Increased-Efficiency Single-Phase Direct Boost A.C.-A.C. Converter

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***Abstract*—A.C.-A.C. converters are currently used in A.C. motor speed control, adjustable switching A.C. sources, A.C. voltage waveform restorers, electric heating, lighting control etc. In many applications, they replaced alternative voltage regulators that use thyristors or triacs due to their benefits: they are smaller, more efficient and the waveforms of the voltage, load current and input current are much better.**

I. Introduction

A.C.-A.C. boost converters are used for supplying power to 220V receivers from a 110V grid and for increasing voltage for the receivers connected to the end of a long line, as well as for electronic boost transformers.

There are numerous papers that present the power and control circuits of A.C.-A.C. converters. Reference [1] presents an improved-switching buck-boost A.C. chopper and reference [2] analyses a buck A.C. chopper with IGBTs. References [3] and [4] research a control strategy for harmonic reduction, and in [5] and [6], the three-phase A.C. chopper with IGBTs that is studied by simulations has a significant low-cost circuit. A three-level A.C.-A.C. chopper is presented in [7]. Reference [8] shows a resonant A.C.-A.C. converter using the sliding control mode, and [9] studies a three-phase A.C.-A.C. converter with nine IGBTs.

Three-phase devices are also approached in reference [10], which introduces a method for assessing the various A.C.-A.C. converter topologies. Reference [12] suggests the modification of the sinusoidal PWM technique [11] in order to improve the power factor in a single-phase A.C.-A.C. converter.

References [13] and [17] present calculation methods for converter components allowing the functioning of unit power factor A.C.-A.C. converters. In [14], a new family of three-level A.C. choppers is introduced and [15] shows the circuit of a voltage restorer with A.C. choppers designed with commercial modules. Reference [16] analyses by means of simulations an improved-performance buck A.C.-A.C. converter. A boost converter allowing the modification of the ratio between the resistance and the load inductance is presented in [18], and in [19] the authors test by simulations and experiments an improved-switching buck converter.

The paper presents a boost A.C.-A.C. converter that uses an extra inductor and a new control technique, thus considerably reducing switching losses. The circuit also allows increased operating frequency. It was submitted to simulations in order to test its adequate functioning.

II. Circuit Analysis

Fig. 1 presents the circuit suggested for the boost A.C.-A.C. converter. In this circuit, L_1 and L_2 are boost inductors, MOS transistors Q_1 and Q_3 operate at the converter switching frequency f_S, and MOSs Q_2 and Q_4 are controlled at the alternating current grid frequency, i.e 50Hz.

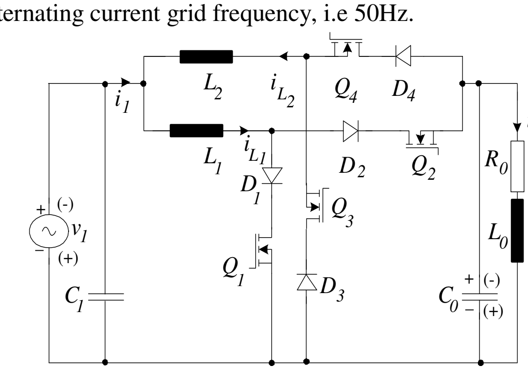

Figure 1 Circuit for the boost A.C.-A.C. converter

Fig. 2 presents the waveforms of the grid voltage v_1 and of the current i_1 , as well as those of the control signals of the four MOS transistors. During the positive half-period of the grid voltage, Q_2 is always in conduction, and Q_1 is under *PWM* control at a converter operating frequency $f_S = \frac{1}{T_s}$. In the meanwhile transistors Q_3 and Q_4 remain switched off. During the negative half-period of the grid voltage, Q_4 is always in conduction, and Q_3 is under *PWM* control. In fact, this control technique results in two D.C.-D.C. boost converters functioning back-to-back and their input voltage is equal to the rectified grid voltage. Therefore, the specifications of the A.C.-A.C. converters can be improved up to those of the D.C.-D.C. converters.

The control circuit is very simple and it is shown in Figure 3. In order to avoid the distortion of the waveform of the current i_1, the fundamental of this current should be in phase with voltage v_1, which is possible if the output capacitor C_O is chosen adequately. If the phase shift between fundamentals v_{O1} and i_{O1} varies widely, a signal generated by the transducer of current i_1, must be applied at the non-inverting input of the comparator. Switching state changes, i.e. Q_2 is switched off and Q_4 is switched on when the current i_1 is 0.

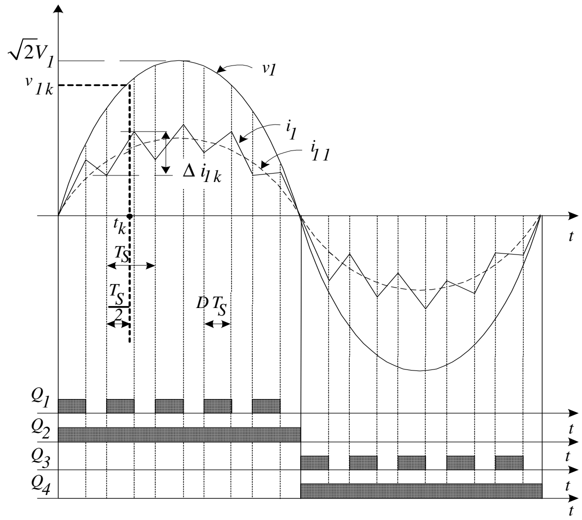

Figure 2 Waveforms of grid voltage v_1 and of input current i_1 and waveforms of the control signals of transistors Q_1 - Q_4

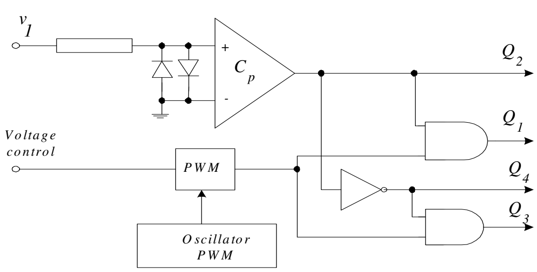

Figure 3 The control circuit diagram

In order to choose the capacitor C_O, we need to analyse the equivalent converter circuit. Assuming that voltages v_1 and v_O are constant throughout the switching period T_S and that the devices Q_1 - Q_4 are ideal switches, when the capacitor C_O is disconnected, this results in the simplified equation presented below [10], [14]:

$$\frac{v_1(t)}{1-D}=\frac{L}{(1-D)^2}\frac{di_1(t)}{dt}+v_O(t) \tag{1}$$

This equation refers to the equivalent converter circuit in Fig. 4, which allows C_O to be selected so that voltages v_1 and i_1 , assumed to be sinusoidal, should be in phase.

See below the input voltage equation:

$$v_1=\sqrt{2}V_1\sin\omega t,\ \omega=2\pi f \tag{2}$$

and the frequency modulation ratio is:

$$m_f=\frac{f_S}{f}=\frac{T}{T_S} \tag{3}$$

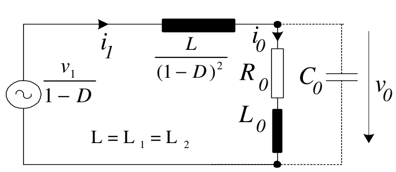

Figure 4 The equivalent circuit of the boost A.C.-A.C. converter

In the switching period k, the input voltage, assumed to be constant, is calculated below:

$$v_{1k}=\sqrt{2}V\sin\omega t_k\ ,\ t_k=(k-1)T_S+\frac{T_S}{2},k=\overline{1,m_f} \tag{4}$$

And the boost converter output voltage is :

$$v_{Ok}=\sqrt{2}V_O\sin(\omega t_k-\varphi)=\frac{\sqrt{2}V_1}{1-D}\sin(\omega t_k-\varphi) \tag{5}$$

It is assumed that, in a given application, the input voltage ranges between

$$V_{1(\min)}\le V_1\le V_{1(\max)} \tag{6}$$

and that the output voltage should vary between

$$V_{O(\min)}\le V_O\le V_{O(\max)} \tag{7}$$

The maximum duty cycle is calculated as:

$$D_{(\max)}=1-\frac{V_{1(\min)}}{V_{O(\max)}} \tag{8}$$

and the minimum duty cycle is:

$$D_{(\min)}=1-\frac{V_{1(\max)}}{V_{O(\min)}} \tag{9}$$

The ripple current through inductors L_1 and L_2 is calculated as:

$$\Delta i_L=\frac{D\,v_1}{L\,f_S}\ ;\ \Delta i_{L(\max)}=\frac{D_{(\max)}\sqrt{2}V_{1(\max)}}{L\,f_S} \tag{10}$$

The last equation refers to the size of inductors L_1 and L_2, as the maximum ripple current should be calculated as a percentage from the maximum load current:

$$I_{O(\max)}=\frac{\sqrt{2}V_{O(\max)}}{\sqrt{R_O{}^2+(L_O\,\omega)^2}} \tag{11}$$

Under the simplifying conditions required, there are several specifications for a switching period k in the positive half-period of the grid voltage. The average current through the inductor is:

$$I_{L1k\,AVR}=\frac{R_O v_{1k}}{(1-D)^2\left[R_O{}^2+(\omega L_O)^2\right]} \tag{12}$$

The average current through the transistor Q_1 and the diode D_1 is:

$$I_{Q1k\,AVR}=I_{D1k\,AVR}=D\cdot I_{L1k\,AVR}=\frac{D\,R_O v_{1k}}{(1-D)^2\left[R_O^2+(\omega L_O)^2\right]} \tag{13}$$

The average current through the transistor Q_2 and the diode D_2 is:

$$I_{Q2k\,AVR}=I_{D2k\,AVR}=(1-D)\cdot I_{L1k\,AVR}=\frac{R_O v_{1K}}{(1-D)\left[R_O^2+(\omega L_O)^2\right]} \tag{14}$$

The maximum repetitive current through the inductor, the transistors and the diodes is:

$$I_{L1k\,RM}=I_{Q1kRM}=I_{D1k\,RM}=I_{Q2kRM}=I_{D2k\,RM}=$$
$$=\frac{R_O v_{1K}}{(1-D)^2\left[R_O^2+(\omega L_O)^2\right]}+\frac{D v_{1K}}{2L_1 f_S} \tag{15}$$

The voltage stress of the transistor Q_1 is:

$$V_{Q1RM} = v_{Ok} = \frac{v_{1k}}{1-D} \tag{16}$$

and the reverse voltage applied to diode D_2 is:

$$V_{D2RRM} = v_{Ok} = \frac{v_{1k}}{1-D} \tag{17}$$

In this alternating period of the grid, transistors Q_3 and Q_4 are switched off. The reverse voltage applied to diode D_3 and the direct voltage applied to transistor Q_4 are:

$$V_{D3RRM} = v_{1k}\ ;\ v_{Q4RM} = v_{Ok} - v_{1k} = \frac{D\ v_{1k}}{1-D} \tag{18}$$

Since v_{1K} varies sinusoidally and the transistors and the diodes are in conduction in only one of the two half-periods of the grid voltage, here are the resulting maximum stress in terms of voltage and current:

The maximum repetitive current through the inductor, the transistors and the diodes is:

$$I_{LRM} = I_{QRM} = I_{DRM} = \\ = \frac{R_O\sqrt{2}V_{1(\max)}}{\left(1-D_{(\max)}\right)^2\left[R_O^2+\left(\omega L_O\right)^2\right]} + \frac{D_{(\max)}\sqrt{2}V_{1(\max)}}{2L_1 f_S} \tag{19}$$

The average current through transistors Q_1 and Q_3 and through diodes D_1 and D_3 is:

$$I_{Q1AVR} = I_{Q3AVR} = I_{D1AVR} = I_{D3AVR} = \\ = \frac{D_{(\max)}\sqrt{2}V_{1(\max)}R_O}{\pi\left(1-D_{(\max)}\right)^2\left[R_O^2+\left(\omega L_O\right)^2\right]} \tag{20}$$

The average current through transistors Q_2 and Q_4 and through diodes D_2 and D_4 is:

$$I_{Q2AVR} = I_{Q4AVR} = I_{D2AVR} = I_{D4AVR} = \\ = \frac{R_O\cdot\sqrt{2}V_{1(\max)}}{\pi\left(1-D_{(\max)}\right)\left[R_O^2+\left(\omega L_O\right)^2\right]} \tag{21}$$

The average current through the inductors is:

$$I_{L1AVR} = I_{L2AVR} = \frac{R_O\cdot\sqrt{2}V_{1(\max)}}{\pi\left(1-D_{(\max)}\right)^2\left[R_O^2+\left(\omega L_O\right)^2\right]} \tag{22}$$

The maximum repetitive voltages applied to transistors Q_1 and Q_3 are:

$$V_{Q1RM} = V_{Q3RM} = \frac{\sqrt{2}V_{1(\max)}}{1-D_{(\max)}} \tag{23}$$

The maximum reverse voltages applied to diodes D_1 and D_3 are:

$$V_{D1RRM} = V_{D3RRM} = \sqrt{2}V_{1(\max)} \tag{24}$$

The maximum repetitive voltages applied to transistors Q_2 and Q_4 are:

$$V_{Q2RM} = V_{Q4RM} = \frac{D_{(\max)}\sqrt{2}V_{1(\max)}}{1-D_{(\max)}} \tag{25}$$

The maximum reverse voltages applied to diodes D_2 and D_4 are:

$$V_{D2RRM} = V_{D4RRM} = \frac{\sqrt{2}V_{1(\max)}}{1-D_{(\max)}} \tag{26}$$

III. SIMULATION RESULTS

The adequate functioning of the circuit was tested by simulation with a resistive load *R=150* ohm and an inductive load *L=32 mH*.

The components of the simulated circuit have the following specifications: input capacitor C_1 = *1uF*, output capacitor C_O = *3uF*, boost inductances $L_1 = L_2$ = *12 mH*. The amplitude of the grid voltage is *155V* and the switching frequency is *f=100KHz*.

The simulations were based on the following values of the duty cycle: *D=0.2, 0.4*, and *0.6*. The paper presents only some of the waveforms obtained, namely the most representative ones. Thus, Fig. 5 shows the waveforms of the grid voltage v_1 and of the load voltage v_0, as well as the waveforms of the current i_1 supplied by the grid and of the load current i_0 for *D=0.2*.

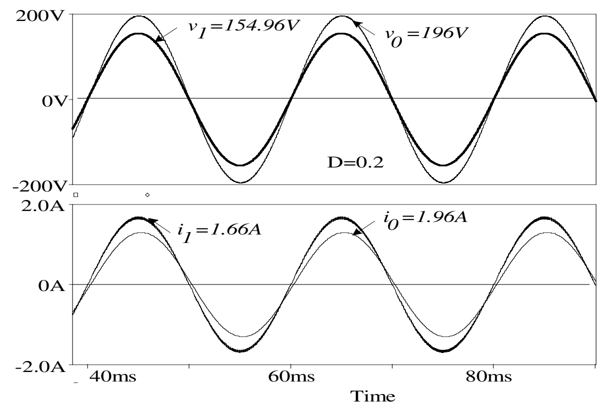

Figure 5. Waveforms of voltage v_1, voltage v_0, current i_1 and current i_0, for D=0.2

Figure 6 presents the same waveforms for D=0.4.

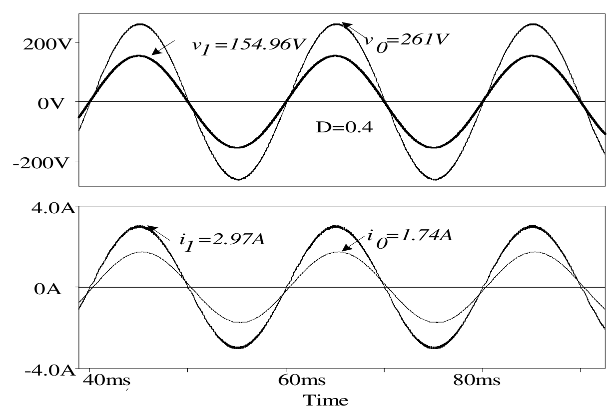

Figure 6. Waveforms of the voltage v_1, voltage v_0, current i_1 and current i_0, for D=0.4

Figure 7 presents the same waveforms for D=0.6. The waveforms of the load voltage and current are nearly sinusoidal.

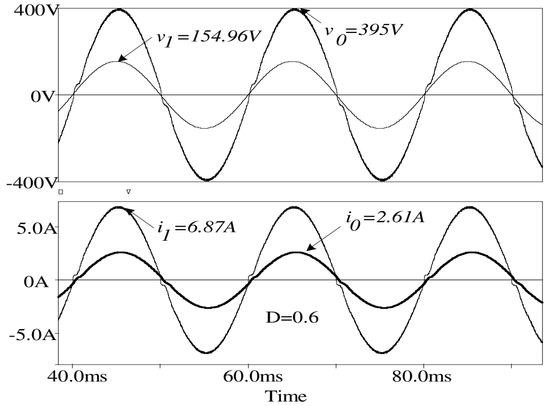

Figure 7. Waveforms of the voltage v_1, voltage v_0, current i_1 and current i_0, for D=0.6

Figure 8 shows the efficiency variations depending on the duty cycle *D*, also based on the simulation results. We notice thus that the efficiency is high for all the values of *D*.

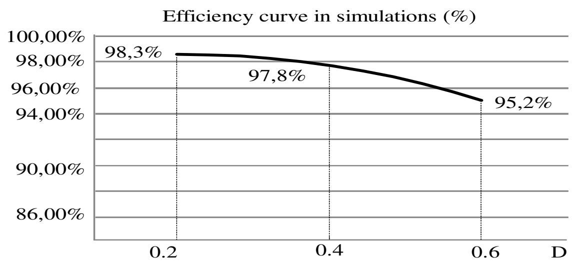

Figure 8 Variation of the efficiency depending on the duty cycle *D*.

Fig. 9 shows the control characteristic of the circuit presented, comparing results in simulation and in theory.

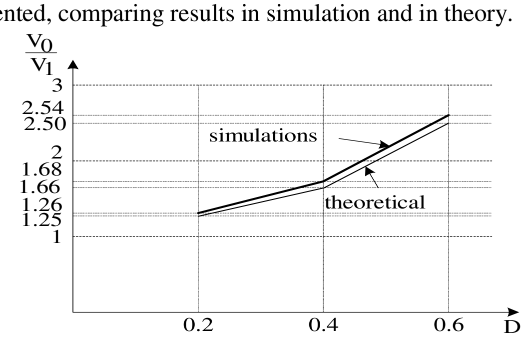

Figure 9 The control characteristic of the circuit presented

CONCLUSIONS

This paper presents the analysis and simulation of a boost A.C.-A.C converter with four controlled transistors, two of them in conduction in one half-period of the sinusoidal alternation. The adequate functioning of the circuit was tested by simulation for various values of the control factor of the MOS transistors. Based on the various values of the control factor D=0.2, 0.4 and 0.6), we presented the main waveforms of the voltages and of the input and output currents, as well as the variation of the converter energy efficiency.

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