SINGLE-PHASE AC-AC CONVERTER

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Abstract. This paper presents a direct AC-AC single-phase buck-boost converter. The circuit is simple and has good performances, whatever the load nature. The correct functioning of the circuit at a 20 kHz switching frequency was tested both by simulation and experimentally.

Key words: choppers; power conversion; circuit simulation.

1. Introduction

AC-AC converters are currently used in numerous fields, such as: AC motor drive, adjustable AC power supplies, electronic transformers, voltage waveform restorers, adjustable impedances, etc. These converters successfully replace alternating voltage variators using thyristors or triacs. Since the functioning frequency is high (more than 20 kHz), there is no noise, filters are small in size, efficiency is high and the current from the power supply is nearly sinusoidal.

The first AC-AC converters analysed were buck converters (AC choppers) (Revenkar, 1977). References (Chose & Park, 1989; Jang & Choe, 1991; Do-Hyum & Choe, 1995) present improved PWM techniques, which increase the power factor and eliminate certain harmonics (in the absence of grid filters). In (Lucanu & Ursaru, 2003), simulations were used to analyse an

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IGBTs chopper at a 5kHz frequency. Reference (Congwei & Bin, 2009) presents a three-phase AC-AC converter with 9 IGBTs, and (Lai & Wang, 2009) suggests an evaluation method for three-phase AC converters.

In Lucanu & Ursaru (2005) and Kim & Min (1998), the choppers presented have improved switching and increased efficiency, but the circuits are complex. References Li & Yang (2001) and Thiago & Clovis (2011) present AC choppers with three-level converters and topologies that use commercial power modules (Neascu, 2013). In Aghion & Lucanu (2012) a high-performance AC-AC single-phase converter with two inductances and four IGBTs is presented.

This paper presents a direct AC-AC single-phase buck-boost converter using two IGBTs, eight diodes, an inductance and a capacitor (except for the grid filters). In fact, it is a classic buck-boost converter structure, where the two switches used are AC switches and a grid filter was added. The converter functions adequately irrespective of the load nature and it can insure a bidirectional energy flow if the load contains an AC source. The circuit design equations are presented below. The adequate functioning of the converter is checked both by simulation and experimentally; the tests were applied to a converter prototype developed by the authors. The prototype was used for connecting a device to the grid, when the nominal power supply is different of nominal voltage load.

2. Circuit Analysis

Fig. 1 presents the circuit of the AC-AC single-phase buck-boost converter.

![Diagram of the AC-AC single-phase buck-boost converter](image)

The AC-AC converter contains the grid filter $L_f$, $C_f$, the inductor $L$, the capacitor $C$ connected in parallel with the load impedance $R$, $L_3$ and two AC switches: one of them is made up of the IGBT $S_1$ and the diodes $D_1$-$D_4$, and the other one includes $S_2$ and $D_5$-$D_8$. The snubber circuits $R_{S1}$, $C_{S1}$ and $R_{S2}$ $C_{S2}$ are connected in parallel with the IGBTs.
Fig. 2 shows the waveform of the voltage \( v \) at the \( C_f \) capacitor terminals, and the generation of the control signals for the IGBT's \( S_1 \) and \( S_2 \). If the current through the inductor \( L \) is \( i > 0 \), for time intervals \([0, DT]\), it will flow through \( D_1, S_1, D_2, L \). The load resistor \( R \) is powered by the capacitor \( C \). For time intervals \([DT, T]\), the current \( i \) will flow through \( L \), the \( R-C \) circuit, \( D_6, S_2, D_5 \).

We used uniform PWM control (Barleanu & Baitoiu, 2012), in which the conduction durations for the two switches are the same in all switching periods \( T \), as in DC converters, and \( f = 1/T \) is the switching frequency (Valachi & Timis, 2009). The equations that describe the functioning of the converter rely on the following simplifying hypotheses: the passive components are ideal, the power devices are ideal switches, the voltage \( V \) on the capacitor \( C_f \) and the voltage \( V_0 \) on the load are sinusoidal and remain constant for a period \( T \), and the load is purely resistive. If \( V_k \) and \( V_{0k} \) are voltages from the middle of the switching period \( K \) and \( \omega \) is the grid voltage frequency, we can write the following equations:

\[
\begin{align*}
    v &= \sqrt{2}V \sin \omega t, \\
    v_k &= \sqrt{2}V \sin \omega t_k = \text{const.}, \quad t_k (k - 1)T + \frac{T}{2}, \\
    v_{0k} &= \sqrt{2}V_0 \sin \omega t_k = \text{const.}
\end{align*}
\]

(1)

For the interval \([0, DT]\), when \( S_1 \) is conducting and \( S_2 \) is off:
\[
\begin{aligned}
    v_{Lk} &= L \frac{di}{dt} v_K, \\
    i_k &= I_{km} + \frac{v_k}{L}.
\end{aligned}
\]  

By replacing \( t = DT \) to the eq. (2), we get the ripple of the current \( i \) through the inductor \( L \) in the switching period \( K \):

\[
\begin{aligned}
    i_k &= I_{KM} = I_{km} + \frac{v_k}{L} DT, \\
    \Delta i_k &= I_{KM} - I_{km} = \frac{D}{Lf} v_K.
\end{aligned}
\]

For the interval \([DT, T]\), when \( S_1 \) is off and \( S_2 \) is conducting:

\[
\begin{aligned}
    v_{Lk} &= v_0 K, \\
    i_k &= I_{KM} = I_{km} - \frac{v_0 K}{L}, \quad t' \in [0, (1 - D)T].
\end{aligned}
\]

Since this is a buck-boost converter, the control characteristic corresponding to the switching period \( K \) can be approximated by the equation below:

\[
v_{0K} = \frac{D}{1 - D} v_K.
\]

In a similar way we obtain the following average values of the currents through the bidirectional switches, corresponding to the switching period \( K \):

\[
\begin{aligned}
    I_{s,K} &= \left( \frac{D}{1 - D} \right)^2 \frac{v_K}{R}, \\
    I_{s,K} &= \frac{D}{1 - D} \frac{v_K}{R}.
\end{aligned}
\]

and the following values for the maximum repetitive currents:

\[
I_{s, KM} = I_{s, KM} = I_k + \frac{\Delta i_k}{2} = \frac{D v_K}{(1 - D)^2 R} + \frac{D v_K}{2 Lf}.
\]

The maximum collector-emitter voltages on the two IGBTs in the switching period \( K \) are the following:

\[
V_{s, KM} = V_{s, KM} = v_0 K = \frac{D v_K}{1 - D}.
\]
The voltages $V_K$ have sinusoidal variation, therefore the average values of the currents through the switches on a period $T_m$ of the AC grid voltage are:

\[
\begin{align*}
I_{S_{\text{avr}}} &= \left( \frac{D}{1-D} \right)^2 \frac{2\sqrt{2}V}{\pi R}, \\
I_{S_{\text{avr}}} &= \frac{D}{1-D} \frac{2\sqrt{2}V}{\pi R}.
\end{align*}
\]  
(9)

and the maximum repetitive currents are:

\[
I_{S_{M}} = I_{S_{M}} = \frac{D\sqrt{2}V}{(1-D)^2 R} + \frac{D\sqrt{2}V}{2Lf}.
\]  
(10)

The IGBT’s stress voltage is:

\[
V_{S_{M}} = V_{S_{M}} = \frac{D\sqrt{2}V}{1-D}.
\]  
(11)

The normalised current ripple results from equations (3) and (6):

\[
\frac{\Delta I_k}{I_k} = \frac{(1-D)^2 R}{Lf}.
\]  
(12)

This equation can be used for calculating the inductor $L$. The normalised ripple of the output voltage can be calculated by:

\[
\frac{\Delta v_0}{v_0} = \frac{1-D}{RCf}.
\]  
(13)

This equation allows for the calculation of the value of the capacitor $C$.

3. Simulation and Experimental Results

The adequate functioning of the circuit was tested by simulation and experimental prototype. The load used in the simulations and in the prototype is the same - a resistive load: $R = 390$ Ω, inductive load $L_S = 750$ mH, $R = 390$ Ω and the switching frequency is $f = 20$ kHz.

Table 1 shows the main parameters used for the simulations and the experimental prototype.
Table 1

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Symbol</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>AC input voltage</td>
<td>$V_m$</td>
<td>110 V (RMS)</td>
</tr>
<tr>
<td>Input frequency</td>
<td>$f_m$</td>
<td>50 Hz</td>
</tr>
<tr>
<td>Resistive load</td>
<td>$R$</td>
<td>150 Ω</td>
</tr>
<tr>
<td>Inductive Load</td>
<td>$R/L_s$</td>
<td>150 Ω/750 mH</td>
</tr>
<tr>
<td>Transistors</td>
<td>$S_1, S_2$</td>
<td>IRGB8B60KPBF</td>
</tr>
<tr>
<td>Diodes</td>
<td>$D_1-D_8$</td>
<td>MUR460</td>
</tr>
<tr>
<td>Capacitor</td>
<td>$C$</td>
<td>2 μH</td>
</tr>
<tr>
<td>Input filter</td>
<td>$L_f$</td>
<td>13 mH</td>
</tr>
<tr>
<td>$C_f$</td>
<td>10 μF</td>
<td></td>
</tr>
<tr>
<td>Microcontroller</td>
<td></td>
<td>PIC16F684</td>
</tr>
<tr>
<td>Snubber</td>
<td>$R_{S1,S2}$</td>
<td>100 Ω</td>
</tr>
<tr>
<td></td>
<td>$C_{S1,S2}$</td>
<td>3.3 nF</td>
</tr>
</tbody>
</table>

Fig. 3 shows the waveforms obtained by simulations, - the waveforms of the $i_0$ current load and of the $i_m$ current source and the waveforms of the $v_0$ voltage load and of the $v_m$ voltage source and for a duty factor $D = 0.3$.

![Waveform Diagram](image)

Fig. 3 – Waveforms of the $i_m$ current, $i_0$ current and $v_m$ voltage, $v_0$ voltage for $D = 0.3$, in the resistive load case $R = 150$ Ω.

Fig. 4 shows the same waveforms for $D = 0.3$, for the inductive load case, where the output impedance is: $R = 150$ Ω and $L = 750$ mH.

Fig. 5 shows the waveforms obtained by simulations, namely the waveforms of the $i_0$ current load and of the $i_m$ current source and the waveforms of the $v_0$ voltage load and of the $v_m$ voltage source and for a duty factor $D = 0.7$.

Fig. 6 shows the same waveforms for $D = 0.7$, for the inductive load case, where the output impedance is: $R = 150$ Ω and $L = 750$ mH.
Fig. 4 – Waveforms of the $i_m$ current, $i_0$ current and $v_m$ voltage, $v_0$ voltage for $D = 0.3$, in the inductive load case ($L = 750 \text{ mH}$ and $R = 150 \Omega$).

Fig. 5 – Waveforms of the $i_m$ current, $i_0$ current and $v_m$ voltage, $v_0$ voltage for $D = 0.7$, in the resistive load case $R = 150 \Omega$.

Fig. 6 – Waveforms of the $i_m$ current, $i_0$ current and $v_m$ voltage, $v_0$ voltage for $D = 0.7$, in the inductive load case ($L = 750 \text{ mH}$ and $R = 150 \Omega$).
As presented in Fig. 7, the prototype circuit is made up of two boards; one of them includes the microcontroller, the LCD, the drivers and the low-voltage supply circuit, and the other one contains the proposed power circuit, based on the schematic presented in Fig. 1.

![Experimental setup](image)

Fig. 7 – Experimental setup.

Fig. 8 shows the waveforms obtained by measurements, the waveforms of the $i_0$ current load and of the $i_m$ current source and the waveforms of the $v_0$ voltage load and of the $v_m$ voltage source and for a duty factor $D = 0.3$.

![Waveforms](image)

Fig. 8 – Waveforms of the $i_m$ current, $i_0$ current and $v_m$ voltage, $v_0$ voltage for $D = 0.3$, in the resistive load case $R = 150 \, \Omega$. 
Fig. 9 show the same waveforms for $D = 0.3$, for the inductive load case, where the output impedance is: $R = 150 \ \Omega$ and $L = 750 \ \text{mH}$.

Fig. 9 – Waveforms of the $i_m$ current, $i_0$ current and $v_m$ voltage, $v_0$ voltage for $D = 0.3$, in the inductive load case ($L = 750 \ \text{mH}$ and $R = 150 \ \Omega$).

Fig. 10 shows the waveforms obtained by measurements, the waveforms of the $i_0$ current load and of the $i_m$ current source and the waveforms of the $v_0$ voltage load and of the $v_m$ voltage source and for a duty factor $D = 0.7$.

Fig. 10 – Waveforms of the $i_m$ current, $i_0$ current and $v_m$ voltage, $v_0$ voltage for $D = 0.7$, in the resistive load case $R = 150 \ \Omega$. 
Fig. 11 show the same waveforms for $D = 0.7$, for the inductive load case, where the output impedance is: $R = 150\, \Omega$ and $L = 750\, \text{mH}$.

![Waveforms](image)

Fig. 11. Waveforms of the $i_m$ current, $i_0$ current and $v_m$ voltage, $v_0$ voltage for $D = 0.7$, in the inductive load case ($L = 750\, \text{mH}$ and $R = 150\, \Omega$).

Fig. 12 $a$ shows THD for the input current obtained by simulations and measurements for resistive load and Fig. 12 $b$ shows THD for the input current obtained by simulations and measurements for inductive load.

![THD Analysis](image)

Fig. 12 – THD analysis for the input current: $a$ – for the resistive load, $b$ – for inductive load.
Fig. 13 illustrates the efficiency obtained by simulations and measurements for the cases of the resistive load. Fig. 13b shows the efficiency obtained by simulations and measurements for the cases of the inductive load.

Fig. 13 – Efficiency: a – for the resistive load, b – for the inductive load.

4. Conclusions

The paper presents a direct AC-AC buck-boost converter circuit containing, beside the grid filter, two bidirectional current switches, each made up of one IGBT, four diodes and a boost inductance.

Switches were controlled by uniform sampling, with the same duty cycle in all the switching periods. The resulting control circuit is simple and the energy flow can be bidirectional. The adequate functioning of the circuit was tested by simulation, as well as on a laboratory prototype.

The results allowed the identification of the control characteristic and of the functioning efficiency. The waveforms obtained for the load voltage and current are very good. The supplied current is sinusoidal, therefore the circuit can be used particularly for converting the RMS voltage from 110 V to lower or upper voltage depending of the duty cycle D.

REFERENCES


Valachi A., Timis M., Danubianu M., Some Contributions to to Synthesis and Implementation of Multifunctional Registers. 11th WSEAS Internat. Conf. on Automatic Control, Modelling & Simulation (acmos09), Istanbul, Turkey, May 30 - June 1, 2009, 146-149.

CONVERTOR AC-AC MONOFAZAT

(Rezumat)

Circuitul propus, alimentat în curent alternativ, are rolul de a converti tensiunea furnizată sarcinii, într-o tensiune alternativă, de aceeași alură, dar care se modifică conform caracteristicii de funcționare întâlnite în convertoarelor dc-de de tip buck-boost (convertor mixt).

Strategia de comandă a tranzistoarelor din compoziția convertorului este simplă, managementului de control al gestionării energiei furnizate unei sarcini este mult simplificat, permitând pe ansamblu obținerea unor performanțe globale net superioare, față de managementul de control utilizat în topologiile clasice.