ZVZCS PWM DC-DC CONVERTER WITH CONTROLLED OUTPUT RECTIFIER

Jaroslav DUDRIK, Vladimír RUŠČIN
Department of Electrical, Mechatronic and Industrial Engineering, Faculty of Electrical Engineering and Informatics,
Technical University of Košice, Letná 9, 042 00 Košice, Slovak Republic, tel.: +421 55 602 2276,
e-mail: jaroslav.dudrik@tuke.sk, vladimir.ruscin@tuke.sk

ABSTRACT

A new zero-voltage zero-current switching (ZVZCS) full-bridge phase-shifted PWM converter with controlled output rectifier is presented in this paper. Zero-voltage turn-on and zero-current turn-off for all power switches of the inverter is achieved for full load range from no-load to short circuit by using new secondary energy recovery clamp and modified PWM control strategy. Moreover by adding secondary energy recovery clamp the zero-current turn-on and zero-voltage turn-off for rectifier switch is ensured. The principle of operation is explained and analysed and simulation results are presented.

Keywords: *soft switching, ZVZCS converter, switched-mode power supply*

1. INTRODUCTION

The soft switching PWM converters are very suitable for high voltage, high power applications where IGBTs are predominantly used as power switches.

The conventional phase shifted PWM converters are often used in many applications because their topology permits all switching devices to operate under zero-voltage switching by using circuit parasitics such as power transformer leakage inductance and devices junction capacitance.

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

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 those of the conventional hard-switching PWM full-bridge converter. To achieve soft switching and decrease the circulating current, various snubbers, auxiliary circuits and/or clamps connected mostly at the secondary side of power transformer are applied [1] – [18].

The first way of decreasing the circulating current is achieved by application of the reverse bias for the output rectifier when the secondary voltage of the transformer 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 (Fig. 1).

Consequently, both primary and secondary currents of the transformer become zero. Only a low magnetising current circulates during freewheeling interval as shown in Fig. 2. Thus, the RMS current of the transformer and switches are considerably reduced in the freewheeling interval.

Hence, the converter achieves nearly zero-current switching for the right leg (transistors T_2, T_3) due to minimised circulating current during the interval of right leg transition and achieves zero-voltage switching for the left leg (transistors T_1, T_4) due to reflected output current ($I_O/n=I_P$, $n=N_P/N_S$) during the interval of left leg transition.

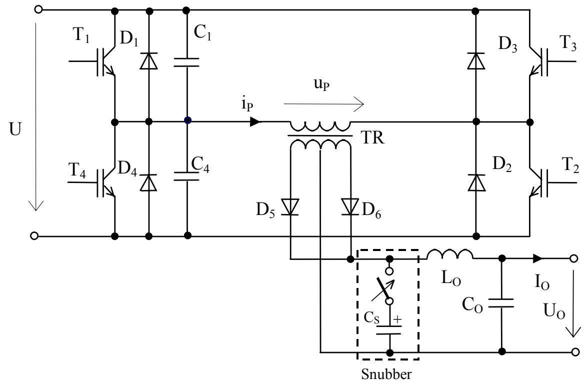

Fig. 1 Principle of the ZVZCS converter operation

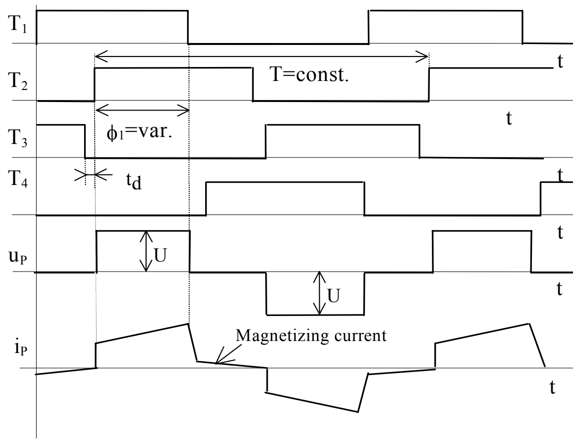

Fig. 2 Operation waveforms of ZVZCS PWM converter

The another method how to reduce the circulating current in the converter is using the controlled output rectifier (Fig. 3) [12], [13]. Turn-off losses are reduced by non dissipative turn-off snubbers (capacitors C_1 – C_4). Reduction of turn-on losses is achieved by using the leakage inductance of power transformer.

In the converters mentioned above, the inverter switches operate under zero-voltage switching either in one leg (converter in Fig. 1) or in both legs of the converter (converter in Fig. 3).

However, the optimal switching for IGBTs is zero-voltage turn-on and mainly zero-current turn-off due to elimination of the current tail influence, which has considerably high involvement in creation of the IGBT turn-off losses.

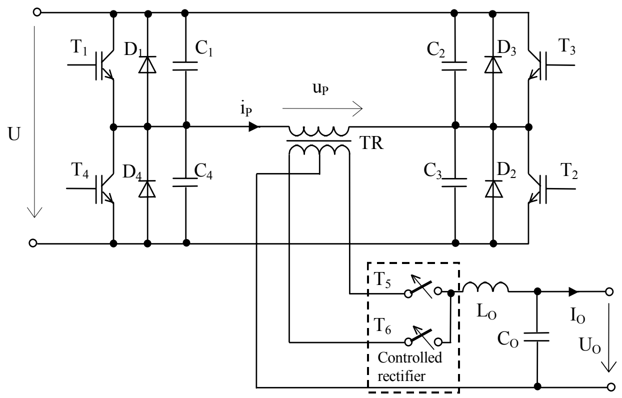

Fig. 3 Principle of the ZVS converter operation

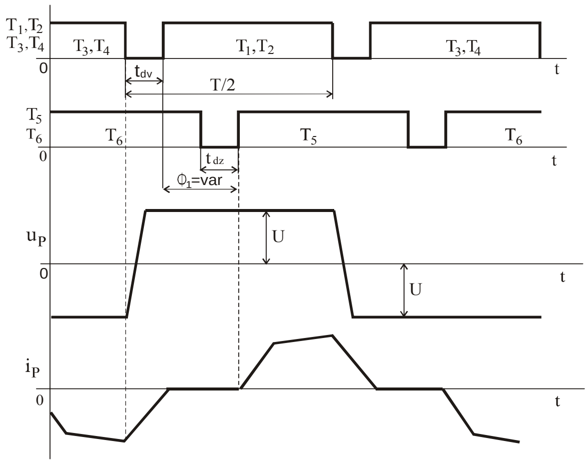

Fig. 4 Operation waveforms of ZVS PWM converter

2. POWER CIRCUITS OF THE PROPOSED CONVERTER

To avoid the problems mentioned above, the topology of the following ZVZCS converter was proposed.

The proposed DC-DC converter shown in Fig. 5 consists of high-frequency inverter, power transformer, output rectifier, output secondary switch and output filter.

The main part of the converter includes high frequency full-bridge inverter consisting of four ultrafast IGBT´s T_1-T_4 and freewheeling diodes D_1-D_4. The secondary winding of the high-frequency step-down power transformer TR is connected through a fast recovery rectifier D_5, D_6 and secondary switch T_S to output filter consisting of smoothing choke L_O and capacitor C_O.

The converter is controlled by modified pulse-width modulation (Fig. 6), and consequently the zero-voltage turn-on and zero-current turn-off all of the transistors T_1-T_4 in the inverter are reached.

The semiconductor switch T_S in the secondary side is used to reset secondary and consequently also primary current. The transistor T_S operates with double switching frequency. At turn-off of the switch T_S the energy stored in leakage inductance is clamped by D_C and C_C and then transferred through D_S and L_S to the load. By using non-dissipative turn-off snubber to reduce turn-off losses of the transistor T_S, the overall efficiency is increased.

The additional energy recovery clamp is very simple, consisting of only few components and so the additional cost is not high.

3. OPERATION PRINCIPLE

The basic operation of the proposed soft switching converter has nine operating modes (intervals) within each half cycle. The switching diagram and operation waveforms are shown in Fig. 6.

It is assumed that all components and devices are ideal.

The turn-off snubber used for decreasing turn-off losses of the secondary switch was not included into circuit analysis.

Interval** **(t_0-t_1): The transistors T_1, T_2 and T_S are turned on at t_0. The primary current (only magnetizing current) flows through diodes D_1, D_2 and consequently the transistors T_1 and T_2 are turned on with ZVS.

The collector current of the transistor T_S starts to flow in the loop T_S-C_C-D_S-L_S-L_O-C_O and capacitor C_C is discharged. So, the rise of the collector current is in resonant way with the resonant frequency ω_{R1} different at no-load and short circuit in a range:

$$\sqrt{(L_O + L_{CS}) \cdot \frac{C_O \cdot C_C}{C_O + C_C}} \le \omega_{R1} \le \sqrt{(L_O + L_{CS}) \cdot C_C} \qquad (1)$$

***Interval** (t_1-t_2):* The transformer leakage inductance L_{LP} reflected to the primary side causes that primary current i_P is linearly increased with the slope U/L_{LP} while the secondary voltage u_S is zero as a result of commutation between output freewheeling diode D_O and rectifier diode D_5.

The discharging of the clamp capacitor C_C causes the current overshoot at turn-on of the transistor T_S, which maximum is limited by the value of the smoothing inductance current i_{LO}.

***Interval** (t_2-t_3):* At t_2 the commutation between diode D_5 and output freewheeling diode D_O is finished. At t_3 the clamp capacitor current commutates to clamp diode D_C.

***Interval** (t_3-t_4):* Transistors T_1 and T_2 are conducting and the energy is delivered from the source to the load via power transformer TR, diode D_5 and smoothing choke L_O and from inductance L_S in the loop L_S-L_O-C_O-D_C-D_S. So, the smoothing inductance current is a sum of the secondary current and inductance L_S current:

$$i_O = i_S + i_{LS} \qquad (2)$$

***Interval** (t_4-t_5):* The primary current increases with the slope:

$$\frac{di_p}{dt} = \frac{U - n \cdot U_O}{L_{LP} + n^2 \cdot L_O} + \frac{U}{L_m} \tag{3}$$

Where $n = \frac{N_P}{N_S}$ is power transformer turns ratio and L_m magnetizing inductance of the power transformer TR.

***Interval** (t_5-t_6):* At t_5 the secondary transistor T_S turns off. At that time the commutation between transistor T_S and clamp diode D_C occurs and charging of the clamp capacitor C_C starts. This commutation time can be neglected, because only parasitic inductance of wires is in the commutation loop T_S-D_C-C_C. Afterwards the commutation between D_C, D_5 and output freewheeling diode D_O starts. Because in the commutation path a relatively large leakage inductance of the transformer is found, the commutation is slow.

In the mentioned commutation path the resonance occurs and rise of the current depends on the resonant frequency ω_{R2}:

$$\omega_{R2} = \sqrt{(L_O + L_{LS}) \cdot \frac{C_O \cdot C_C}{C_O + C_C}} \quad \text{for } R_O = \infty \tag{4}$$

$$\omega_{R2} = \sqrt{(L_O + L_{LS}) \cdot C_C} \quad \text{for } R_O = 0 \tag{5}$$

During the commutation the energy stored in the leakage inductance is transferred to the clamp capacitor C_C and consequently an over-voltage ΔU_S appears on transformer secondary voltage.

Its value can be calculated from equation (the output current ripple is neglected):

$$\frac{1}{2} L_{LS} \cdot I_O^2 = \frac{1}{2} C_C \cdot U_{CC}^2 \tag{6}$$

where L_{LS} is the transformer leakage inductance reflected to the secondary side and U_{CC} is maximum clamp capacitor voltage.

Then

$$\Delta U_S = U_{CC} - U \frac{N_S}{N_P} \tag{7}$$

***Interval** (t_6-t_7):* Only small magnetizing current i_m flows through primary winding of TR. The output current flows trough output freewheeling diode D_O.

***Interval** (t_7-t_8):* In this interval the transistors T_1 and T_2 are turned off with ZCS. Only small magnetizing current i_m is switched off by transistors T_1 and T_2. The magnetizing current charges or discharges the internal output capacitances C_{OSS1} – C_{OSS4} of the IGBT transistors T_1 – T_4 respectively.

The minimum dead time t_d for the transistors in the leg is given by:

$$t_{d,\min} \geq t_{recom} \tag{8}$$

where t_{recom} is minority carrier recombination time of IGBTs due to stored charges that could not be removed at turn-off process.

When we take into account also charging and discharging of the capacitances C_{OSS1} – C_{OSS4} by magnetizing current, then minimum dead t_d for achieving of zero voltage turn-on must be:

$$t_{d,\min} \geq \frac{4C_{OSS} \cdot U}{I_{m,\max}} \tag{9}$$

So, in the end of the interval the situation from first interval is repeated for the transistors T_3 and T_4, which turn-on at zero-voltage during conduction of the freewheeling diodes D_3, D_4 (at t_9).

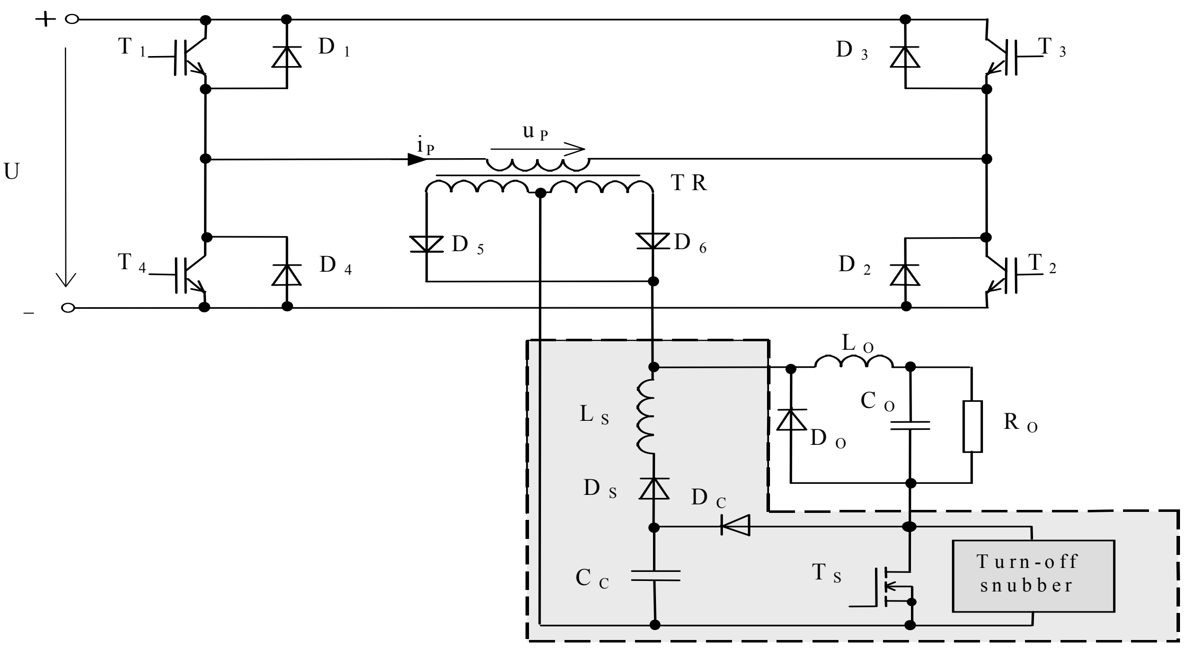

Fig. 5 Scheme of the proposed ZVZCS PWM DC-DC converter

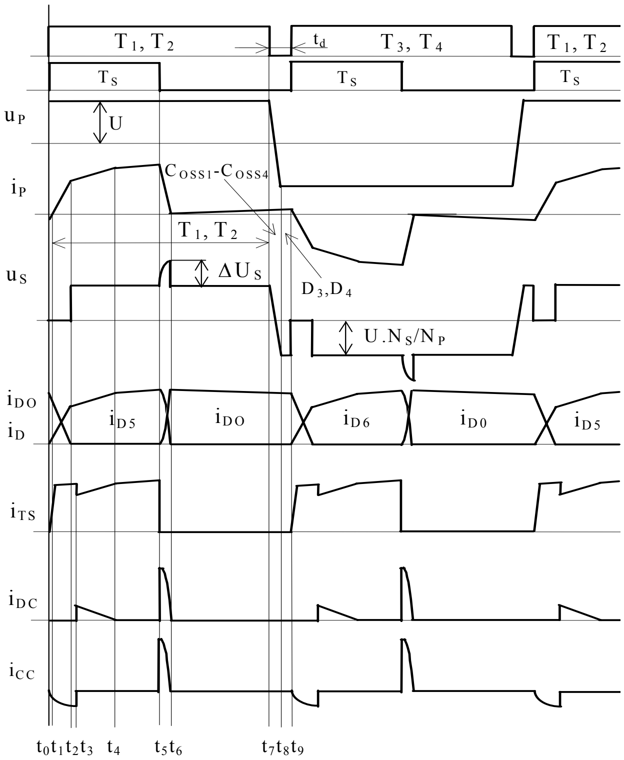

Fig. 6 Operation waveforms of the converter

***Interval* (*t_8-t_9*):** At t_8 the freewheeling diodes D_3, D_4 starts to lead primary current and thus conditions for the zero-voltage turn-on for the transistors T_3 and T_4 are set up.

4. SIMULATION RESULTS

A simulation model in programme Orcad was created to verify the properties of the proposed converter. The simulations were made at input voltage U = 320V.

Parameters:

Transformer TR parameters:
Turns ratio n = 6.5,
Magnetizing inductance L_m = 800 μH,
Leakage inductance L_{LP} = 5 μH.
Clamp circuit parameters:
Clamp capacitor C_C = 220 nF,
Clamp inductance L_S = 1 μH.

The following waveforms were obtained at resistive load.

Fig. 7 shows switch voltage u_{CE4} and switch current $i_{C4}+i_{D4}$ during turn-on and turn-off of the transistor T_4 in the converter. The switch (transistor T_4 including diode D_4) is turned-on under zero-voltage because at turn-on of the transistor T_4 its freewheeling diode D_4 is in on-state. Moreover the rate of rise of the collector current is limited by the leakage inductance L_{LP} of the transformer.

The transistor turn-off losses are negligible because transistor T_4 turns-off only small magnetizing current (about 1 Amp in this case) as can be seen in Fig. 7.

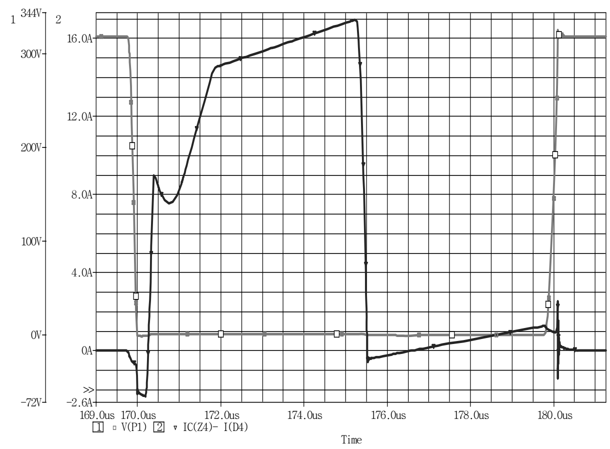

Fig. 7 Switch (transistor T_4 + diode D_4) voltage u_{CE4} and switch current $i_{C4}+i_{D4}$

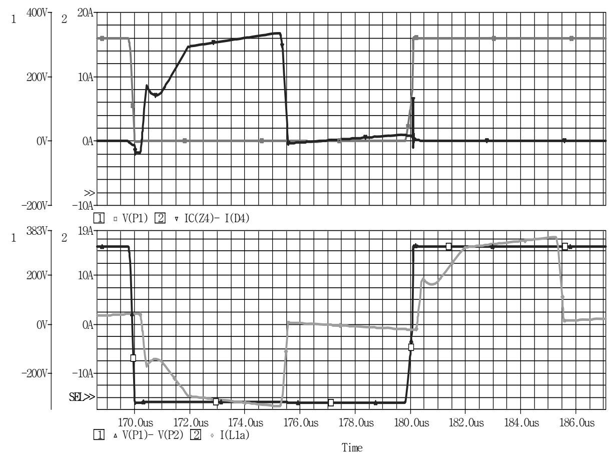

Fig. 8 Switch voltage u_{CE4} and switch current $i_{C4}+i_{D4}$.(upper waveforms), Power transformer TR primary voltage up and primary current i_P, (bottom waveforms)

Fig. 8 shows primary voltage u_P and current i_P of the power transformer TR at output load current above I_0 = 100A (bottom waveforms) in comparison with switch voltage u_{CE4} and switch current i_{C4} of the transistor T_4 (upper waveforms). It is evident that no circulating current flows through primary winding of the power transformer.

After turn-off of the transistor T_4 only a small magnetizing current is conducting through primary winding of the power transformer. Maximum magnetizing current $I_{m,max}$ is approximately 1 Amp. Depending on the dead time t_d it should be high enough for charging or discharging output capacitances C_{OSS1} – C_{OSS4} of the IGBT switches and thus to achieve zero-voltage turn-on.

Collector voltage u_{DS} and collector current i_D of the secondary transistor T_S (bottom waveforms) is shown in Fig. 9. The secondary switch (transistor T_S) is turned-on under zero-current due to influence of the leakage inductance of the transformer L_{LS} reflected to the secondary side and clamp inductance L_S.

The turn-off loss is reduced by clamp capacitor C_C acting as the non-dissipative snubber as it is evident in Fig. 9.

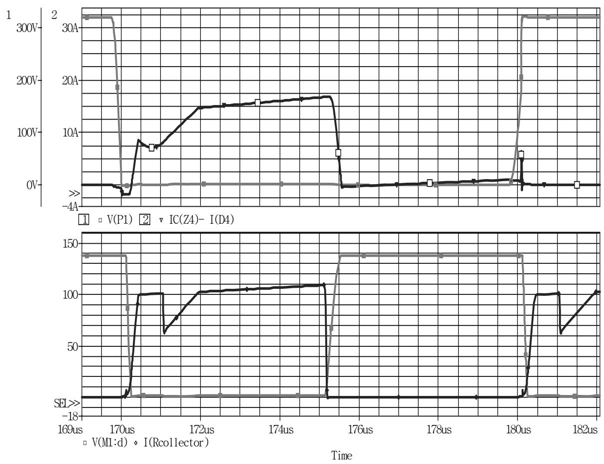

Fig. 9 Switch voltage u_{CE4} and switch current i_{C4}+i_{D4} (upper waveforms) ,Collector voltage u_{DS} and collector current i_D of the transistor T_S, (bottom waveforms)

The clamp diode current is displayed in Fig. 10 together with rectified secondary voltage. Sum of the collector current and clamp diode current equals the value of the smoothing inductance current.

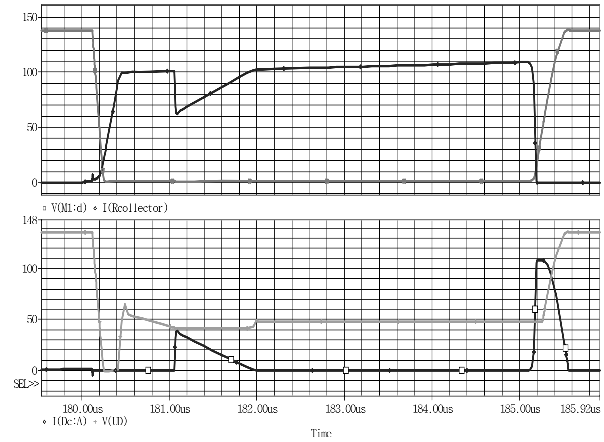

Fig. 10 Collector voltage u_{DS} and collector current i_D of the secondary transistor T_S (upper waveforms)
Rectified secondary voltage ud of the power transformer TR and clamp diode current i_{DC} (bottom waveforms)

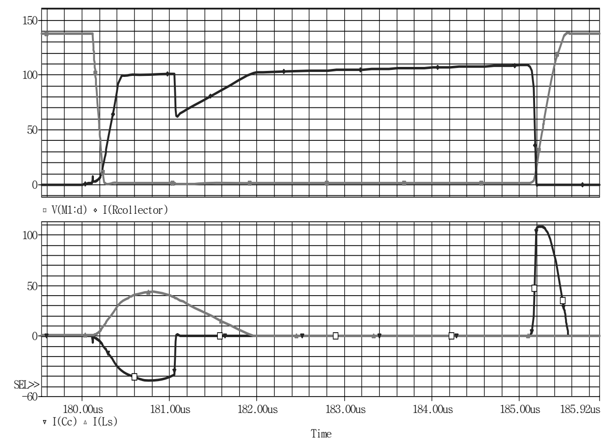

Fig. 11 Collector voltage u_{DS} and collector current i_D of the secondary transistor T_S (upper waveforms)
Clamp capacitor current i_{CC} and snubber inductance current i_{LS} (bottom waveforms)

During commutation between secondary diode, output freewheeling diode, and secondary switch the secondary voltage and accordingly rectified secondary voltage is zero. At turn-off of the secondary switch the secondary and also rectified voltage rises as a result of energy stored in leakage inductance. The over-voltage can be decreased to acceptable value by proper design of the clamp capacitor and clamp inductance.

For completeness Fig. 11 shows also the clamp capacitor current and clamp inductance current (bottom waveforms).

5. CONCLUSION

Soft switching and reduction of circulating currents in the proposed converter are achieved for full load range using secondary side energy recovery clamp in combination with modified PWM.

By proper design it is possible to utilize the magnetizing current of power transformer for charging or discharging output capacitances of the IGBT switches and thus zero-voltage turn-on of the IGBTs to achieve.

If the magnetizing current is not high enough for charging or discharging output capacitances of the IGBT switches, during chosen dead time, then at least zero-current turn-on is reached as a result of leakage inductance of the power transformer.

The IGBT transistors are turned-off almost under zero current. Only small magnetizing current of the power transformer is turned-off by IGBT transistors.

The main task of the proposed secondary energy recovery clamp is transfer of the leakage inductance energy to the load at turn-off of the secondary switch.

Moreover it ensures zero current turn-on and zero voltage turn-off of the secondary switch.

Because this function of clamp is not fully effective when clamp inductance current is continuous, an additional turn-off snubber is employed to improve turn-off process of the secondary switch.

IGBTs in the full bridge inverter operate at almost ideal switching conditions – ZV turn-on and ZC turn-off, which is the main advantage of the proposed converter.

Soft switching of the secondary switch and leakage energy transfer to the load is ensured by energy recovery clamp containing only non-dissipative components.

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BIOGRAPHIES

Jaroslav Dudrik received the M.S. and Ph.D. degrees in electrical engineering from the Technical University of Košice, Slovakia, in 1976 and 1987, respectively. He is currently full professor of Electrical Engineering at the Department of Electrical, Mechatronic and Industrial Engineering, Technical University of Košice, where he is engaged in teaching and research. His primary interest is power electronics. His current field of research includes dc-to-dc converters, high power soft switching converters, converters for renewable energy sources and automotive technique.

Vladimír Ruščin received the M. S. degree in electrical engineering from the Technical University of Košice, Slovakia, in 2006. In the same year, he joined the Department of Electrical, Mechatronic and Industrial Engineering, Technical University of Košice as a Ph.D. student. His primary interest is power electronics. His field of research includes high frequency DC/DC converters, microcontrollers, and soft switching technique in the power converters.