A Dual Full-Bridge Resonant Class-E Bidirectional DC–DC Converter

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***Abstract*—A new bidirectional dc–dc converter composed of two class-E resonant converters is presented in this paper. Bidirectional power flow is controlled by transistor control pulse frequency changes, with a constant break between the succeeding pulses as in quasi-resonant converters. The boost or buck mode converter operation depends on the mutual relation between the control pulses of the transistor pairs which are located diagonally in the converter bridge. The advantages of the system involve are high operation frequency (200 and 450 kHz) and zero value of the transistor switching losses. This paper includes an analytical description that is useful for the converter design. The investigations are confirmed by PSpice simulations and an experiment of a laboratory model (1.6 kW).**

***Index Terms*—Bidirectional, bridge, class-E, dc–dc, resonant.**

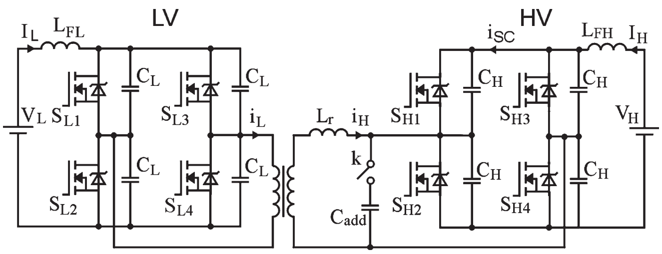

Fig. 1. Proposed BDC topology.

I. Introduction

In recent years, due to the requirements of electric vehicles [1], [2], uninterruptible power supplies [3], [4], and energy storage systems [5], bidirectional dc–dc converters (BDCs) have aroused much interest. Literature provides many solutions in this area, the main of which can be classified into several characteristic types.

The first type is, named as, a dual active-bridge (DAB) converter [6]. The main drawback of this solution is that the converter cannot achieve zero-voltage switching (ZVS) in a wide range of load variations while input or output voltage rises. In order to eliminate this problem in control system, the phase shift was additionally enhanced with a pulsewidth modulation [7], [8].

A three-port active bridge (TAB) was introduced as an extension of the DAB topology [9].

Another type of BDC is characterized by a current-fed inverter/rectifier on the low-voltage (LV) side of the transformer and a voltage-fed inverter/rectifier on the high-voltage (HV) side.

Some of the BDC characteristics are common for other types of converters [10]. For low-power applications, simplified configurations based on elementary converters are proposed [11], [12]. A quasi-resonant bidirectional converter seems to be an attractive solution here [12]. Owing to ZVS or zero-current switching, it can operate with higher frequency. In addition, it is characterized by a low level of electromagnetic noise. The main drawback of this proposition is the necessity of transfer of different power levels in both directions when the same resonant circuit elements are used. In this paper, a completely new type of BDC is discussed. It is composed of current-sourced resonant class-E inverter/rectifier converters. The class-E amplifier/converter was first proposed in [14]. The theoretical background of high-frequency resonant class-E inverters/rectifiers is developed and presented in [15]. Even though the converters have had many different applications so far, there has been no attempt to use them for bidirectional power flow with a high range of control.

The LV-side converter is controlled as a boost converter [16] while the HV-side converter is controlled as a buck converter [17]. The advantages of this proposition are similar as in quasi-resonant BDCs [12]. To ensure a comparable level of power transfer for both directions, the resonant circuit has different capacitors for each direction. For that reason, an additional capacitor C_{add} is switched on or not by the switch k. The switching depends on the change in the direction of power flow. Depending on the system application (switching frequency), a mechanical switch or a bidirectional thyristor or transistor switch can be used as the switch k.

II. Proposed BDC Scheme and Operation Principle

The class-E resonant bridge BDC is shown in Fig. 1. It consists of two bridge inverters fed by current sources due to the input inductances L_{FL} and L_{FH}.

Inverter outputs are connected by a transformer, which is characterized by secondary-to-primary turns ratio k_T.

During the power transfer from the LV source V_L to the HV source V_H, an additional capacitor C_{add} is introduced by closing the key k. The LV converter transistors are controlled and the converter operates as a class-E boost converter [16], while the HV converter transistors are not controlled and the

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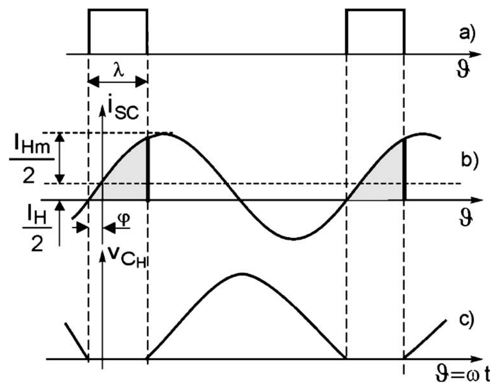


Fig. 2. Waveforms for the optimal operation point with the minimal value of the transferred power.

converter operates as a class-E rectifier [15] composed of transistor body diodes. The resonant circuit $[(C_{\mathrm{add}} + 2C_H)L_r]$ is formed by the inductance L_r, capacitance C_{add} (switch k is closed), and capacitances C_H which are placed parallel to the diode.

During power flow in the opposite direction, from the (V_H) source to the (V_L) source, the HV converter transistors are controlled and the converter operates as a class-E buck converter [17], while the LV converter transistors are not controlled and the converter operates as a class-E rectifier composed of transistor body diodes. The switch k is opened, and the resonant circuit is formed by the capacitances C_L and the inductance L_r.

To ensure ZVS in a class-E resonant converter, it has to be controlled by frequency change, keeping a constant break between the succeeding control pulses of the transistor, as it is the case with quasi-resonant converters. The LV-boost and HV-buck mode operation is characterized by an overlap or a break of the control pulse of the transistor pairs which are located diagonally in the bridge.

III. Theoretical Background of the Converter Operation

The properly designed converter assumes the optimal operation point with maximal transferred power for the LV-boost direction and with minimal transferred power for the HV-buck direction. The main task of the designer is to calculate the adequate capacitance values of the capacitors C_L and C_H to ensure the optimal operation point.

A. HV-Buck Operation Mode

The waveform characteristic for the optimal operation point with minimal transferred power (maximal control frequency and minimal control pulsewidth λ) is shown in Fig. 2.

We assume that the converter input current I_H is constant in shape and the converter output current i_H is sinusoidal in shape. Owing to the converter configuration symmetry, its branch current i_{SC} can be described as

$$i_{SC} = \frac{I_{Hm}}{2} \sin \omega t + \frac{I_H}{2} \tag{1}$$

where I_{Hm} is the HV converter output current amplitude and ω is the converter control pulsation.

The part of current i_{SC} marked in Fig. 2 with oblique strokes flows through the transistor, and the remaining unmarked part of the current flows through the capacitor C_H, first charging it from zero voltage and then, when the current changes direction, discharging to zero voltage. Therefore, the following equation is fulfilled:

$$\int_{\lambda-\varphi}^{2\pi-\varphi} \left(\frac{I_{Hm}}{2} \sin \omega t + \frac{I_H}{2} \right) d(\omega t) = 0 \tag{2}$$

where λ is the transistor control pulsewidth and φ is the initial phase of the branch current.

The solution to the aforementioned equation yields the dependence of the relation between the converter input I_H and output I_{Hm} currents on the control pulsewidth

$$\frac{I_{Hm}}{I_H} = \sqrt{1 + \left(\frac{\sin \lambda + \pi \left(2 - \frac{\lambda}{\pi} \right)}{1 - \cos \lambda} \right)^2}. \tag{3}$$

The desired value of the capacitance C_H can be evaluated, taking into account that, during the steady-state operation, the mean value of the capacitor C_H voltage is equal to the half value of the source voltage V_H

$$\frac{V_H}{2} = \frac{1}{2\pi} \int_{\lambda-\varphi}^{2\pi-\varphi} \nu_{C_H} \, d(\omega t) \tag{4}$$

where the capacitor voltage ν_{C_H} can be calculated as

$$\nu_{C_H} = \frac{1}{\omega C_H} \int_{\lambda-\varphi}^{\omega t} \left(\frac{I_{Hm}}{2} \sin \omega t + \frac{I_H}{2} \right) d(\omega t). \tag{5}$$

The solutions to (4) and (5) yield the value of capacitance C_H

$$C_H = 0.0254 \frac{P_H}{f V_H^2} (2\pi - \lambda) \left(\frac{\cos(\lambda - \varphi)}{\sin \varphi} - \lambda + \varphi \right) + 0.0254 \frac{P_H}{f V_H^2} \left[\frac{1}{2} (2\pi - \lambda)^2 + \frac{\sin \lambda}{\sin \varphi} \right]. \tag{6}$$

The capacitance value depends on the transferred power P_H, the control frequency f, and the control pulsewidth λ.

B. Boost Operation Mode

A similar analysis for the buck converter should be performed to evaluate the capacitance C_L value for the boost converter. Such an analysis was presented in detail in [16]; therefore, we now refer to the equations deduced in the article. The relation of the input I_L to the output I_{Lm} current of the boost converter is given by

$$\frac{I_L}{I_{Lm}} = \frac{1 + \cos \left(\frac{2t_{\mathrm{ov}}}{T} \pi \right)}{\pi \left(1 - \frac{2t_{\mathrm{ov}}}{T} \right)} \tag{7}$$

where t_{ov} is the time overlap of the transistor control pulses.

The capacitance C_L value is given by

$$C_L = 8.866 \cdot 10^{-6} \frac{T I_{Lm}}{V_L}. \tag{8}$$

Using (7), the last equation can be transformed to the form

$$C_L = 0.00887 \frac{P_L}{f V_L^2} \frac{\pi \left(1 - \frac{t_{\text{ov}}}{T}\right)}{1 + \cos\left(\frac{2t_{\text{ov}}}{T}\pi\right)}. \tag{9}$$

The capacitance value of the parallel placed capacitor depends on the transferred power P_L, the control frequency f, and the transistor control pulse overlap time t_{ov}.

IV. Design Example and PSpice Simulations

The design example is calculated for following specifications: LV source $V_L = 50$ V, HV source $V_H = 300$ V, a controlled power range of 0.3–1.6 kW, and a control frequency range of 200–450 kHz.

The results of the evaluation of the converter elements are as follows: the capacitor shunting transistors $C_L = 40$ nF and $C_H = 3.5$ nF, the additional capacitor of the resonant circuit $C_{\text{add}} = 8$ nF, and the resonant circuit inductance $L_r = 35\ \mu$H.

The observable oscillations on the transistor current waveforms are caused by a parasitic inductance. Generally, the voltage and current waveforms obtained in the experiment are in accordance with the waveforms obtained during simulations.

The BDC system efficiency versus the power value, measured for both directions of power flow, is shown in Fig. 5. Input filter inductances L_{FL} and L_{FH} were assumed as 0.3 and 1.3 mH, respectively. Generally, these inductances should be at least ten times bigger than the resonant inductance.

PSpice simulations have been done and validate the aforementioned analysis and derived equations. The power losses of the particular switches as a function of control frequency (result of simulation) are shown in Fig. 3. Obviously, LV converters are characterized by higher power losses in switches than HV converters. The important result of the simulation is the fact that the power losses in the HV converter are bigger when its switches operate as transistors (curve 3) than when they operate as diodes (curve 4). In the LV converter, the opposite situation takes place: Bigger losses in the switches occur when they operate as diodes (curve 1) than when they operate as transistors (curve 2).

Because of the ZVS, the losses are not significant for the system (their value is much smaller than the allowable catalogue value of the used transistors). What is more, the value decreases rapidly together with the increase in the value of the power being sent (the increase of control frequency).

V. Experimental Results

The laboratory prototype BDC was implemented using the following: *IRFP4227PbF transistors for the LV-boost converter and IXFN34N100 for the HV-buck converter, a transformer with toroidal core (Vitroperm 500F WAC T60004-L2130-W352) wound with a Litz wire* (3×6 mm^2 *primary and 4* mm^2 *secondary), input filters* L_{FL} *and* L_{FH} *wound with a Litz wire on ferrite core with air gap, propylene capacitors* $C_H =$ 3.5 nF,$C_L = 40$ nF, *and* $C_{\text{add}} = 10$ nF, *and the air resonant inductor wound with a 4-* mm^2 *Litz wire*.

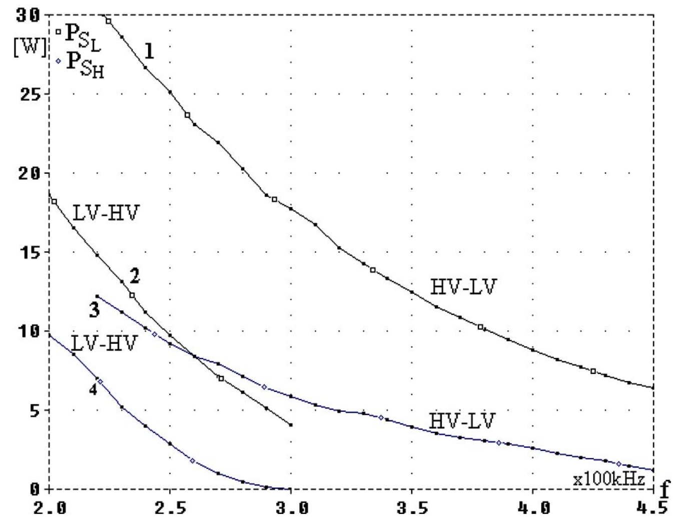


Fig. 3. Power losses of the converter switches as a function of control frequency. *Graphs 1 and 2 concern S_L transistors while graphs 3 and 4 concern S_H transistors, HV–LV power is transferred from V_H to V_L, and LV–HV power is transferred from V_L to V_H.*

The experimental waveforms are shown in Fig. 4.

The observable oscillations on the transistor current waveforms are caused by a parasitic inductance. Generally, the voltage and current waveforms obtained in the experiment are in accordance with the waveforms obtained during simulations.

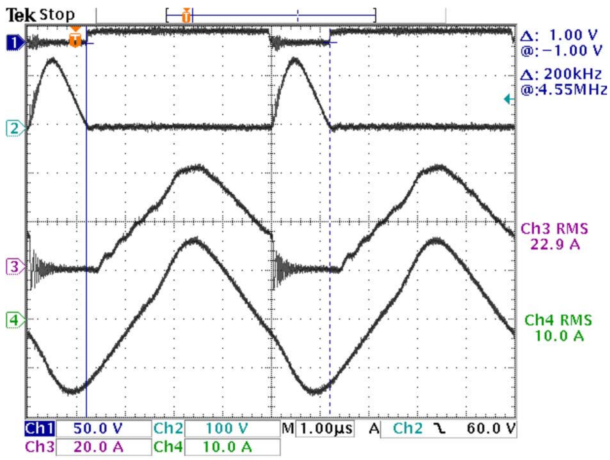
The BDC system efficiency versus the power value, measured for both directions of power flow, is shown in Fig. 5.

Observing the results presented in the figure, one has to notice that, in every solution, when the power being sent in the system decreases to zero, its efficiency also decreases to zero, due to the so-called losses of no-load mode (circulating currents). When using class-E converters, it is not uncommon to use even higher control frequencies (several megahertz). Power frequency over 200 kHz used in this paper was limited by the availability of isolated transistor drivers for bridge systems. Further work on the increase of frequency value is possible; however, the aim of this paper is to introduce and verify for the first time the idea of application of class-E converters for bidirectional power flow.

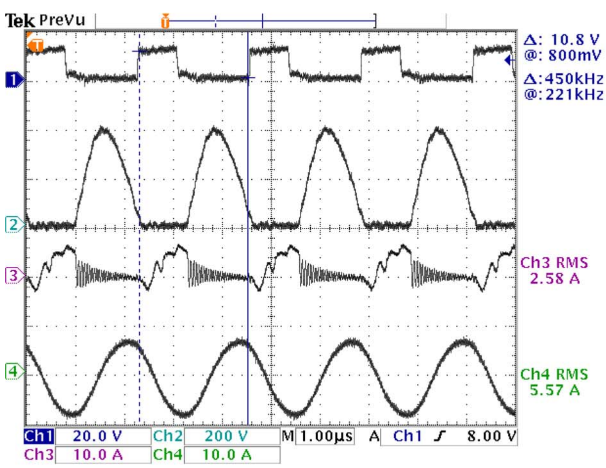
Although the converter power dissipation in the system was minimized, its efficiency remains limited because of transformer power dissipation. The problem of power dissipation in transformers working at high frequency is of more general character and requires additional investigation.

VI. Conclusion

A new BDC system based on the resonant class-E converters has been proposed in this paper. The basic advantage of this proposition is a relatively high operation frequency, which enables the decrease of the system's mass and cost. The high operation frequency guarantees also high dynamic properties. The proposed converter operates at a frequency of several hundreds of kilohertz, whereas the systems described in the



(a)



(b)

Fig. 4. Waveforms obtained for the optimal operation point. (a) LV-boost operation ($f = 200$ kHz). (b) HV-buck operation ($f = 450$ kHz) (experimental results). *(1) Transistor control pulses. (2) Transistor voltages (LV: 100 V/div; HV: 200 V/div). (3) Transistor currents (LV: 20 A/div; HV: 10 A/div). (4) Transformer high side current (10 A/div).*

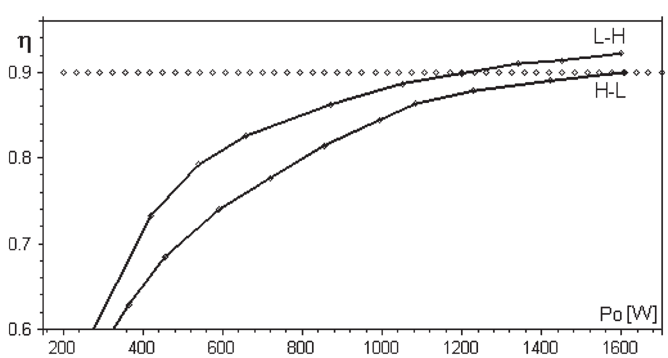


Fig. 5. BDC system efficiency as a function of the transferred power. LV–HV *is the power transfer from* V_l *to* V_H, and HV–LV *is the power transfer from* V_H *to* V_L.

literature operate at a switching frequency of several dozens of kilohertz. Both simulation and experimental results have proven that transistor power dissipation is small despite high control frequency. Both converters employed in the system are current sourced. Therefore, the parasitic inductances of connections are of no importance. System is characterized by a low level of electromagnetic noise, which makes it environmentally friendly. This nonlinear system is easily described basing on the fundamental current harmonic; thus, it is simple to design and evaluate. The system simulation results and the results of the laboratory model investigations confirm the correctness of the presented analytical description.

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