AC-DC Bidirectional Single-Phase Step-Down Converter with High Power Factor

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***Abstract* – This paper presents a converter that allows the operation as rectifier or inverter, with high power factor, which the voltage output can be lower, equal or greater than the peak of the input voltage, besides that working with only a cell of conventional switching. The operation stages, equations, control strategy and design of the converter are presented. Finally, simulation results are shown for operation as rectifier and inverter.**

I. INTRODUCTION

Most of the single-stage topologies that allow operating as rectifier have an input AC voltage lower than the output DC voltage. Then, to obtain lower output voltage it is necessary to add another stage as a buck converter, to achieve the voltage magnitude required. Besides, they used at least two switching cells adding losses and diminishing the efficiency of the topology. Among the topologies used at the moment, only one allows the operation with greater or lower voltage in the output converter. This converter was proposed by Cáceres and Barbi [1],[2]. It consists of two individual step-up converters as shown in Fig. 1. In the step-up inverter the load connected in differential mode can theoretically, obtain the output voltage with any value and shape. Besides, respected the condition that individually V_{c1} and V_{c2} are greater than V_{dc}, a degree of freedom exists in the selection of these voltages, where only the difference between them interests to the load.

This way, a sinusoidal output voltage can be obtained with a capacitor with fixed voltage and imposing one sinusoidal variation in the other capacitor, or using two sinusoidal references with a phase-shift of 180° between them, as shown in Fig. 2. Each converter produces a sinusoidal output voltage single-pole with a continuous component, as shows in Fig. 2, with waveforms of V_{c1} and V_{c2}. The load is connected in differential mode between the converters, annulling the continuous components.

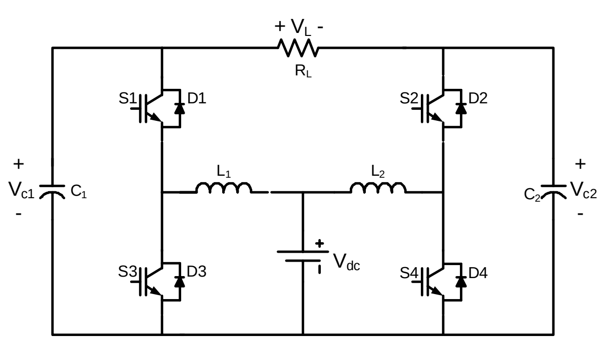

Fig. 1. Basic diagram of the inverter proposed by Cáceres and Barbi [1].

The modulation in each converter is 180° degrees phase-shift in relation to the other, it maximizes the excursion of voltage through the load. However, the difference of phase between the two converters can have any value. This is presented as an alternative for the control output voltage of (V_{c1}-V_{c2}).

The generation of the bipolar output voltage is obtained by a push-pull arrangement, forcing the converter to operate as source and the another one as load, being bidirectional in current.

Colling and Barbi [4] propose the circuit operating as rectifier. Operating as rectifier it is necessary to invert its

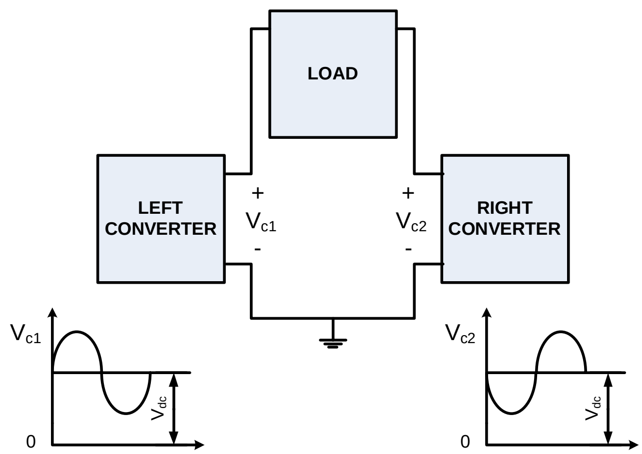

Fig. 2. Basic configuration to get inversion AC-DC [3].

power flow and to establish some control on the absorbed current of AC source. Therefore, an inductor in series with the AC source is included, as shown in Fig. 3. Due to that reversibility of power flow, the elements (inductive and source) are not identified as input or output any more, but as AC-side or DC-side.

This paper proposes a variation of the circuit proposed by Colling and Barbi [4] simplifying the topology presented in Fig. 3.

This simplification consists of substituting the converter of the right side for a capacitor, such modification requires to increase the value of the capacitor, therefore it is able to maintain a higher DC voltage to V_{dc}. With this, it is possible to diminish the quantity of semiconductors and magnetic elements without modifying the operation of the converter.

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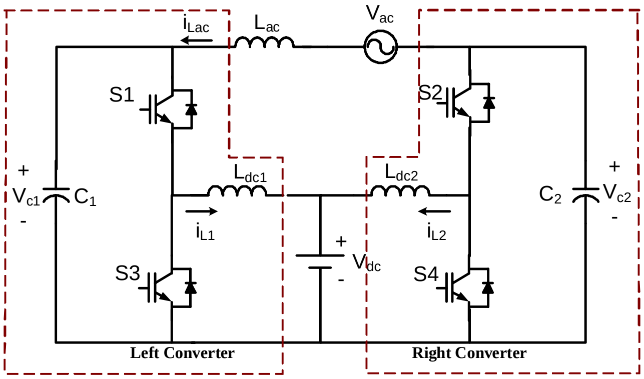

Fig. 3. Rectifier diagram proposed by Colling and Barbi.

II. TOPOLOGY PROPOSAL

The original topology operating as rectifier is shown in Fig. 3. The function of the converter of the right side consists of keeping a sinusoidal voltage with a DC level, in the capacitor C_2, greater that V_{dc}. The voltage in the capacitor C_1, which it must be complementary to V_{c2} (Fig. 2), is controlled for the left converter. This also adjusts the value of the current i_{Lac} to obtain the power factor correction.

The idea of the proposal consists of substituting the right side converter for a capacitor and to control its voltage to follow a DC value. This value must be greater than the output voltage V_{dc}. Therefore, the function of the left side converter will be to keep a sinusoidal voltage in the C_1 capacitor with one average component, equal to V_{c2}. The simplified and redesigned circuit is shown in Fig. 4.

This way, in the C_1 capacitor the magnitude of its voltage will be of $V_{ac}+V_{c2}$. Therefore, the structure operating as rectifier is similar to a step-down converter, with filter an input, the difference is that their input voltages have a component alternated above a DC voltage level.

A. Considerations for the operation of the circuit

The circuit is considered as a step-up inverter voltage. For the C_2 capacitor, its voltage must be (1).

$$v_{c2}^{*}(t) = V_{c2_{dc}}^{*} \tag{1}$$

The total sinusoidal excursion will be applied to the C_1

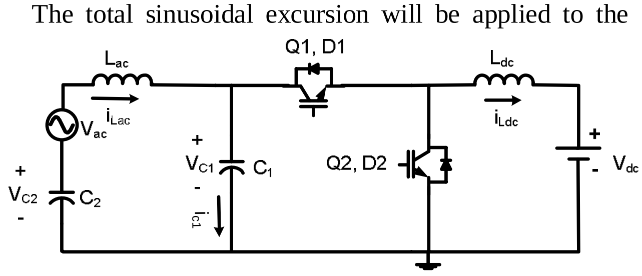

Fig. 4. Circuit inverter/rectifier proposed.

capacitor, leaving the C_2 capacitor only with a DC voltage level (contrary to the proposed by Cáceres in [3], where the AC voltage is divided between both capacitors, C_1 and C_2). Thus, the condition below must be satisfied with a margin of security, in way that never V_{c2} is lower than V_{dc} (2) .

$$V_{c2_{dc}}^{*} > V_{dc} + V_{acp} \tag{2}$$

Therefore,

$$v_{c2}^{*}(t) = \left(V_{dc} + V_{acp}\right) + \Delta V_{dc} \tag{3}$$

With ΔV_{dc} as the security voltage margin. It is considered that the AC mains voltage is $V_{acp} \cdot sin(\omega t)$. The sum of the AC voltage and the voltage in the C_2 capacitor is:

$$v_{ac}(t) + v_{c2}^{*}(t) = V_{acp} \cdot sin(\omega t) + \left(V_{dc} + V_{acp} + \Delta V_{dc}\right) \tag{4}$$

therefore, the voltage in the C_2 capacitor can be a DC voltage with minimum magnitude, definite by (2).

The reference function for the current is defined in (5).

$$i_{Lac}^{*}(t) = I_{Lacp}^{*} \cdot sin(\omega t) \tag{5}$$

L_{ac} is designed to filter the oscillations in high frequencies originated from commutation. Its impedance and the drop voltage in the AC mains frequency, in steady state are low. It is concluded then that the voltage $v_{c1}(t)$ oscillates close to (6).

$$v_{c1}(t) = V_{acp} \cdot sin(\omega t) + \left(V_{dc} + V_{acp} + \Delta V_{dc}\right) \tag{6}$$

DC level applied in C_2 also is established in the C_1 capacitor.

The duty cycle of the $Q2$ switch is given by (7).

$$d(t) = 1 - \frac{V_{dc}}{V_{c2_{dc}}^{*} + V_{acp} \cdot sin(\omega t)} \tag{7}$$

For the energy balance the current in the L_{dc} inductor can be calculated by (8).

$$i_{Ldc}(t) = \frac{V_{c1}(t) \cdot \left(i_{Lac}(t) + i_{c1}(t)\right)}{V_{dc}} \tag{8}$$

In steady state the current in the inductor is given by (9).

$$\begin{aligned} i_{Ldc}(t) = {} & \frac{V_{c2_{dc}}^{*} \cdot I_{Lacp}^{*} \cdot sin(\omega t)}{V_{dc}} + \frac{I_{Lacp}^{*} \cdot V_{acp}}{2} \cdot \\ & \cdot \frac{\left(1 - cos(2\omega t)\right)}{V_{dc}} + \omega \cdot C_1 \cdot \left\{ \frac{V_{c2_{dc}}^{*}}{V_{dc}} \cdot \right. \\ & \left. \cdot V_{acp} \cdot cos(\omega t) + \frac{V_{acp}^{2}}{2 \cdot V_{dc}} \cdot sin(2\omega t) \right\} \end{aligned} \tag{9}$$

The current i_{Ldc} is described by (9). It still has the excursions in high frequencies caused for the switching of

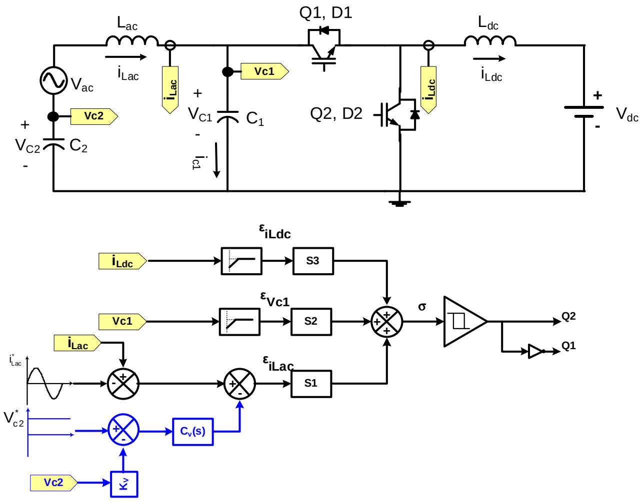

Fig. 5. Rectifier circuit and blocks diagram of the control circuit.

the switches *Q1* and *Q2*. Can be noted that the current presents a continuous component, responsible for the transference of energy, and alternating components of first and second order of the power system frequency. The circulation of these current components is an intrinsic characteristic of the circuit. The alternating components of i_{Ldc} (9) correspond to the circulating power reactive in the circuit.

The active power transferred to the sinusoidal source is given by (10). It is equivalent to the power supplied for a sinusoidal source with a current in phase with it.

$$\mathrm{P} = \mathrm{V}_{dc} \cdot \mathrm{I}_{\mathrm{Ldc}_{med}} = \mathrm{V}^{*}_{\mathrm{c2}_{acp}} \cdot \mathrm{I}^{*}_{\mathrm{Lacp}} \tag{10}$$

The operation as rectifier all the terms with i_{Lac} (or I^{*}_{Lacp}) are inverted.

B. Analysis of the circuit

The switches *Q1* and *Q2* receive pulses complementary to prevent the discontinuous conduction in the L_{dc} inductor. It is becoming possible the existence of only two structures, extracted of Fig. 4. When *Q2* and *D2* are opened and *Q1* and *D1* are closed meeting in the first stage operation. The second stage operation is represented where *Q1* and *D1* are closed and *Q2* and *D2* are opened.

First operation stage. I_{Lac} is assumed to be at its reference value and the switches state is described by variable γ.

$\gamma = 1 \;\; \Rightarrow$ *Q2 and D2 ON; Q1 and D1 OFF* .

The first stage of operation of the circuit is described by (11), (12) and (13).

$$\frac{di_{Ldc}}{dt} = \frac{V_{dc}}{L_{dc}} \tag{11}$$

$$\frac{di_{Lac}}{dt} = \frac{v_{ac} + v^{*}_{c2}}{L_{ac}} + \frac{v_{c1}}{L_{ac}} \tag{12}$$

$$\frac{dv_{c1}}{dt} = -\frac{i_{Lac}}{C_1} \tag{13}$$

Second operation stage.

$\gamma = 0 \;\; \Rightarrow$ *Q1 and D1 ON; Q2 and D2 OFF* .

The second stage operation of the circuit is described by (14), (15) and (16).

$$\frac{di_{Ldc}}{dt} = \frac{-v_{c1} + V_{dc}}{L_{dc}} \tag{14}$$

$$\frac{di_{Lac}}{dt} = \frac{v_{c1}}{L_{ac}} - \frac{v_{ac} + v^{*}_{c2}}{L_{ac}} \tag{15}$$

$$\frac{dv_{c1}}{dt} = \frac{i_{Ldc} - i_{Lac}}{C_1} \tag{16}$$

Defining $\bar{\gamma} = 1 - \gamma$ and grouping the expressions (11) to (16) in a matrix form, we can obtain (17). It can be written as form $\dfrac{d\mathbf{v}}{dt} = \mathbf{A} \cdot v + \mathbf{B} \cdot u$.

$$\begin{bmatrix} \dfrac{di_{Lac}}{dt} \\ \dfrac{dv_{c1}}{dt} \\ \dfrac{di_{Ldc}}{dt} \end{bmatrix} = \begin{bmatrix} 0 & \dfrac{1}{L_{ac}} & 0 \\ -\dfrac{1}{C_1} & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} \cdot \begin{bmatrix} i_{Lac} \\ v_{c1} \\ i_{Ldc} \end{bmatrix} + \begin{bmatrix} 0 \\ \dfrac{i_{Ldc}}{C_1} \\ -\dfrac{v_{c1}}{L_{dc}} \end{bmatrix} \cdot \bar{\gamma} + \begin{bmatrix} -\dfrac{v_{ac} + v^{*}_{c2}}{L_{ac}} \\ 0 \\ \dfrac{V_{dc}}{L_{dc}} \end{bmatrix} \tag{17}$$

III. CONTROL STRATEGY

For the original Step-up inverter, different techniques had been tested and proposed [1]: application of modulation by width pulse, using itself the model of "PWM" switch to obtain the transfer function; control with feeding direct or feed-forward and sliding mode or sliding regime.

From among the three related solutions, the last one shown the best characteristics of stability and robustness to the system. Based on this result, it is chosen to apply the sliding regime to the circuit.

The system is of third order (L_{ac}, C_1 and L_{dc}), being controlled by means of sliding regime [4]. This strategy of control was chosen due to their qualities of robustness, little sensibility to small variations of the parameters of the circuit and simplicity of implementation [5].

The C_2 capacitor is controlled through the classic control. The difference between the current i_{Lac} and its reference is feedback by the output voltage of the compensator of C_2. This way is used to diminish the average component of the current i_{Lac}, to control DC voltage in C_2.

Fig. 5 shows, by means of blocks, the form to control the converter. The circuit exposed does not present restrictions with relation to the relative values of voltages (DC and AC): DC level can be lower, equal or higher than the peak of the input voltage.

A. System control of i_{Lac}, v_{C1} and i_{Ldc}

To carry out the control of these variables the control is implemented by sliding mode. The control for sliding mode extends to the properties of the control for hysteresis for an environment multi-variable, and it is possible to force the states of the system to follow a trajectory which is located on a convenient surface in the space of the states (sliding surface). With this purpose, each one of the zones of the space of the states, separate by the sliding surface, is associated with a state of the switches [3].

Two conditions are essential so that it has success in the implementation of a sliding regimen: the condition of existence and the condition of encounter. If it relates with the capacity of the system when beginning with initial conditions in t=t_0, it can find the sliding surface (σ) in some t>t0; this refers to the maintenance of the sliding regimen after the encounter, or either, to the ability with that the system maintains the variable of state in a neighborhood enough next to σ [5]. The existence condition implies that around of σ the trajectories always must point with respect to own surface σ. Mathematically, this convergence is express by (18).

$$\lim_{\sigma\to 0^-}\left(\frac{d\sigma}{dt}\right)>0 \qquad \lim_{\sigma\to 0^+}\left(\frac{d\sigma}{dt}\right)<0 \tag{18}$$

This indicates that, next to the sliding surface, if σ will have negative value, its derivative will have to be positive and if σ will have positive value, its derivative will have to be negative, so that in any situation the representative point if approaches to the null space $\sigma = 0$ [5].

From (17) it looks for it error variation, subtracts in both sides the expression (17) the derivatives from the references from i^*_{Lac}, v^*_{c1} e i^*_{Ldc}. The error is defined as $\varepsilon = v - v^*$ and their derivatives as $\frac{d\varepsilon}{dt}=\frac{dv}{dt}-\frac{dv^*}{dt}$ and disrespecting the derivatives of the references of the right side it is obtained in (19).

$$\begin{bmatrix}\frac{d\varepsilon_{iLac}}{dt}\\ \frac{d\varepsilon_{vc1}}{dt}\\ \frac{d\varepsilon_{iLdc}}{dt}\end{bmatrix}=\begin{bmatrix}0 & \frac{1}{L_{ac}} & 0\\ -\frac{1}{C_1} & 0 & 0\\ 0 & 0 & 0\end{bmatrix}\cdot\begin{bmatrix}i_{Lac}\\ v_{c1}\\ i_{Ldc}\end{bmatrix}+\begin{bmatrix}0\\ \frac{i_{Ldc}}{C_1}\\ -\frac{v_{c1}}{L_{dc}}\end{bmatrix}\cdot\bar{\gamma}+\begin{bmatrix}-\frac{v_{ac}+v^*_{c2}}{L_{ac}}\\ 0\\ \frac{V_{dc}}{L_{dc}}\end{bmatrix} \tag{19}$$

The parameter regent of the converter is ε_{iLca} (error in the current that circulates for L_{ac}). To keep the current of the L_{ac} inductor enough near to its value of reference is the main objective of this converter. A good waveform of the i_{Lac} current and the difference of voltage in the capacitors C_1 and C_2 is expected. It fits, of this form, to the converter to diminish the possible discrepancies that appear. The error in the L_{dc} inductor is the main responsible for the determination of the instants of the commutations and ε_{vc1} is the stabilizer parameter. This parameter is indispensable, therefore in this application the reference is variable with the time.

Defined ε, as the error between the variable of control and its reference. Sliding surface is defined as σ (20).

$$\sigma = S\cdot\varepsilon = S_1\cdot\varepsilon_1 + S_2\cdot\varepsilon_2 + \ldots + S_n\cdot\varepsilon_n \tag{20}$$

Choosing S invariant with the time, it is obtained:

$$\frac{d\sigma}{dt} = S\cdot\frac{d\varepsilon}{dt} \tag{21}$$

with S_1, S_2 and S_3 >0.

$$\sigma = S\cdot\varepsilon = S_1\cdot\varepsilon_{iLac} + S_2\cdot\varepsilon_{vc1} + S_3\cdot\varepsilon_{iLdc} \tag{22}$$

Substituting the variables of (19) in (21) and evaluating it meets expression (23).

$$\begin{aligned}\frac{d\sigma}{dt} &= S_1\cdot\left[\frac{v_{c1}-v_{ac}-v^*_{c2}}{L_{ac}}\right]+S_2\cdot\left[\frac{\bar{\gamma}\cdot i_{Ldc}-i_{Lac}}{C_1}\right]\\ &\quad + S_3\cdot\left[\frac{V_{dc}-\bar{\gamma}\cdot v_{c1}}{L_{dc}}\right]\end{aligned} \tag{23}$$

State $\gamma = 1\left(\bar{\gamma}=0\right)$ is associated with the increase of the energy in the system, therefore it is always applied that the representative point this below of $\sigma = 0$; in opposed mode $\gamma = 0$ is applied when the point it is located above of the switching line. Thus the following inequations are established.

When γ=1 $\Rightarrow$ dσ/dt>0 is shown in (24).

$$S_1 < \frac{L_{ac}}{v_{c1}-v_{ac}-v^*_{c2}}\cdot\left[S_2\cdot\frac{i_{Lac}}{C_1}-S_3\cdot\frac{V_{dc}}{L_{dc}}\right] \tag{24}$$

When γ=0 $\Rightarrow$ dσ/dt<0 is shown in (25).

$$S_1 < \frac{L_{ac}}{v_{c1}-v_{ac}-v^*_{c2}}\cdot\left[S_2\cdot\frac{\left(i_{Lac}-i_{Ldc}\right)}{C_1}-S_3\cdot\frac{\left(v_{c1}-V_{dc}\right)}{L_{dc}}\right] \tag{25}$$

It is considered initially that the L_{ac} inductor behaves as a current source i_{Lac} in order to determine the limit for α. It is Defined as the reason between S_2 and S_3, with $Z_n = \sqrt{L_{dc}/C_1}$. As in this in case S_1=0, can be calculated the value of α and the restriction for S_1.

$$\gamma = 1 \;\Rightarrow\; \alpha < \frac{V_{dc}}{i_{Lac}\cdot Z_n^2} \tag{26}$$

$$\gamma = 0 \;\Rightarrow\; v_{c1} > V_{dc} + máx\left\{\alpha\cdot Z_n^2\cdot\left(i_{Ldc}-i_{Lac}\right),0\right\} \tag{27}$$

The S_1 coefficient is finally defined, considering the maximum variation of $\left|v_{Lac}\right| = \left|v_{c1}-v_{ac}-v_{c2}\right|$

$$S_1 \cdot |v_{Lac}| < L_{ac} \cdot min\left\{\left[S_3 \cdot \frac{V_{dc}}{L_{dc}} - S_2 \cdot \frac{i_{Lac}}{C_1}\right], \left[S_3 \cdot \frac{(V_{dc} - v_{c1})}{L_{dc}} - S_2 \cdot \frac{(i_{Ldc} - i_{Lac})}{C_1}\right]\right\} \quad (28)$$

The switching frequency for the sliding regime depends on the band of hysteresis used in the comparison of straight line σ with zero level.

$$f_{cd}(t) = \frac{d(t)}{\Delta\sigma} \cdot \left[S_3 \cdot \frac{V_{dc}}{L_{dc}} - S_2 \cdot \frac{i^*_{Lac}(t)}{C_1}\right] \quad (29)$$

It is noted that the increase of the duty cycle as of the value of the AC current contribute positively for the increase of the frequency.

So that the value of $f_{cd_{min}}$ really is verified, the band of hysteresis of the comparator must be chosen of appropriate form:

$$\Delta\sigma \le \frac{d_{min}}{f_{cd_{min}}} \cdot \left[\frac{S_2 \cdot V_{dc}}{L_{dc}} - S_1 \cdot \frac{I^*_{Lacp}}{C_1}\right] \quad (30)$$

B. Control of C_2

In this case the classic control is used, through the theorem of the average value. It is substituted source V_{dc} and the L_{dc} inductor for constant current sources. The transfer function of the current i_{Lac} in function of the voltage in the C_2 capacitor is shown in (31).

$$\left.\frac{\hat{i}_{Lac}(s)}{\hat{v}_{c2}(s)}\right|_{\substack{\hat{v}_{ac}(s)=0 \\ \hat{v}_{c1}(s)=0}} = \frac{1}{L_{ac} \cdot s} \quad (31)$$

Therefore, the variation of the current i_{Lac} with respect to the voltage of the C_2 capacitor only depends on the filtering inductance of AC-side.

For this control the response it must be very slow (almost it continues), given that the voltage v_{c2}(t) of reference is a DC signal. As the transfer function already is an integrator, it can be used a proportional control, but a proportional integral control with filter for a better response will be used. The cut-off frequency chosen is near 10 Hz to obtain a slow response.

IV. DESIGN AND SIMULATIONS

The converter in study has the following electrical characteristics: $V_{dc}=100$ V; $V_{ac}=155$ V; output power $P=500$ W; minimum switching frequency 15 kHz. In agreement with established criteria the minimum voltage in the C_2 capacitor must be 255 V(2). A safety margin of 45 V is applied, fixing the voltage of the C_2 capacitor in 300 V. The C_1 capacitor is calculated by (32) and the L_{dc} inductor by (33).

$$C_1 \ge \frac{I^*_{Lacp} \cdot d_{min}}{\Delta v_{c1_{max}} \cdot f_{cd_{min}}} = 15\mu F \quad (32)$$

$$L_{dc} \ge \frac{V_{dc} \cdot d_{min}}{\Delta I_{Ldc_{max}} \cdot f_{cd_{min}}} = 1.1mH \quad (33)$$

Following are showed the results obtained through numerical simulations. Considering S_3=0.244, S_2=0.022 and S_1=0.13. Change the operation mode it is simply necessary to change the reference signal of the current i_{Lac} and keeping the other unchanged parameters. Two high pass filters are used: for i_{Ldc}, filter of second order, f_{pa}=1 kHz, ξ=0.7; for V_{c1}, filter of second order, f_{pa}=1 kHz, ξ=0.7.

The results obtained in the operation as inverter are presented in the following figures. The voltages in the two capacitors are presented initially ($v_{c1}(t)$ and $v_{c2}(t)$), in Fig. 6. The current in the inductor is shown in Fig. 7. The current and voltage in the AC side is shown in Fig. 8, this present a THD≈3.27% he current i_{Lac}, taking into account the firsts fortieth-ninth harmonics.

Inverting the signal of reference for i_{Lac}, the system starts to work as rectifier. The main results obtained by numerical simulations for the nominal load condition (2.5 kW) are shown in the figures that follow. Fig. 9 shown the voltage in the two capacitors. The current in the inductor is shown in Fig. 10. The current and voltage in the AC side is shown in Fig. 11, this presents a THD≈2.04% he current i_{Lac}, taking into account the firsts fortieth-ninth harmonics.

V. CONCLUSION

The presented converter is a new contribution to the family of the single-phase rectifiers with high power factor. This presents some advantages with respect to the already existing converters, these are:

- It operates, as rectifier or inverter, with the use of only a cell of conventional switching.
- It is bidirectional in current.
- The voltage output can be lower, equal or greater than the peak of the input voltage.
- The converter is able to maintain the input current, i_{Lac}, very close to the imposed sine reference (in phase with the input voltage), achieving a power factor next to the unit.

The control technical used shows some inconveniences because their nature is a control for hysteresis, the switching frequency is variable and depends on the operation point and the design of the control parameters that can be complex. That is compensated for a practical easy implementation.

The studied topology can find a large field of application in the generation of energy from sources of DC voltage (photovoltaic panels, for example). In the operation as rectifier, they offer a solution to connect it to the power system, without degrading the power factor, systems that demand DC voltage, especially in the cases where the levels of these are lower to the value of peak of the sinusoidal voltage available.

VI. ACKNOWLEDGMENT

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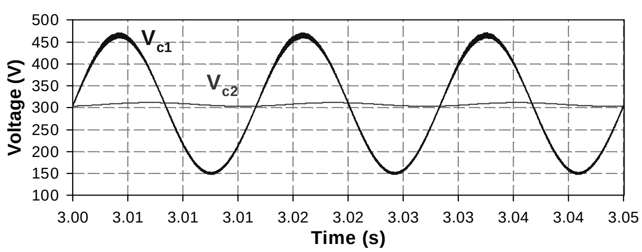

Fig. 6. Voltage on the capacitors C_1 and C_2, in the operation as inverter.

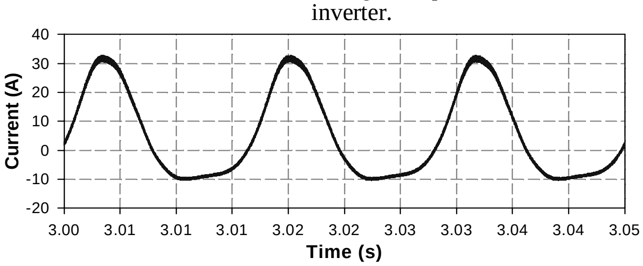

Fig. 7. Circulating current in the inductor L_{dc}, in the operation as inverter.

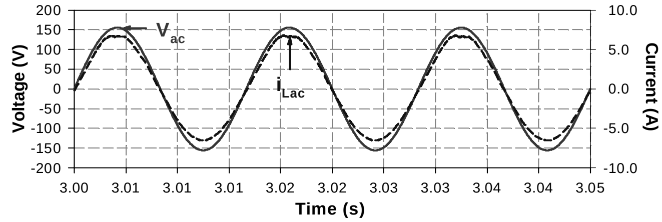

Fig. 8. Voltage (v_{ac}(t)) and current (i_{Lac}) in the AC side, in the operation as inverter.

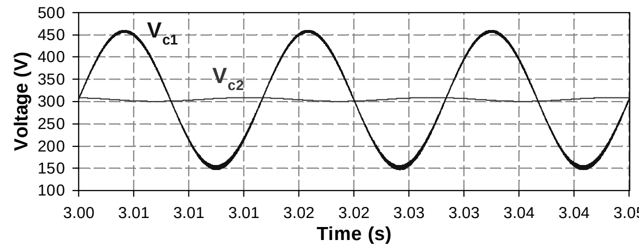

Fig. 9. Voltage on the capacitors C_1 and C_2, in the operation as rectifier.

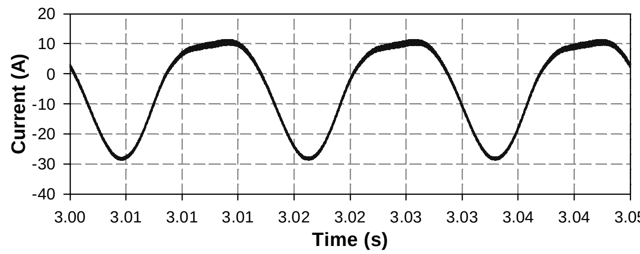

Fig. 10. Circulating current in the inductor L_{dc}, in the operation as rectifier

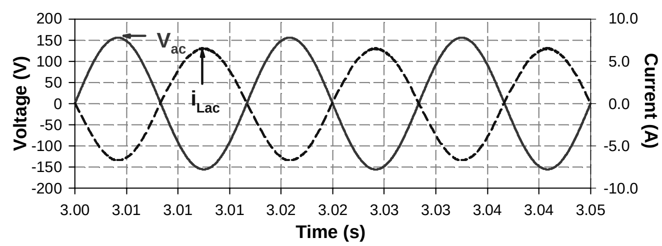

Fig. 11. Voltage ($v_{ac}(t)$) and current (i_{Lac}) in the AC side, in the operation as rectifier.