

An Analysis of a Bidirectional Series Resonant DC/DC Converter

Nikolay Dimitrov Bankov, Aleksandar Stoyanov Vuchev
and Stoyan Aleksandrov Vuchev

Abstract – A bidirectional series resonant DC/DC converter operating at frequencies higher than the resonant frequency is examined. It consists of two identical bridge commutation circuits whereby a phase-shift control is achieved. An analysis of the processes in the converter with the impact of only the first harmonic is carried out. The condition that defines the border between the two typical operating modes is established. As a result of the analysis, expressions for basic parameters of the converter are obtained and both output and control characteristics are drawn.

Keywords – Series Resonant DC/DC Converter, Controlled Rectifier

I. INTRODUCTION

For years, series resonant DC/DC converters operating at frequencies higher than the resonant frequency have been widely used for the realization of different power supply devices. This is due to their inherent advantages, the most significant of which are small size and weight, natural short circuit protection and so on. Last but not least, these converters operate at low switching losses due to the real opportunity the power switches to commute when the voltage applied on them has a value of zero (ZVS – Zero Voltage Switching).

In order to control the output power of the series resonant converters, two groups of methods are applied [1] – at variable and at constant frequency. Most of these methods are realized by the means of the inverter, which assumes uncontrolled rectifiers to be put in the converter structure.

Such a decision, however, does not support energy return to the power supply, which is indispensable for a number of applications. A solution to the problem is the use of controlled rectifier [1] – [3]. Such circuits for phase-shift control of series resonant DC/DC converters have long been known and investigated [4] – [6]. With these circuits, the output power varies practically from zero to the maximum value. In this case, however, the used rectifier is asymmetric, which does not provide energy return back to the power supply source.

N. Bankov is with the Department of Electrical Engineering and Electronics, Technical Faculty, University of Food Technologies - Sofia, 26 Maritza Blvd., 4000 Plovdiv, Bulgaria, e-mail: nikolay_bankov@yahoo.com

A. Vuchev is with the Department of Electrical Engineering and Electronics, Technical Faculty, University of Food Technologies - Sofia, 26 Maritza Blvd., 4000 Plovdiv, Bulgaria, e-mail: avuchev@yahoo.com

S. Vuchev is with the Department of Power Electronics, Faculty of Electronic Engineering and Technologies, Technical University of Sofia, Kl. Ohridski 8, 1000 Sofia, Bulgaria, e-mail: vu4ev_100yan@abv.bg

For investigation of series resonant DC/DC converters, a number of authors [5, 6] apply harmonic analysis taking into consideration only the influence of the first harmonic. The results obtained by this method are to a significant extent appropriate for engineering calculations.

A purpose of the present work is the examination of a bidirectional series resonant DC/DC converter with harmonic analysis with the impact of only the first harmonic. As a result of the analysis, expressions for basic parameters of the converter must be obtained and both its output and control characteristics must be drawn.

II. PRINCIPLE OF THE CONVERTER OPERATION

The scheme of the examined bidirectional converter is presented in fig.1. It consists of two similar bridge switching circuits, a resonant tank circuit (L , C), matching transformer Tr , capacitive input and output filters (C_F and C_0) and load resistor (R_0). Likewise the conventional series resonant DC/DC converter, the first switching circuit (transistors $Q_1 - Q_4$ with reverse diodes $D_1 - D_4$) interprets the function of an inverter, and the second one (transistors $Q_5 - Q_8$ with reverse diodes $D_5 - D_8$) respectively – the function of a rectifier. In the scheme, the required for ZVS mode snubber capacitors are not shown.

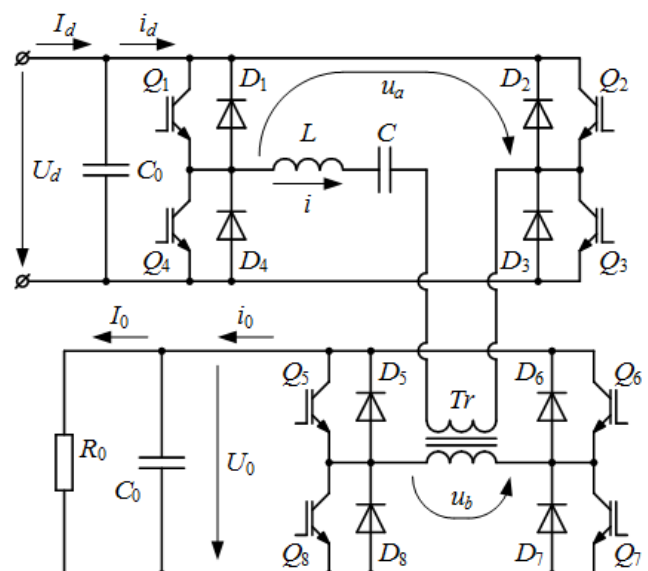


Fig. 1. Scheme of the bidirectional resonant DC/DC converter

The operation of the converter is illustrated by the waveforms, presented in fig.2. In this case, however, in order to control the output power, a different from the suggested in [4] – [6] strategy is applied.

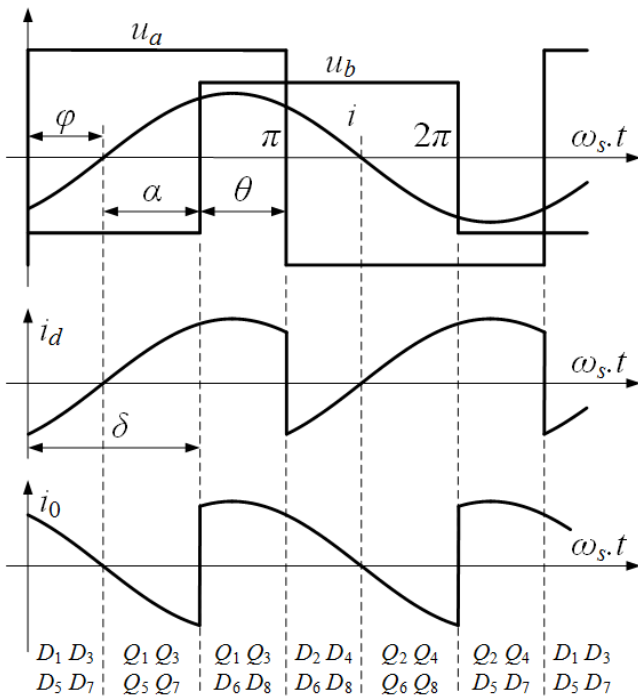


Fig. 2. Waveforms of the basic voltages and currents

The converter operates at constant frequency higher than that of the resonant circuit. Therefore, the inverter transistor couples (Q_1, Q_3 or Q_2, Q_4) switch at ZVS. They are controlled in a way that the voltage u_a at the inverter output has almost rectangular shape. The resonant current i falls behind the voltage u_a at angle φ . The rectifier transistors operate at ZVS also. Therefore, with the current i pass by zero, the respective couple (Q_5, Q_7 or Q_6, Q_8) starts to conduct. This couple is switched off after time corresponding to the angle δ after the turn-off of the inverter conducting transistors for the same half period. The voltage u_b at the input of the rectifier, the shape of which is similar to the one of u_a , changes its direction when the rectifier transistors are switched off. In this way, the control of the output power is obtained by the variation of the angle δ .

III. ANALYSIS OF THE CONVERTER OPERATION

The examination of the converter is carried out with harmonic analysis with the impact of only the first harmonic taken into account. The obtained results are suitable for engineering calculations.

For the purposes of the analysis, the following assumptions are made: the matching transformer is ideal with a transformation ratio k , all the circuit elements are ideal, the influence of the snubber capacitors and the pulsation of the supply voltage U_d and the output voltage U_0 are neglected, i.e. the voltages u_a and u_b have rectangular shape.

The analysis of the waveforms in fig.2 shows the possible energy transfer modes – from the power supply source to the load and back.

In accordance with the assumptions made, when the conduction time of the inverter reverse diodes D_1, D_3 (respectively D_2, D_4) is more than that of the transistors Q_1, Q_3 (respectively Q_2, Q_4), the energy transferred back to the

power supply source is more than the consumed one, so that the average value of the current I_d is negative. This happens when $\varphi > \pi/2$.

Because of the rectifier diodes operation, the output voltage U_0 cannot change its sign. Therefore, the energy transferred back from the load is more than the consumed for the rectifier when $\alpha > \pi/2$ or when the average value of the current I_0 becomes negative. This means that the rectifier transistors Q_5, Q_7 (respectively Q_6, Q_8) must conduct longer than its diodes D_5, D_7 (respectively D_6, D_8).

Considering the above mentioned and taking into consideration that $\delta = \alpha + \varphi$, energy transfer back to the supply source is possible when $\delta > \pi$.

On the basis of the assumptions made, the following symbols are introduced:

$$\omega_0 = 1/\sqrt{LC} \text{ and } \rho_0 = \sqrt{L/C} - \text{resonant frequency and characteristic impedance of the resonant tank circuit (L, C);}$$

$\nu = \omega_s/\omega_0$ – frequency detuning of the resonant circuit, where ω_s is the inverter operating frequency.

In accordance with the chosen method of analysis, it is assumed that only the first harmonics of the current i and the voltages u_a and u_b have impact in the examined circuit. This gives the opportunity the processes in the converter to be illustrated with vector charts. The first operation mode when $0 \leq \delta \leq \pi$ and the energy is transferred to the load is illustrated in fig.3a. The second mode when $\pi \leq \delta \leq 2\pi$ and the energy is transferred back from the load is illustrated with the vector chart in fig.3b.

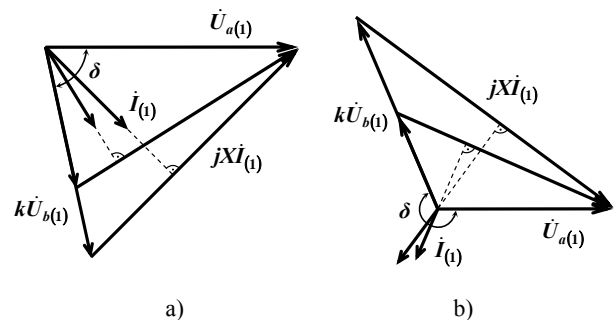


Fig. 3. Vector diagrams of the basic voltages and currents

The following equation is valid for the examined circuit in standing mode:

$$\dot{U}_{a(t)} = k\dot{U}_{b(t)} + jX\dot{I}_{(t)} \tag{1}$$

where $\dot{U}_{a(t)}$, $\dot{U}_{b(t)}$ and $\dot{I}_{(t)}$ are the first harmonic complexes of u_a, u_b and i , and

$$X = \omega_s L - \frac{1}{\omega_s C} = \left(\nu - \frac{1}{\nu} \right) \rho_0$$

According to the assumptions made, the average value of the output current is defined as:

$$\begin{aligned} I_0 &= \frac{1}{\pi} \int_0^\pi \sqrt{2} k I_{(t)} \sin(\omega t + \alpha) d\omega t = \\ &= \frac{2\sqrt{2}}{\pi} k I_{(t)} \cos \alpha \end{aligned} \tag{2}$$

from where for the effective value of the first harmonic of the resonant current is derived that:

$$I_{(1)} = \frac{\pi}{2\sqrt{2}} \cdot \frac{I_0}{k \cos \alpha} \quad (3)$$

In the first case when $0 \leq \alpha \leq \pi/2$, the following expressions are valid for the complex values $\dot{U}_{a(1)}$, $\dot{U}_{b(1)}$ and $jXI_{(1)}$:

$$\begin{aligned} \dot{U}_{a(1)} &= \frac{2\sqrt{2}}{\pi} U_d \\ \dot{U}_{b(1)} &= \frac{2\sqrt{2}}{\pi} U_0 \cdot e^{-j\delta} \\ jXI_{(1)} &= \frac{\pi}{2\sqrt{2}} \cdot \frac{I_0}{k \cos \alpha} \cdot \frac{v^2 - 1}{v/\rho_0} \cdot e^{j(\pi/2 - \varphi)} \end{aligned} \quad (4)$$

In the second case when $\pi/2 \leq \alpha \leq \pi$, the expressions for the complex values $\dot{U}_{a(1)}$, $\dot{U}_{b(1)}$ and $jXI_{(1)}$:

$$\begin{aligned} \dot{U}_{a(1)} &= \frac{2\sqrt{2}}{\pi} U_d \\ \dot{U}_{b(1)} &= \frac{2\sqrt{2}}{\pi} U_0 \cdot e^{j(2\pi - \delta)} = \frac{2\sqrt{2}}{\pi} U_0 \cdot e^{-j\delta} \\ jXI_{(1)} &= \frac{\pi}{2\sqrt{2}} \cdot \frac{I_0}{k \cos \alpha} \cdot \frac{v^2 - 1}{v/\rho_0} \cdot e^{j(\pi/2 - \varphi)} \end{aligned} \quad (5)$$

From expressions (4) and (5) it is observed that independently from the interval of variation of the control angle δ , equation (1) assumes the same description:

$$\begin{aligned} \frac{2\sqrt{2}}{\pi} U_d &= \frac{2\sqrt{2}}{\pi} k U_0 \cdot e^{-j\delta} + \\ &+ \frac{\pi}{2\sqrt{2}} \cdot \frac{I_0}{k \cos \alpha} \cdot \frac{v^2 - 1}{v/\rho_0} \cdot e^{j(\pi/2 - \varphi)} \end{aligned} \quad (6)$$

In order to obtain generalized results, the magnitudes output voltage U'_0 and output current I'_0 are introduced in relative units:

$$U'_0 = kU_0/U_d \text{ and } I'_0 = \frac{I_0/k}{U_d/\rho_0} \quad (7)$$

Then from equation (6) the following normalized expression is obtained:

$$\begin{aligned} 1 &= U'_0 \cos \delta - jU'_0 \sin \delta + \\ &+ \frac{\pi^2}{8} \cdot \frac{v^2 - 1}{v} \cdot \frac{I'_0}{\cos \alpha} \cdot \cos(\pi/2 - \varphi) + \\ &+ j \frac{\pi^2}{8} \cdot \frac{v^2 - 1}{v} \cdot \frac{I'_0}{\cos \alpha} \cdot \sin(\pi/2 - \varphi) \end{aligned} \quad (8)$$

As the imaginary part of the above expression equals zero, it is obtained that:

$$U'_0 \sin \delta = \frac{\pi^2 (v^2 - 1)}{8v} \cdot \frac{I'_0}{\cos \alpha} \cdot \cos \varphi \quad (9)$$

From the vector charts presented in fig.3 it is observed that:

$$U_{a(1)} \cos \varphi = kU_{b(1)} \cos \alpha \quad (10)$$

and therefore:

$$\cos \varphi = U'_0 \cos \alpha \quad (11)$$

On the basis of the expressions (9) и (11), the final expression is obtained as follows:

$$I'_0 = \frac{\sin \delta}{c} \quad (12)$$

$$\text{where: } c = \frac{\pi^2}{8} (v - 1/v)$$

From the real part of equation (8) and taking (12) into consideration, it is derived that:

$$U'_0 = \frac{1 - \frac{\sin \varphi}{\sin \alpha} \sin \delta}{\cos \delta} \quad (14)$$

The above expression allows the boundaries of the ZVS operating mode to be obtained for the transistors of the inverter (at $\varphi = 0$ and $\varphi = \pi$) and the rectifier (at $\alpha = 0$ and $\alpha = \pi$) respectively.

It is easy to see that these boundaries are meaningful at two cases when control angle is changing in range $0 \leq \delta \leq \pi/2$ or $3\pi/2 \leq \delta \leq 2\pi$ respectively. For $\pi/2 \leq \delta \leq 3\pi/2$ the transistors in both the rectifier and the inverter operate at ZVS.

IV. OUTPUT AND CONTROL CHARACTERISTICS

A full enough picture of the examined series resonant DC/DC converter features can be obtained via analysis of its graphically illustrated normalized characteristics drawn for a value of the transformation ratio equal to one ($k = 1$).

Thus, on the basis of equation (12), the normalized dependencies of the output voltage U'_0 from the output current I'_0 are obtained when the converter operates at a constant frequency ($v = 1,15$). As mentioned above, according to the control angle are two possible cases. For first one at $\delta = -\pi/2; -2\pi/5; -3\pi/10; -\pi/5; -\pi/10; \pi/10; \pi/5; 3\pi/10; 4\pi/5; \pi/2$ such dependencies are presented on fig.4.

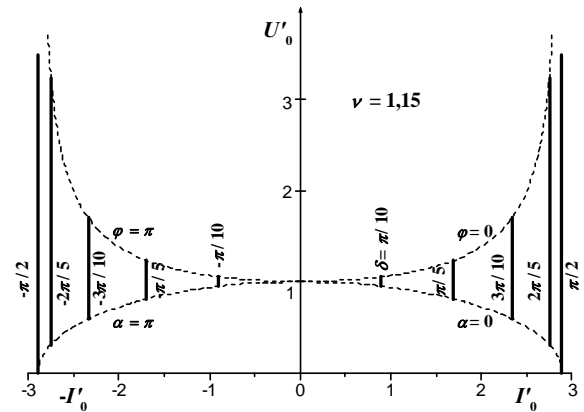


Fig. 4. Output characteristics at $-\pi/2 \leq \delta \leq \pi/2$

The output characteristics are presented with thick lines. The ones situated in the first quadrant for $0 \leq \delta \leq \pi/2$ correspond to the mode of energy transfer to the load, and the others in the second quadrant for $-\pi/2 \leq \delta \leq 0$ – to the mode of energy return back to the power source respectively. In the area of the output characteristics, the boundaries of ZVS mode are shown with dotted line for inverter's transistors (at $\varphi = 0$ and $\varphi = \pi$) and for rectifier's transistors (at $\alpha = 0$ and $\alpha = \pi$) respectively. The analysis of the dependencies presented in fig.4 shows that in this case the converter operation is very limited and practically no-load mode is not possible.

For control angle $\delta = \pi/2; 3\pi/5; 7\pi/10; 4\pi/5; 9\pi/10; \pi; 11\pi/10; 6\pi/5; 13\pi/10; 7\pi/5; 3\pi/2$ normalized dependencies of the output voltage U_0 from the output current I_0 are shown in fig.5.

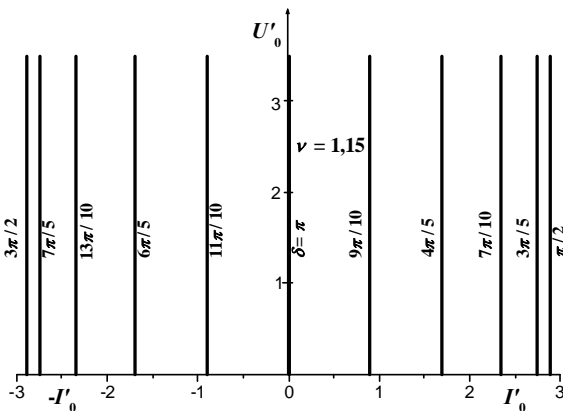


Fig. 5. Output characteristics at $\pi/2 \leq \delta \leq 3\pi/2$

The output characteristics are presented with thick lines again. The ones situated in the first quadrant for $\pi/2 \leq \delta \leq \pi$ correspond to the mode of energy transfer to the load, and the others in the second quadrant for $\pi \leq \delta \leq 3\pi/2$ – to the mode of energy return back to the power source respectively. The ordinate appears to be the boundary between the two modes. Referring to it, the characteristics for the equally distanced from π control angles are situated symmetrically.

The analysis of the output characteristics shows that independently from the control angle value the converter behaves such as an ideal current source.

Normalized control characteristics for the output current I_0 as a function of the control angle δ are presented in fig.6. They are drawn on the basis of equation (12) when the converter operates at several different values of the frequency detuning $\nu = 1,08; 1,10; 1,15; 1,20; 1,30$.

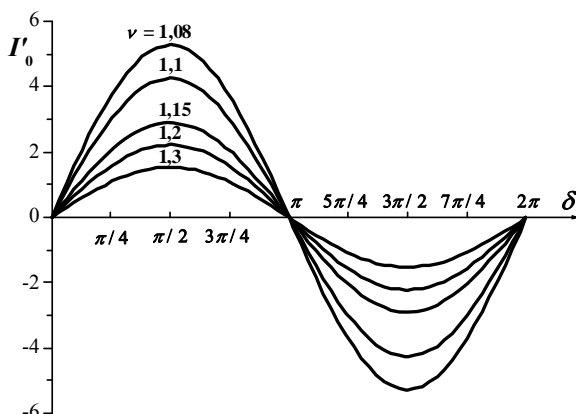


Fig. 6. Control characteristics of the converter

In this case, the abscissa appears to be the boundary between the two modes of energy transfer.

Fig. 6 shows that independently from the value of the operating frequency these characteristics have two extreme

values. The output current gains same values for these points.

Obviously, from the point of view of the converter output power control, the most appropriate interval of variation of the angle δ is the one limited by the two extreme values. Moreover, the control characteristics can be assumed as linear for a comparatively wide range.

V. CONCLUSION

A bidirectional resonant DC/DC converter operating over the resonant frequency with phase-shift control of the output power is examined. A theoretical examination of the converter is carried out on the basis of harmonic analysis with taking into consideration only the impact of the first harmonic. As a result, dependencies for the basic parameters of the converter are obtained and its output and control characteristics are drawn. The control intervals are determined for the two energy transfer modes – to the load and back to the power supply source. The boundary conditions for the power switch area of commutation at zero voltage are defined. Several specialties in the converter output power control are discussed. The most suitable interval of variation of the control angle is pointed.

The presented output and control characteristics show that the examined resonant DC/DC converter is bidirectional only by means of current as its output voltage cannot change its sign. Furthermore, the change of the energy transfer direction is unambiguously determined by the control angle. Moreover, the converter behaves such as an ideal current source for both modes.

The obtained analytical results can be used for design of this and similar resonant DC-DC converters.

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