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LOW POWER DC/DC CONVERTER FROM 3 KV TO 300 V – SIMULATION ANALYSIS

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Abstract. *The article presents a 150 W converter, supplied from the railway 3 kV DC traction grid, lowering the voltage to a level of 300 V. The described structure enables the selection of low cost IGBT devices. A parallel connection of five identical two-transistor flyback converters, with division of input voltage on a series of capacitors, has been proposed. The operation of converter, together with the developed control method, has been tested by performing simulation studies. In states of input voltage variations, the converter powers the load with stable output voltage.*

Keywords: DC-DC power converters, computer simulation, traction converters

PRZETWORNICA NAPIĘCIA STAŁEGO 3 KV NA NAPIĘCIE STAŁE 300 V MAŁEJ MOCY – ANALIZA SYMULACYJNA

Streszczenie. *W artykule zaprezentowano przetwornicę o mocy 150 W, zasilaną z kolejowej sieci trakcyjnej 3 kV DC, obniżającą napięcie do poziomu 300 V. Opisana struktura umożliwia zastosowanie tanich tranzystorów IGBT do budowy przetwornicy. Układ oparty jest na równoległym połączeniu pięciu dwu tranzystorowych przetwornic typu flyback, z podziałem napięcia wejściowego na kondensatorach. Praca przekształtnika, wraz z opracowaną metodą sterowania, została sprawdzona poprzez badania symulacyjne. W sytuacjach zmian napięcia wejściowego, układ zapewnia stabilne zasilanie obciążenia.*

Słowa kluczowe: przekształtniki DC/DC, symulacja komputerowa, przetwornice trakcyjne

Introduction

Automated and fully supervised railway routes require specialized sensors and actuators situated along the tracks to control train traffic. Most of these utilities are low power electrical devices that need power supply. The industrial AC power networks are very often far away from railway routes, therefore the necessary long power cable connections would be expensive. A reasonable source of power for such applications is the railway 3 kV DC traction system. The high input voltage needs to be decreased with the help of a DC/DC converter. High voltage converters suitable for low power applications are not common on the market. One of the main obstacles in wide application of such converters is related to the quality of the energy available in this way. The voltage variation in the railway grid, which is allowed by standards [4], can be of quite large range. It is evident, that in such systems the reliability of the power supply is of great importance to security of trains. This issue is emphasized by a high probability of overvoltages, increasing the already harsh insulation requirements for power switches. Therefore, high voltage, and thus more expensive, transistors are necessary. The aim of this work was to find a converter topology enabling the use of low cost IGBT transistors in 3 kV low power applications. In this paper a parallel connection of five identical low power "two transistor flyback converters with division of input voltage" has been proposed. The designed system has a modular form. The high voltage 3 kV DC is divided on an active input circuit, then further lowered by the work of the flyback converter modules. The operation of converter, together with the developed control method, has been tested by simulation studies using PSIM.

1. Converter structure

Insulation requirements of power electronic devices designed for railway traction grid applications can be met by the use of high voltage components, which are expensive. Another solution would be to find a topology reducing the voltage stress of power switches to a level enabling the selection of low cost IGBT transistors.

The described system enables power supply from railway traction grid for electronic devices situated along the railways, by lowering the input 3 kV voltage to the level of 300 V. The proposed 150 W system structure has a modular form, consisting of five low power DC/DC converters (modules), and is presented in Fig. 1.

In the introduced solution the high input traction voltage U_{RTV} is divided by a series of capacitors C_{xm}. In this way, the voltage requirements for single module components are lowered.

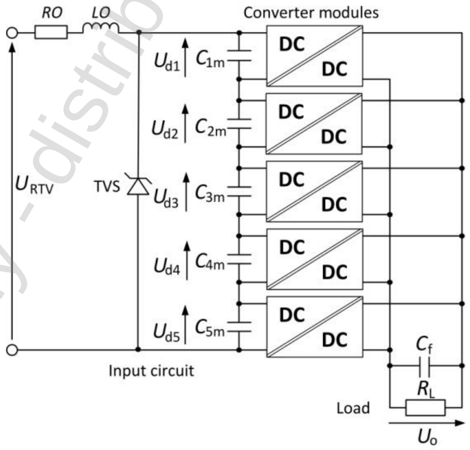

Fig. 1. Simplified block diagram of modular converter structure

The output pins of the modules are parallel connected, which can give a number of advantages [7]. The effective output voltage frequency is higher. As a result the load current pulsations are decreased. What is more, converter reliability is improved, and the power ratings of necessary components can be reduced. Assuming the use of 1200 V transistors, the number of necessary capacitors k, and thus converter modules, was obtained from:

$$k = \frac{U_{\max 3}}{1200} \cong 5 \tag{1}$$

where: $U_{\max3}$ – the highest long term overvoltage according to standard [4].

Devices supplied from the traction grid are exposed to overvoltages. In order to protect the load, a special input circuit was developed, which is a combination of a low-pass filter with TVS (*Transient Voltage Suppressor*) diodes. This circuit is incorporated in the structure presented in Fig. 1, and the parameters are listed in Table 1. The values were obtained through simulation studies of the response of the proposed circuit to two surge models defined in standards [5, 6]. The first surge waveform was *long voltage surge* [6], and the second one was *normalized voltage surge* 1.2/50 µs of 12 kV amplitude [5]. The complete design procedure is presented in [2].

Table 1. Parameter list of designed input protection circuit

RO	1100 Ω
LO	0.1 H
$C_{1m} – C_{5m}$	0.68 µF each
TVS	15 piece: 1.5 kW, V_{BR} = 400 V
U_{imax}	7165 V

U_{imax} is the clamped voltage of a series of 15 TVS diodes dissipating the overvoltage energy. This value is the maximum input voltage for the designed converter. Note that capacitors $C_{1m} – C_{5m}$ have two functions: input voltage division and transient voltage protection. From the value of maximum input circuit voltage U_{imax} the maximum single module input voltage can be determined:

$$U_{d\max} = \frac{U_{i\max}}{k} = 1433\ \text{V} \tag{2}$$

The module structure has influence on voltage stress of the power switches. The chosen two-transistor flyback converter, presented in Fig. 2, enables the reduction of switch voltage to the level equal the input voltage U_d [3]. The incorporated transformer provides galvanic isolation of the load.

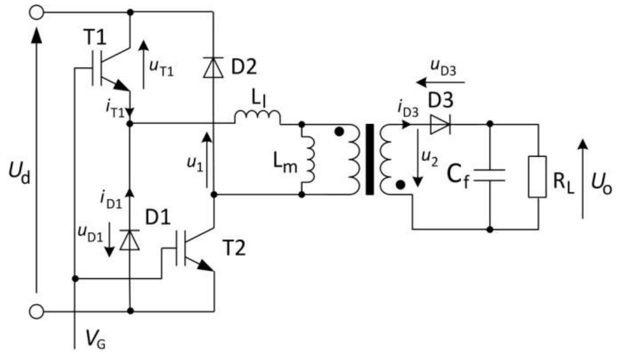

Fig. 2. Single module DC/DC structure: two-transistor flyback converter

2. Control principle

The module transistors T1 and T2 are simultaneously controlled. All modules work with the same switching frequency. Fast transistor switching enables low output voltage ripple, and the transformer core volume can be reduced. On the other hand, with higher frequencies, the impact of skin effect in transformer windings becomes more of a problem. For the power switches, IGBT transistors were chosen, so the maximum frequency is limited to about 20 kHz. Considering the above, the flyback module switching frequency equals:

$$f_m = 15\,\text{kHz} \tag{3}$$

Thanks to parallel connection of outputs of five modules, the output current frequency is increased to a maximum level of:

$$f_p = f_m \cdot k = 75\,\text{kHz} \tag{4}$$

In the selected flyback module structure the energy transfer process is divided in two stages. When the transistors are on, their currents are linearly increasing, and the energy is stored in transformer core. The load voltage U_o is not affected, until the transistors are turned off. The control system includes load voltage feedback. Therefore, the response of converter can be delayed. The process of charging the transformer magnetizing inductance has to be obligatory stopped, to avoid transformer saturation or even the damage of power switches. Another reason to limit the duty cycle of each modules transistors is not to excessively discharge the input circuit capacitors and preserve their voltage balance.

All DC/DC modules work with the same switching frequency, but control signals are phase shifted. The modules work in different time intervals, which are predetermined. The maximum transistor duty cycle can be obtained from:

$$D_{\max} = \frac{1}{k} = 0.2 \tag{5}$$

The proposed converter control system has been presented in form of a block diagram in Fig. 3. As a result of subtraction of measured load voltage U_o from the reference U_{ref} = 300 V, the value of error signal ε is calculated, and sent to the input of PI regulator. The output U_{PI} voltage is compared with sawtooth voltage signal U_{saw} of frequency f_m. The result V_{PWM} signal is further divided to form the right module control signals $V_{G1} – V_{G5}$ of maximum duty cycle D_{max} each. The input voltage U_d has influence on the increase rate of energy stored in transformer core. Module input voltages $U_{d1} – U_{d5}$ are measured, and depending on their voltage levels, D_{max} can be additionally limited, according to values presented in Table 2. In this way, the quality of output converter voltage U_o is improved, what will be discussed later. The division process of V_{PWM} signal is presented in Fig. 4.

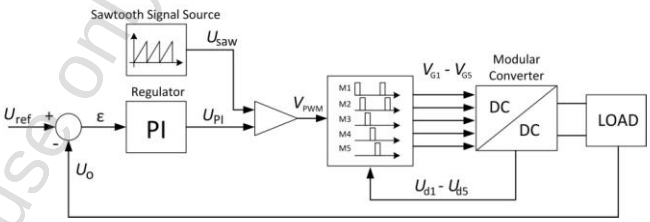

Fig. 3. Converter control system block diagram

Table 2. Limitation of duty cycle D according to module input voltages

U_{dx} [V]	D_{max} [-]
< 500	0.2
500 - 800	0.165
> 800	0.11

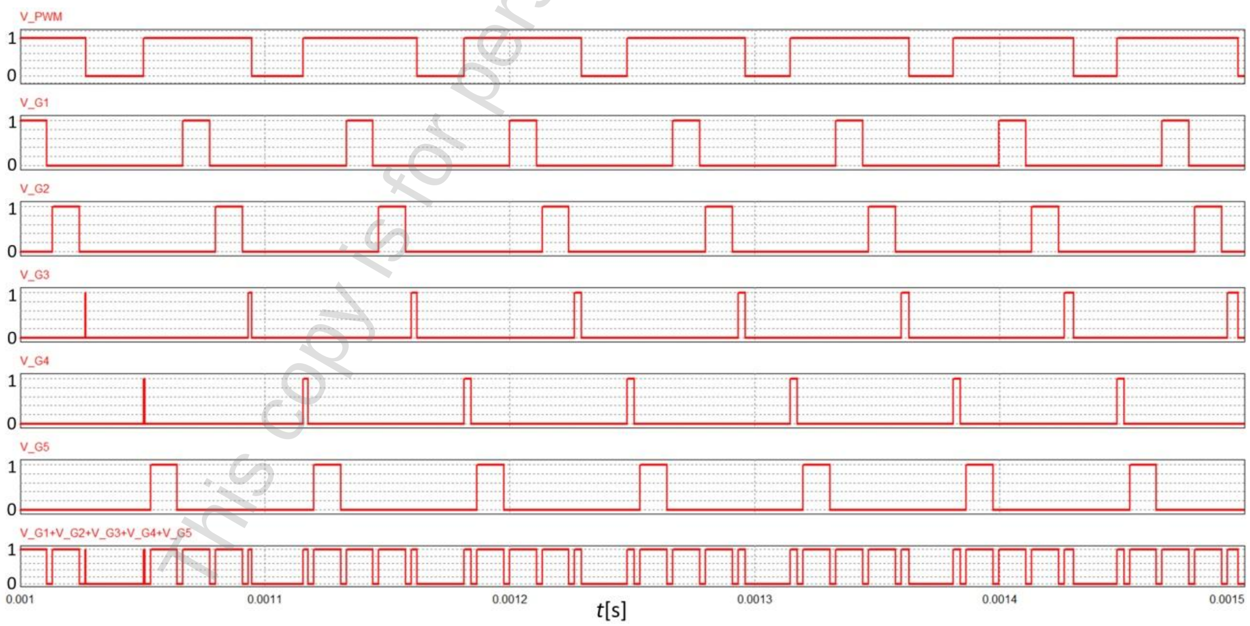

Fig. 4. VPWM signal division process

The last waveform from Fig. 4 presents the reconstruction of the main control signal as a sum of module gate drive signals. As can be seen, dead times occur. These are periods of time, when none of the modules is working. Note that in the 15 kHz cycle not all modules are obliged to operate. The interval in which each module can work is fixed, and the control gate drive signal sequence is: $V_{G1} - V_{G5}$. For example, in the last part of the cycle, the fifth module is allowed to work, but if at that moment the V_{PWM} signal is 0, there is no need for it. In this way, the output voltage frequency varies in time, which can be considered as a drawback.

3. Single module operation

For simulation purposes and analytical description a simplified circuit model of the two-switch flyback converter has been adopted [1] and presented in Fig. 2.

The power supply is represented by a DC voltage U_d. Transistors T1 and T2 are ideal two-state switches. The pulse transformer has been presented as an ideal transformer, with turns ratio defined as:

$$n_T = \frac{N_p}{N_s} > 1 \tag{6}$$

where: N_p – number of primary windings, N_s – number of secondary windings. The leakage L_l and magnetizing L_m inductances are connected in series and parallel to the primary winding respectively. The diodes denoted by D1, D2, D3 are also ideal elements. R_L is the load resistance and C_f is the output filter capacitance.

The processes occurring during DCM (*discontinuous current mode*) operation of two-transistor flyback converter, in a simplified analysis, can be divided into 4 main stages, and the typical waveforms characterizing work of converter are shown in Fig. 5. The circuit elements taking part in each switching cycle stage, are presented bold in Fig. 6. The second stage has been exaggerated for a better understanding of converter operation. The capacitance of load filter C_f is very large, so that the load voltage U_o ripples can be neglected.

STAGE 1 ($t_0 \le t < t_1$):
At the time of t_0, the switches T1 and T2 are turned on by the gate drive signal V_G. The voltage drops across diodes D1 and D2 equal $u_{D1} = u_{D2} = U_d$. Therefore, they are reverse biased and the currents $i_{D1} = i_{D2} = 0$. Assuming $L_l << L_m$, the voltage across the magnetizing inductance and the primary winding u_1 is U_d, then the voltage drop of diode D3 can be obtained from:

$$u_{D3} = \frac{U_d}{n_T} + U_o \tag{7}$$

so D3 is reversed biased, and the current i_{D3} is 0. The current flowing through transistors T1, T2, L_l and L_m is linearly increasing:

$$i_{T1} = i_{T2} = i_{L_m} = i_{L_l} = \frac{U_d}{L_m + L_l}(t - t_0) + i_{L_m}(t_0) \tag{8}$$

In this stage, the energy drawn from the power supply is stored in the magnetizing inductance L_m of transformer. The maximum value of transistor current equals:

$$i_{T1}(t_1) = i_{T2}(t_1) = \frac{U_d \cdot D \cdot T}{L_m + L_l} = \frac{U_d \cdot D}{f_m(L_m + L_l)} \tag{9}$$

where: T – period of the transistor switching cycle.

The voltages of switches T1 and T2 are 0. This stage ends at the moment of t_1, when gate drive signal V_G is 0, and the transistors are turned off.

STAGE 2 ($t_1 \le t < t_2$):
Switches T1 and T2 are turned off, diodes D1, D2 are forward biased. From t_1, when the transistors are turned off, energy stored in leakage inductance L_l is transferred through the conducting diodes D1 and D2 to the power supply. The current linearly decreases, with a slope:

$$\begin{aligned} i_{D1} = i_{D2} = i_{L_l} &= -\frac{u_{L_l}}{L_l}(t - t_1) + i_{L_m}(t_1) = \\ &= -\frac{U_d - u_{L_m}}{L_l}(t - t_1) + i_{L_m}(t_1) \end{aligned} \tag{10}$$

The voltage across transistors T1 and T2 is reduced to the value of U_d. After they are turned off at t_1, in order for the transformer flux to remain continuous, the primary winding voltage becomes negative. The same goes for the secondary winding voltage, so that D3 becomes forward biased. The energy stored in transformer core is transferred to the load. The D3 current linearly rises until the energy transfer from leakage inductance to the power supply is over at t_2.

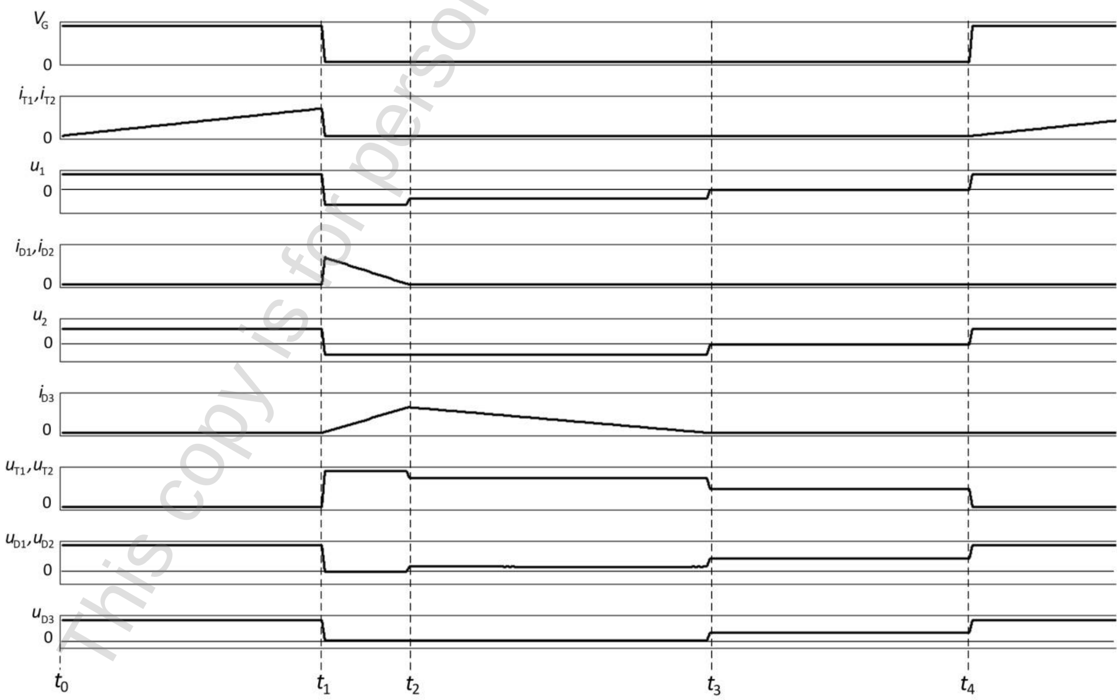

Fig. 5. Waveforms characterizing the work of two transistor flyback converter during DCM operation

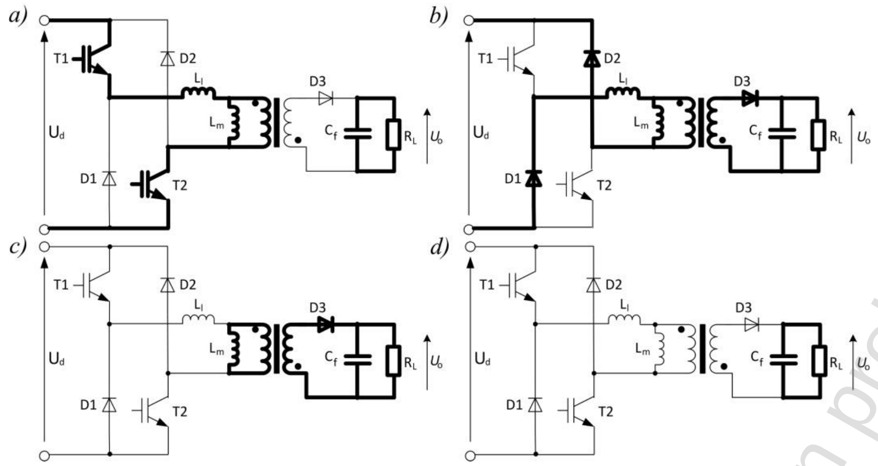

Fig. 6. Equivalent circuits representing two-transistor flyback switching cycle work stages: a) stage: 1 (t_0 – t_1); b) stage: 2 (t_1 – t_2); c) stage: 3 (t_2 – t_3); d) stage: 4 (t_3 – t_4)

STAGE 3 ($t_2 \le t < t_3$):
Transistors T1 and T2 are turned off. The currents i_{D1}, i_{D2} are 0. D3 is forward biased and the secondary winding current linearly decreases:

$$i_{D3} = -\frac{n_T^2 \cdot U_o}{L_m}(t - t_2) + n_T \cdot i_{L_m}(t_2) \tag{11}$$

Assuming switches T1 and T2 are identical, the voltages across those elements equal:

$$u_{T1} = u_{T2} = \frac{n_T \cdot U_o + U_d}{2} \tag{12}$$

Assuming diodes D1 and D2 are identical, their voltage drops equal:

$$u_{D1} = u_{D2} = \frac{U_d - n_T \cdot U_o}{2} \tag{13}$$

This work stage ends at t_3, when i_{D3} drops to 0, which means the energy stored in transformer core was completely extracted to the load.

STAGE 4 ($t_3 \le t < t_4$):
Transistors T1 and T2 are turned off. Diodes D1, D2, D3 are reverse biased, and their currents are 0. The voltages on transformer windings are 0. The voltages across T1, T2, D1, D2 equal:

$$u_{T1} = u_{T2} = u_{D1} = u_{D2} \cong \frac{U_d}{2} \tag{14}$$

The voltage of D3 equals the load voltage U_o. Energy stored in electric field of capacitor C_f is transferred to the load. The stage ends at t_4, when transistor gate drive signal V_G becomes high, and the cycle of operation starts anew.

4. Simulation results

The two-transistor flyback model presented in Fig. 2 has been tested in PSIM, to determine the maximum values of currents and voltages necessary for further component selection.

During simulation the input voltage U_d was equal to the calculated (2) maximum U_{dmax} value. Converter module parameters are listed in Table 3. The duty cycle D remained constant and was 0,2. The output filter capacitance C_f was very high, to keep the load voltage on a stable 300 V level. Simulation results are presented in form of waveforms depicted in Fig. 7.

Table 3. Converter module simulation parameters

Description	Symbol	Value
Transformer magnetizing inductance	L_m	6.5 mH
Transformer leakage inductance	L_l	0.03 mH
Transformer turns ratio	n_T	1.2
Module switching frequency	f_m	15 kHz
Load resistance	R_L	600 Ω

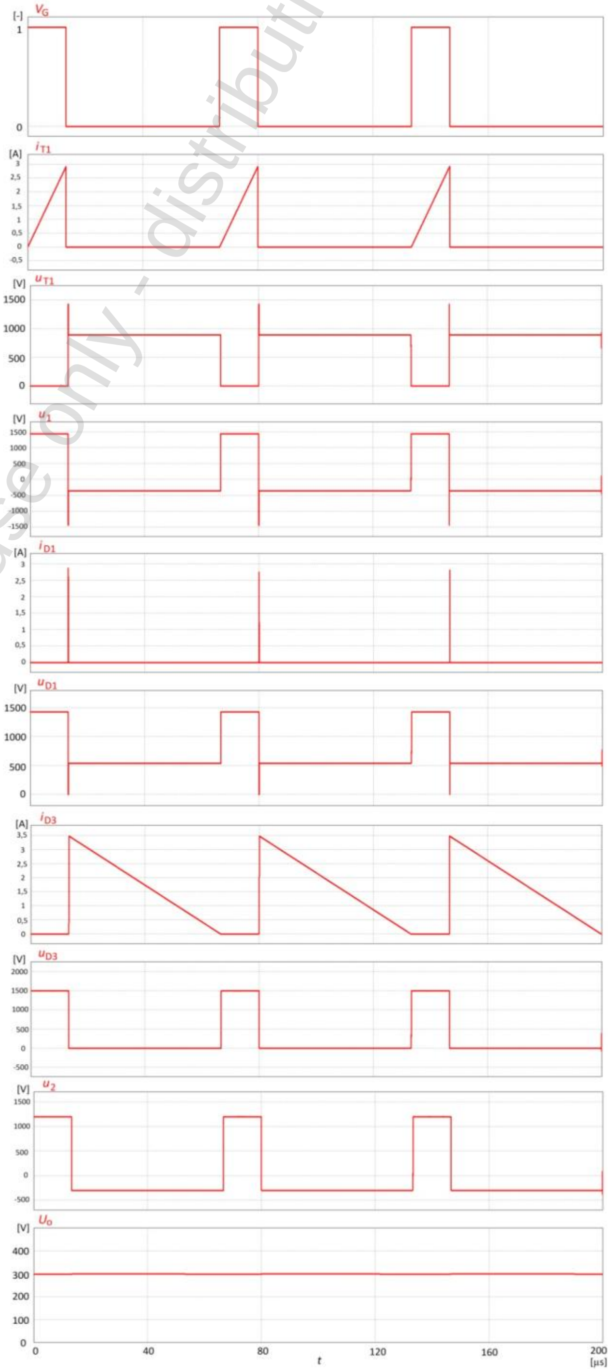

Fig. 7. Simulation results of two-transistor flyback module operation, U_d = 1433 V

Characteristic current and voltage values of converter elements are listed in Table 4 and Table 5. For comparison purposes, results for nominal module input voltage 600 V are also presented.

Thanks to the two-transistor flyback structure, the maximum collector-emitter voltage $U_{CE(max)}$ is clamped to the module input voltage value U_d. Due to the possibility of overvoltages, simulation results show, that for this system structure, 1700 V transistors are required.

Table 4. Converter module simulation results

		T1, T2				D1, D2			
U_d [V]	D [-]	$U_{CE(max)}$ [V]	$I_{C(max)}$ [A]	$I_{C(AV)}$ [A]	$I_{C(RMS)}$ [A]	U_{RRM} [V]	$I_{F(max)}$ [A]	$I_{F(AV)}$ [A]	$I_{F(RMS)}$ [A]
600	0,2	600	1.23	0.12	0.316	600	1.21	0.001	0.033
1433	0,2	1433	2.93	0.29	0.755	1433	2.93	0.002	0.058

Table 5. Converter module simulation results

		D3			
U_d [V]	D [-]	U_{RRM} [V]	$I_{F(max)}$ [A]	$I_{F(AV)}$ [A]	$I_{F(RMS)}$ [A]
600	0.2	802	1.448	0.24	0.484
1433	0.2	1502	3.477	1.38	1.79

To demonstrate the performance of proposed modular 3 kV/300 V converter, a full system simulation model was developed and presented in Fig. 8. The structure incorporates five modules presented in Fig. 2. The control scheme presented in Fig. 3 is contained in the "CONTROL" block. System parameters are listed in Table 3. The output filter capacitance C_f was 5 μF. The operation of the converter was tested by examining the system response to step changes of input voltage U_i. The range of variation was 2 kV – 7.2 kV.

For a better presentation of the influence of limitation of module duty cycle due to input voltage feedback, at first the simulation was conducted without this additional feature. D_{max} was fixed and equalled 0.2. Waveforms of converter input voltage U_i, currents of secondary winding module diodes $i_{D3\_1} - i_{D3\_5}$, and load voltage U_o are presented in Fig. 9a. Simulation results of converter system including input voltage feedback and limitation of module duty cycle according to values listed in Table 2 are presented in Fig. 9b. All system parameters remained the same.

Comparing load voltage waveforms U_o from Fig. 9a and Fig. 9b, it is clear, that thanks to maximum duty cycle limitation according to input voltage feedback, load voltage ripples have been reduced. In conditions of input voltage step change, the converter response is faster. Comparing the module secondary winding diode currents $i_{D3\_1} - i_{D3\_5}$, the currents from Fig. 9b reach lower peak values, and are more evenly distributed in time. Their instantaneous values are not summed. Transistors of each module conduct more frequently. For this reason the quality of regulation has improved, and the load voltage frequency is higher.

5. Conclusion

In this paper, a DC/DC 3 kV/300 V low power converter with input voltage division circuit has been proposed. The converter structure has a modular form, consisting of five two-transistor flyback modules. The operation of converter, together with the developed control method, has been tested by simulation studies using PSIM. Results of research showed proper functionality of the converter, and generation of stable output voltage during traction voltage variations. The additional limitation of module maximum duty cycle improved the regulation process. The level of output voltage ripples didn't exceed 5 V. Thanks to the input circuit and converter topology, the converter itself can be built using low cost 1700 V transistors.

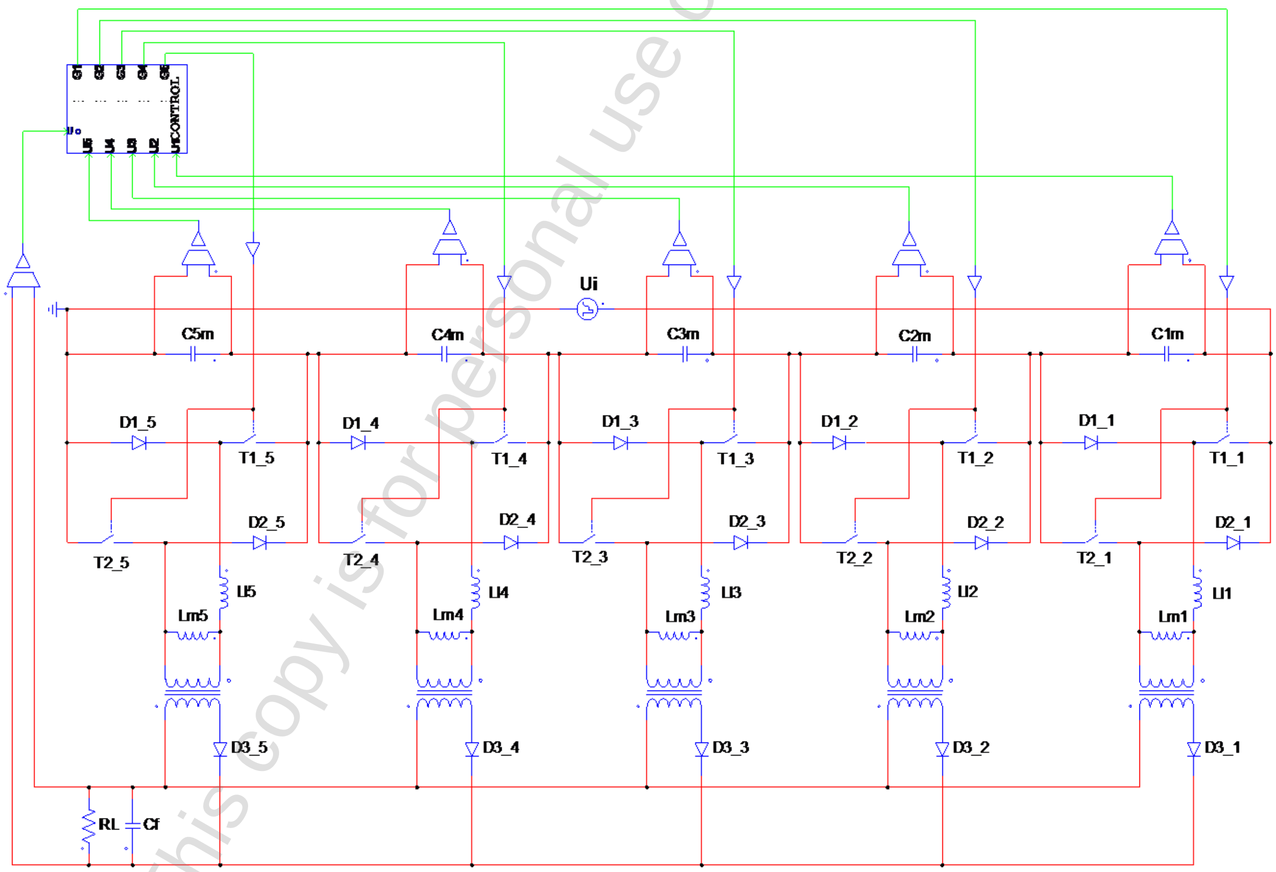

Fig. 8. Simulation model of 3 kV/ 300 V DC/DC converter

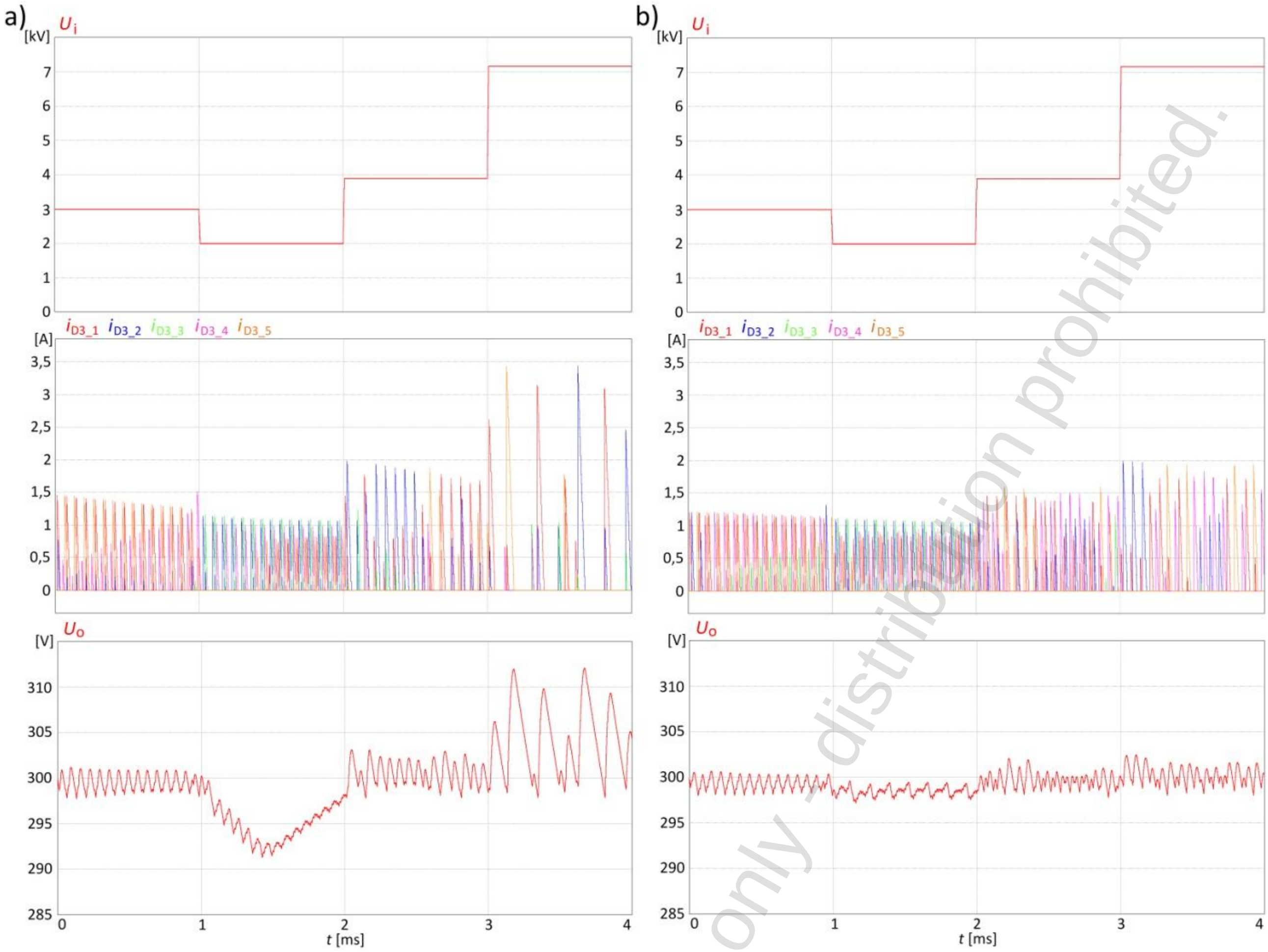

Fig. 9. Simulation results of 3 kV/300 V converter: a) without input voltage feedback; b) including input voltage feedback

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