverse bias for the secondary side rectifier when transformer secondary voltage in the freewheeling interval becomes zero. The output rectifier (D_5, D_6) is then reverse biased and the secondary windings of the transformer are opened.

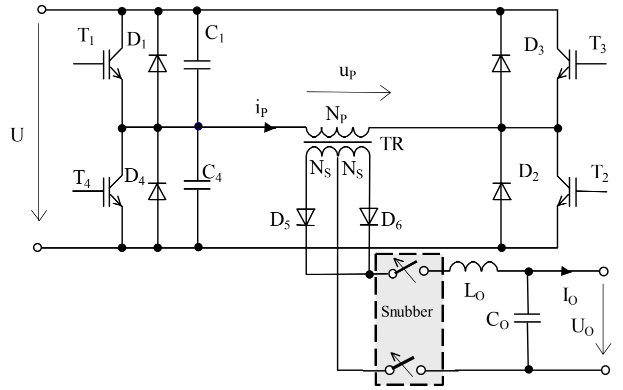

Figure 16

Principle of the ZVZCS converter operation

Therefore, both primary and secondary currents of the transformer become zero. Only a low magnetising current circulates during freewheeling interval as shown in Fig. 17. Thus, the RMS current of the transformer and switches are considerably reduced in the freewheeling interval.

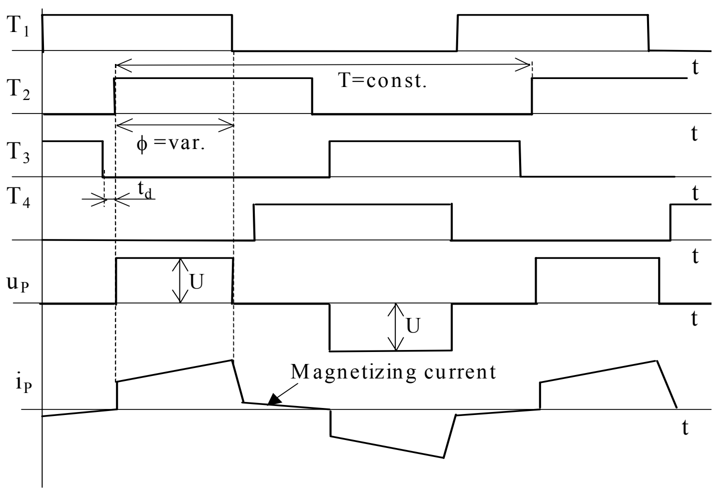

Figure 17

Operation waveforms of ZVZCS PWM converter

Hence, the converter achieves nearly zero-current switching for the lagging leg (transistors T_2, T_3) due to minimised circulating current during interval of lagging leg transition and achieves zero-voltage switching for leading leg (transistors T_1,

T_4) due to reflected output current ($I_O/p=I_P$, $p=N_p/N_S$) during the interval of leading leg transition.

Several passive and active snubber and clamp circuits were developed to resolve the problem concerning the resetting the primary current of the transformer to achieve zero-current switching of the switches in the lagging leg of the converter [21].

An example of ZVZCS PWM converter is shown in Fig. 18. ZVS of the leading leg is achieved by the same manner as that of the ZVS full-bridge PWM converter, while the ZCS of the lagging leg is achieved by resetting the primary current during freewheeling period by using active clamp in the secondary side, which needs an additional active switch. Oscillogram of the collector-emitter voltage u_{CE2} and collector current i_{C2} in the lagging leg at turn–on and turn–off is shown in Fig. 19. The transistor is turned-on at zero voltage and turned-off at zero current. The circulating current does not occur, only negligible magnetizing current flows during freewheeling interval through primary winding of transformer. This combination of switching is very effective for IGBT transistors, which have problems at turn-off due to tail current effect.

The converter in Fig. 18 is operating very well at nominal load, but it is not capable operating over wide load range (from no-load conditions to short circuit) with zero-voltage or zero-current switching for all power switches.

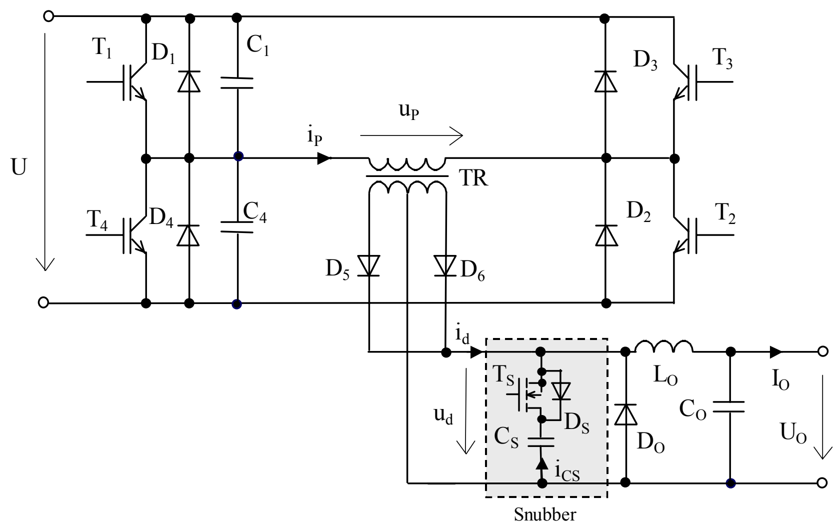

Figure 18
ZVZCS DC-DC PWM converter

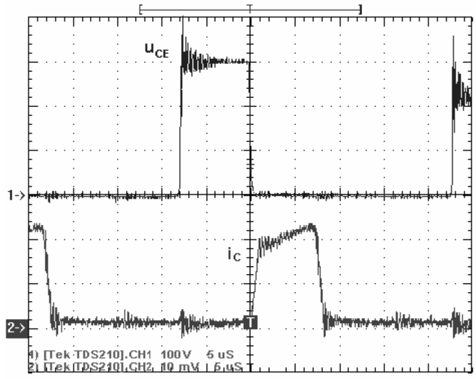

Figure 19

Transistor voltage u_{CE2} and current i_{C2} in the lagging leg at turn–on and turn–off

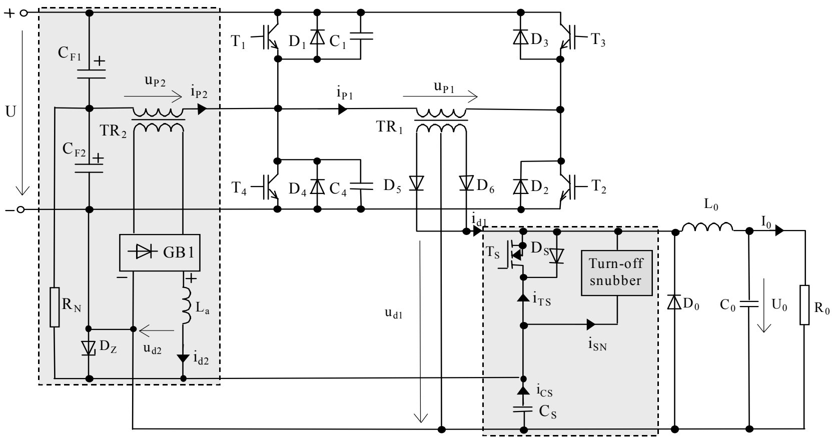

Figure 20

Improved ZVZCS DC-DC PWM converter

In order to achieve soft switching at no-load conditions and at short circuit the auxiliary circuits are needed. The example of such patented arrangement is shown in Fig. 20 [7], [22].

The auxiliary transformer TR_2 is the main part of the auxiliary circuits in this converter.

The transformer TR_2 should have considerably large air-gap to ensure sufficiently high magnetizing current i_{m2} and at the same time to prevent core saturation. The saw-tooth magnetizing current i_{m2} ensures the zero-voltage turn-off of the transistors T_1, T_4 not only at light load but also at no-load conditions.

Simultaneously, charging or discharging of the capacitors C_1, C_4 by magnetizing current i_{m2} avoids high current spikes at transistors T_1, T_4 turn-on at light load and no-load.

In order not to loose the zero-current turn-off of the transistors T_2, T_3 at short circuit, it is necessary to charge up the capacitor C_s to the rated value of the voltage. The capacitor C_s can be charged from the rectifier GB1, which is connected to the secondary winding of the auxiliary transformer TR_2. Soft switching and reduction of circulating currents for full load range are achieved in this converter. The converter is especially suited for application where short circuit and no-load are normal states of the converter operation, e.g. arc welding.

Conclusion

Principles of the zero-voltage and zero-current switching in PWM full-bridge high-frequency converters are described. The overview and division of the prospective soft-switching PWM converters for high power application is presented.

Acknowledgement

This work has been supported by Grant Agency of the Slovak Republic VEGA No. 1/2178/05.

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