AC-DC THREE-PHASE REVERSIBLE CONVERTER WITH HIGH POWER FACTOR CONTROLLED BY SLIDING REGIME

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***Abstract* – This paper presents a three-phase converter that allows the operation as rectifier or inverter without any change the electric parameters. This reversion can be obtained simply by inverting of the reference of the AC currents. The system is analyzed as a connection of three independent lower order subsystems, controlled by sliding mode with decentralized switching scheme. The output voltage can be lower, equal or greater than the peak of the input voltage. The operation stages, equations, control strategy and design of the converter are presented. Finally, simulation results are shown for operation as inverter and rectifier.**

***Keywords* - AC–DC power conversion, DC–AC power conversion, power quality, power supplies, variable structure systems.**

I. INTRODUCTION

Most of the three-phase topologies that allow operating as rectifier or inverter 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 step-down converter, to achieve the voltage magnitude required. Among the single-phase 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], [3]. 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 be fed by an output voltage with any value and shape. Besides that, 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.

In this way, a sinusoidal output voltage can be obtained, for example, by keeping fixed the voltage across one capacitor and imposing a sinusoidal variation on the other one [9], [10], or by using two sinusoidal reference with a phase-shift of 180° between them, as shown in Fig. 1.

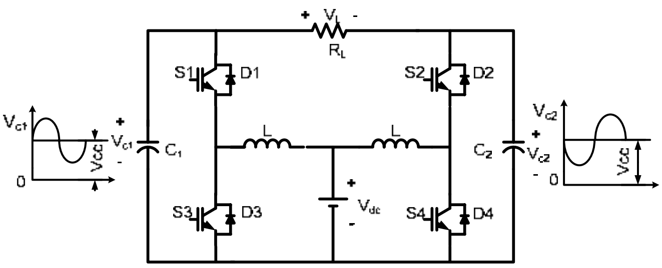


Fig. 1. Circuit of the inverter proposed by Cáceres and Barbi [1].

Each converter produces a unipolar sinusoidal output voltage with a continuous component, as shown by the waveforms of V_{c1} and V_{c2} in Fig. 1. The load is connected in differential mode between the converters, annulling the continuous components.

The modulation in each converter is 180° phase-shifted in relation to the other, this maximizes the excursion of voltage across 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 three-phase topology of this inverter (step-up inverter), proposed by Romaneli [4] is shown in Fig. 4.a.

Colling and Barbi [5], [7], [8] proposed the single-phase circuit operating as rectifier. For the rectifying operation, it is necessary to invert the direction of the power flow and to establish some control on the AC current absorbed from the source. Therefore, an inductor in series with the AC source is included, as shown in Fig. 2. Due to that reversibility of power flow, the elements (inductors and sources) are not identified as input or output anymore, but as AC-side or DC-side.

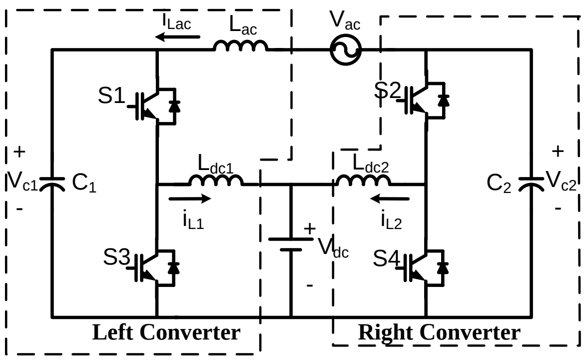


Fig. 2. Rectifier diagram proposed by Colling and Barbi [5], [6].

Fuentealba and Barbi [9],[10] proposed a variation of the topology presented by Colling and Barbi, by removing the right converter and replacing it by a capacitor of great capacity. In this way, it is possible to obtain a converter with a cell of commutation operating like inverter or rectifier as shown in Fig. 3.

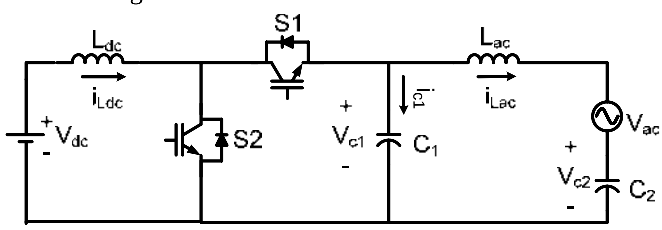
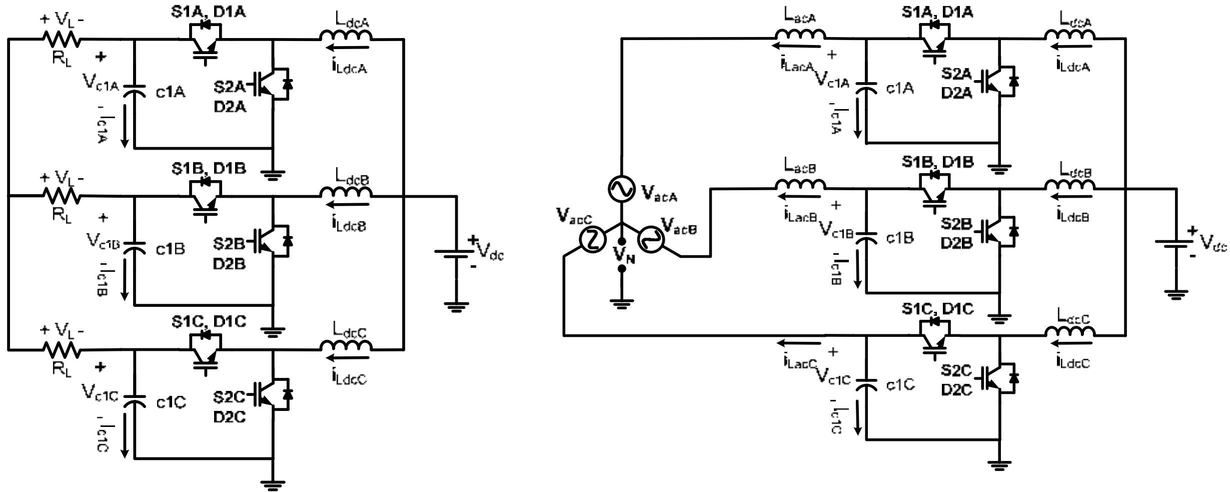


Fig. 3. Rectifier proposed by Fuentealba and Barbi [9], [10].



(a) (b)

The three-phase converter proposed by Colling and Barbi is shown in Fig. 4.b. It is based on the structure of the Fig. 2.

This paper proposes the model, strategy of the control and simulation of the three-phase topology derived of single-phase topology presented by Fuentealba and Barbi [9],[10].

II. TOPOLOGY PROPOSAL

The original single-phase topology operating as rectifier is shown in Fig. 3. It is the basis to obtain the three-phase structure. One such converter is used in each phase, regulating its control variables, so that a set of three converters enables the operation in the three-phase system.

Before beginning the analysis some initial conditions must be considered as for example, the operation of the circuit demands a DC level in such a way that its voltages never be inferior to V_{dc}. Therefore, the function of each converter is to keep to keep a sinusoidal voltage plus an average component (V_{c2A}) on the capacitors c1A, c1B, and c1C.

In this way, the voltage on the capacitor c1A will be $V_{acA}+V_{c2A}$, and in a similar manner on c1B and c1C. They also adjust the values of the currents i_{LacA}, i_{LacB} and i_{LacC} to obtain the power factor correction.

Therefore, the structure operating as inverter is similar to a step-down converter with a filter at the output, the difference being that their input voltages have an alternating component above a DC voltage level.

In order to initiate the analysis of the topology, the three-phase system is separated in three subsystems of lower order, associated to each one associated to one of the phases at the input. It is only analyzed here the circuit associated to phase A, since the others operate in a similar way, as shown Fig. 8.

A. Considerations for the operation of the circuit of phase A

The circuit is considered as a voltage step-up inverter. For the capacitor c2A, its voltage must be (1).

$$v^*_{c2A}(t) = V^*_{c2_{dc}} \tag{1}$$

The total sinusoidal excursion applied to the capacitor c1A, leaving c2A only with a DC voltage level (contrary to the proposed by Cáceres and Barbi in [1], [2], [3], where the AC voltage is divided between capacitors C_1 and C_2). Thus, the condition below must be satisfied with a margin of security so that, never V_{c2A} is lower than V_{dc} (2)

$$V^*_{c2A_{dc}} > V_{dc} + V_{acAp} \tag{2}$$

Therefore,

$$v^*_{c2A}(t) = \left(V_{dc} + V_{acAp}\right) + \Delta V_{dc} \tag{3}$$

With ΔV_{dc} as the security voltage margin. It is considered that the AC mains voltage is $V_{acAp} \cdot \sin(\omega t)$. The sum of the Ac voltage and the voltage across c2A is:

$$v_{acA}(t) + v^*_{c2A}(t) = V_{acAp} \cdot \sin(\omega t) + \left(V_{dc} + V_{acAp} + \Delta V_{dc}\right) \tag{4}$$

therefore, the voltage in the c2A capacitor can be a DC voltage with minimum magnitude, definite by (2).

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

$$i^*_{LacA}(t) = I^*_{LacAp} \cdot \sin(\omega t + \phi_A) \tag{5}$$

L_{acA} 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).

DC level applied in c2A also is established in the c1A capacitor.

$$v_{c1A}(t) = V_{acAp} \cdot \sin(\omega t) + \left(V_{dc} + V_{acAp} + \Delta V_{dc}\right) \tag{6}$$

The duty cycle of the switch S2A is given by (7).

$$d(t) = 1 - \frac{V_{dc}}{V^*_{c2A_{dc}} + V_{acAp} \cdot \sin(\omega t)} \tag{7}$$

Considering the energy balance, the current through the inductor L_{dcA} can be calculated by (8).

$$i_{LdcA}(t) = \frac{V_{c1A}(t) \cdot \left(i_{LacA}(t) + i_{c1A}(t)\right)}{V_{dc}} \tag{8}$$

In steady state, this yields equation by (9).

$$i_{LdcA}(t)=\frac{V^*_{c2A_{dc}}\cdot I^*_{LacAp}\cdot\sin(\omega t+\phi_A)}{V_{dc}}+\frac{I^*_{LacAp}\cdot V_{acAp}}{2}\cdot\frac{(1-\cos(2\omega t))}{V_{dc}}+\omega\cdot c1A\cdot\left\{\frac{V^*_{c2A_{dc}}}{V_{dc}}\cdot V_{acAp}\cdot\cos(\omega t)+\frac{V^2_{acAp}}{2\cdot V_{dc}}\cdot\sin(2\omega t)\right\}\qquad(9)$$

The current i_{LdcA} is described by (9). It still has the excursions in high frequencies caused for the switching of the switches S1A and S2A. 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.

$$P=V_{dc}\cdot I_{LdcA_{med}}=V^*_{c2A_{acAp}}\cdot I^*_{LacAp}\qquad(10)$$

In the operation as rectifier all the terms with i_{LacA} (or I^*_{LacAp}) are inverted.

B. Analysis of the circuit

The switches S1A and S2A receive complementary pulses to prevent the discontinuous conduction in the inductor L_{dcA}. Due to this complementarity, the existence of only two structures is possible, which can be extracted from the diagram presented in Fig. 5. When S2A and D2A are turned on and S1A and D1A are turned off one has the first stage of operation ($\gamma = 1$). The second stage occurs when S1A and D1A are turned on and S2A and D2A are turned off ($\gamma = 0$).

I_{Lac} is assumed to be at its reference value and the state of the switches is described by variable γ. Defining $\bar{\gamma}=1-\gamma$ and grouping the expressions obtained from Fig. 5 in a matrix form, written as $d\mathbf{v}/dt=\mathbf{A}\cdot v+\mathbf{B}\cdot u$, it can be obtained (11).

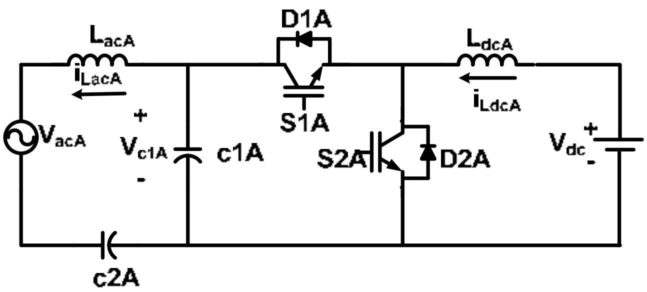


Fig. 5. Circuit phase A, inverter/rectifier proposed.

$$\begin{bmatrix}\frac{di_{LacA}}{dt}\\ \frac{dv_{c1A}}{dt}\\ \frac{di_{LdcA}}{dt}\end{bmatrix}=\begin{bmatrix}0&\frac{1}{L_{acA}}&0\\ -\frac{1}{c1A}&0&0\\ 0&0&0\end{bmatrix}\cdot\begin{bmatrix}i_{LacA}\\ v_{c1A}\\ i_{LdcA}\end{bmatrix}+\begin{bmatrix}0\\ \frac{i_{LdcA}}{c1A}\\ -\frac{v_{c1A}}{L_{dcA}}\end{bmatrix}\cdot\bar{\gamma}+\begin{bmatrix}-\frac{v_{acA}+v^*_{c2A}}{L_{acA}}\\ 0\\ \frac{V_{dc}}{L_{dcA}}\end{bmatrix}\qquad(11)$$

III. CONTROL STRATEGY

The system is of third order (L_{acA}, c1A and L_{dcA}), being controlled by means of sliding regime [6], [8], [10] This strategy of control was chosen due to its qualities of robustness, little sensibility to small variations of the electric parameters of the circuit and simplicity of implementation. The voltage across C2A is controlled through the classic control. The difference between the current i_{LacA} and its reference is feedback by the output voltage of the compensator of c2A this way is used to diminish the average component of the current i_{LacA}, to control DC voltage in c2A.

Fig. 6 shows schematically the way the control of the converter is established.

The circuit exposed does not present restrictions with relation to the relative values of voltages (DC and AC): The output voltage can be lower, equal or higher than the peak of the input voltage. In addition, this circuit is able to work as rectifier or inverter, only with the direction change of the three reference currents (i^*_{LacA}, i^*_{LacB} and i^*_{LacC}).

A. Control System of i_{LacA}, v_{C1A} and i_{LdcA}

The control of the third-order system defined by i_{LacA}, v_{C1A} and i_{LdcA} is represented by the sliding surface of the phase A. This is defined by the equation (12), with S_1, S_2 and S_3 >0.

$$\sigma_A=S\cdot\varepsilon_A=S_{1A}\cdot\varepsilon_{iLacA}+S_{2A}\cdot\varepsilon_{vc1A}+S_{3A}\cdot\varepsilon_{iLdcA}\qquad(12)$$

Initially, it is considered that the inductor L_{acA} behaves as a current source i_{LacA} in order to determine de limit for α_A, which is defined as the reason. between S_{2A} and S_{3A}. As in this case S_{1A} = 0, both the value of α_A and the restriction for S_{1A} can be calculated. Obs. : $Z_{nA}=\sqrt{L_{dcA}/c1A}$.

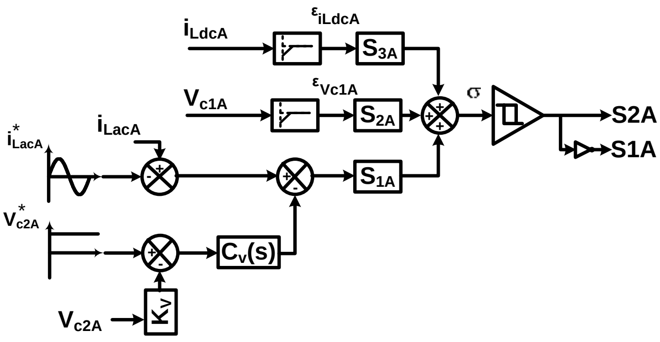


Fig. 6. Blocks diagram of the control of phase A, shown in Fig. 5.

$$\gamma=1\ \Rightarrow\ \alpha_A<\frac{V_{dc}}{i_{LacA}\cdot Z^2_{nA}}\qquad(13)$$

$$\gamma=0\ \Rightarrow\ v_{c1A}>V_{dc}+\text{máx}\left\{\alpha_A\cdot Z^2_{nA}\cdot\left(i_{LdcA}-i_{LacA}\right),0\right\}\qquad(14)$$

Coefficient S1A is finally defined after considering the maximum variation of $|v_{LacA}|=|v_{c1A}-v_{acA}-v_{c2A}|$

$$S_{1A}\cdot\left|v_{LacA}\right|<L_{acA}\cdot\min\left\{\left[S_{3A}\cdot\frac{V_{dc}}{L_{dcA}}-S_{2A}\cdot\frac{i_{LacA}}{c1A}\right],\left[S_{3A}\cdot\frac{(V_{dc}-v_{c1A})}{L_{dcA}}-S_{2A}\cdot\frac{(i_{LdcA}-i_{LacA})}{c1A}\right]\right\} \quad (15)$$

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

$$f_{cdA}(t)=\frac{d_A(t)}{\Delta\sigma_A}\cdot\left[S_{3A}\cdot\frac{V_{dc}}{L_{dcA}}-S_{2A}\cdot\frac{i^*_{LacA}(t)}{c1A}\right] \quad (16)$$

It can be noted that the increase of the duty cycle, as well as the increase of the value of the AC current, contribute positively for augmenting the switching frequency. Consequently, a minimum value for the switching frequency can be used as a parameter for choosing an appropriate hysteresis bandwidth, as expressed by (17).

$$\Delta\sigma_A\le\frac{d_{\min A}}{f_{cd_{\min}A}}\cdot\left[\frac{S_{2A}\cdot V_{dc}}{L_{dcA}}-S_{1A}\cdot\frac{I^*_{LacAp}}{c1A}\right] \quad (17)$$

B. Control of V_{c2A}

In this case, the classical control is used, through the theorem of the average value. V_{dc} and i_{LdcA} are replaced by constant sources. The transfer function of the current i_{LacA} as a function of the voltage of capacitor c2A is given by (18).

$$\left.\frac{\hat{i}_{LacA}(s)}{\hat{v}_{c2A}(s)}\right|_{\substack{\hat{v}_{acA}(s)=0\\ \hat{v}_{c1A}(s)=0}}=\frac{1}{L_{acA}\cdot s} \quad (18)$$

Therefore, the variation of the current i_{LacA} with respect to the voltage of the capacitor c2A depends only on the filtering inductance at the AC-side.

For this controller the response must be slow (almost continuous), given that the reference voltage for v_{c2A} is a DC signal. As the transfer function already is an integrator, it can be used a proportional control, but a proportional control would serve but, in order to achieve a better response, a proportional-integral control with filtering is actually used. The cut-off frequency chosen is near 10 Hz to obtain a slow response.

C. Three-Phase system

As already mentioned, the methodology for designing the converters for phases B and C is the same as that applied for phase A, since each converter configures a single-phase unit, composing the three-phase system when they operate together as a whole. In this way, the analysis shown in the previous item for phase A is applied without modifications to the other phases.

The voltages imposed on the capacitors c1A, c1B, and c1C, which allow a proper system operation, are given by equations (19).

$$\begin{aligned}v_{c1A}(t)&=V_{caAp}\cdot\sin(\omega t)+v^*_{c2A}\\ v_{c1B}(t)&=V_{caBp}\cdot\sin\left(\omega t-\frac{2\pi}{3}\right)+v^*_{c2B}\\ v_{c1C}(t)&=V_{caCp}\cdot\sin\left(\omega t-\frac{4\pi}{3}\right)+v^*_{c2C}\end{aligned} \quad (19)$$

The reference currents imposed in the AC side for each of the phases are expressed by (20).

$$\begin{aligned}i^*_{LcaA}(t)&=I^*_{LcaAp}\cdot\sin(\omega t)\\ i^*_{LcaB}(t)&=I^*_{LcaBp}\cdot\sin\left(\omega t-\frac{2\pi}{3}\right)\\ i^*_{LcaC}(t)&=I^*_{LcaCp}\cdot\sin\left(\omega t-\frac{4\pi}{3}\right)\end{aligned} \quad (20)$$

Equations (20) are applicable when the circuit functions as inverter. For the operation as rectifier, the currents are represented by the equation (21).

$$\begin{aligned}i^*_{LcaA}(t)&=I^*_{LcaAp}\cdot\sin(\omega t+\pi)\\ i^*_{LcaB}(t)&=I^*_{LcaBp}\cdot\sin\left(\omega t+\frac{\pi}{3}\right)\\ i^*_{LcaC}(t)&=I^*_{LcaCp}\cdot\sin\left(\omega t-\frac{\pi}{3}\right)\end{aligned} \quad (21)$$

It is important to notice that the phase-shifts of voltages and currents depend on the sequence that is adopted when the three-phase system is connected. In the present case, ABC sequence is used, as shown Fig. 7.

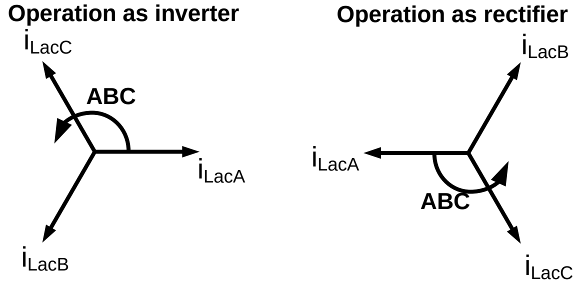


Fig. 7. Fasorial diagram of the reference current in the AC side, with ABC sequence.

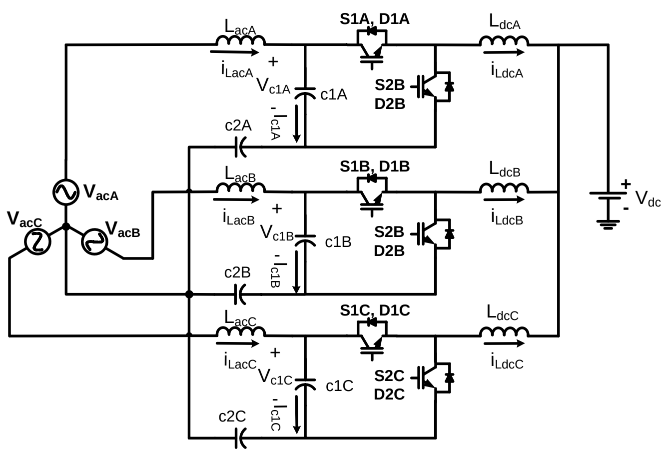


Fig. 8. Three-Phase Rectifier proposed by Fuentealba and Barbi

III. DESIGN AND SIMULATION

The converter in study has the following electrical characteristics: $V_{dc} = 100$ V; $V_{acp} = 155$ V; output power $P = 1000$ W; minimum switching frequency is 27 kHz. In agreement with the established criteria, the minimum voltage in the c2A, c2B and c2C capacitors must be 310 V.

The electric parameters to converter A, B and C are the following:

- $L_{acA}=L_{acB}=L_{acC}=450$ µH, $L_{dcA}=L_{dcB}=L_{dcC}=130$ µH;
- c1A=c1B=c1C=5.6µF; c2A=c2B=c2C=4.7mF,
- $S_{1A}=S_{1B}=S_{1B}=0.2$ Ω, $S_{2A}=S_{2B}=S_{2C}=0.022$ V/V and $S_{3A}=S_{3B}=S_{3C}=0.07$ Ω.

The hysteresis band is 0.49V. Second-order high-pass filters for i_{LdcA}, i_{LdcB} and i_{LdcC} with f_{pa}=1.2 kHz and v_{c1A}, v_{c1B} and v_{c1C} with f_{pa}=1.2 kHz. The duty cycle is between 0.351 to 0.785.

The signals corresponding to v_{c2A}, v_{c2B}, and v_{c2C}, after being attenuated, are compared to fixed references of 3.1V. The resulting errors are compensated by a PI controller with filtering. Its transfer function is presented in equation (22).

$$C_{SI}(s) = -\frac{R_{2i}\cdot C_{2i}\cdot s+1}{s\cdot R_{1i}\cdot\left(C_{2i}+C_{1i}\right)\cdot\left[\left(\frac{R_{2i}\cdot C_{2i}\cdot C_{1i}}{C_{2i}+C_{1i}}\right)\cdot s+1\right]} \quad (22)$$

The results obtained in the operation as inverter are presented in the following figures. The voltages in the three capacitors are presented initially, in Fig. 9. The currents in the DC inductors are shown in Fig. 10. The currents in the AC inductors are shown in Fig. 11. The current and voltage in the AC side, phase A, are shown in Fig. 12. Table I, shown the total harmonic distortion of the AC currents in the operation as inverter. This analysis was made taking into account the first forty-nine harmonics.

TABLE I
Total harmonic distortion, for operation as inverter

	Style		
	Phase A	**Phase B**	**Phase C**
Current (A)	4,36/0,56°	4,35/-120°	4,36/-240°
True RMS	3,03	3,03	3,03
THD	2,83%	1,04%	2%

Inverting the signals of reference for i_{LacA}, i_{LacB} and i_{LacC} the system starts to work as rectifier. The main results obtained by numerical simulations for the nominal load condition (1 kW) are shown in the figures that follow. Fig. 13 shown the voltages in the three capacitors. The currents in the DC inductors are shown in Fig. 14. The currents in the AC inductors are shown in Fig. 15. The current and voltage in the AC side, phase A, are shown in Fig. 16. Table II shows numerical results from the operation as rectifier. As before, this analysis was also made taking into account the first forty-nine harmonics.

TABLE II
Total harmonic distortion, for operation as rectifier

	Style		
	Phase A	**Phase B**	**Phase C**
Current (A)	4,63/-179,2°	4,63/60,5°	4,62/-59,5°
True RMS	3,14	3,14	3,13
THD	3%	2,44%	2,7%

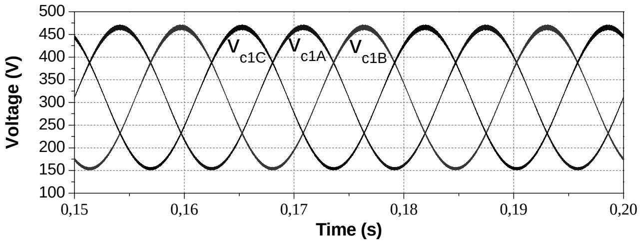


Fig. 9. Voltage on the capacitors v_{c1A}, v_{c1B} and v_{c1C} in the operation as inverter.

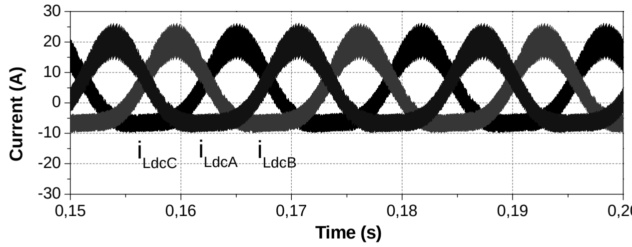


Fig. 10. Circulating currents in the inductors L_{dcA}, L_{dcB} and L_{dcC} in the operation as inverter.

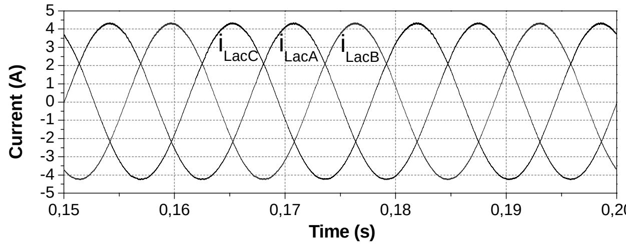


Fig. 11. Circulating currents in the inductors L_{acA}, L_{acB} and L_{acC} in the operation as inverter.

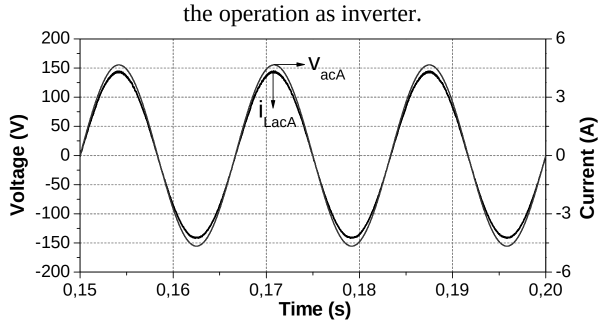


Fig. 12. Voltage (v_{acA} (t)) and current (i_{LacA}) in the AC side (Phase A), in the operation as inverter.

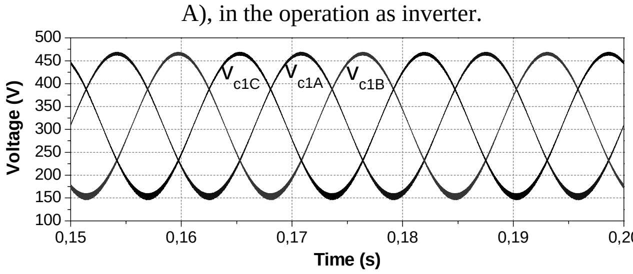


Fig. 13. Voltage on the capacitors v_{c1A}, v_{c1B} and vc_{1C} in the operation as rectifier.

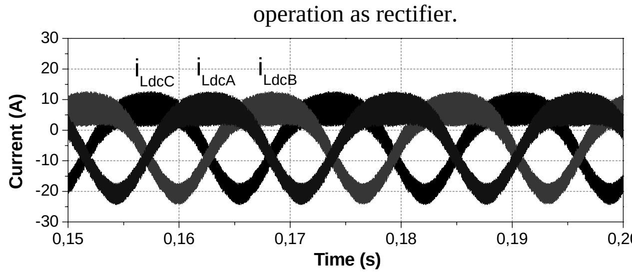


Fig. 14. Circulating current in the inductor L_{dcA}, L_{dcB} and L_{dcC} in the operation as rectifier.

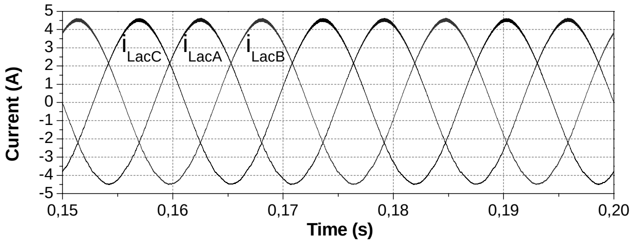


Fig. 15. Circulating current in the inductor L_{acA}, L_{acB} and L_{acC} in the operation as rectifier.

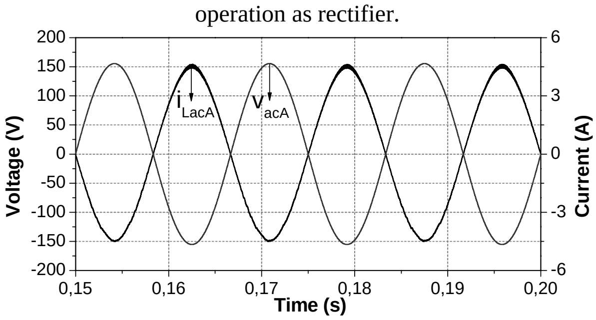


Fig. 16. Voltage ($v_{acA}(t)$) and current (i_{LacA}) in the AC side (Phase A), in the operation as rectifier.

IV. CONCLUSION

The presented converter is a new contribution to the family of the three-phase rectifiers with high power factor. It presents some advantages when compared to the already existing converters, among which one can mention::

- It is bidirectional in current.
- The output voltage can be lower, equal or greater than the peak of the input voltage.
- The converter is able to maintain the input currents, i_{LacA}, i_{LacC} and i_{LacB} very close to the imposed sine reference (in phase with the input voltage), achieving a power factor next to the unit.

The control technique used shows some inconveniences because its nature is hysteretical, the switching frequency is variable and depends on the point of operation, and the design of the control parameters can become complex. On the other hand, it is compensated by a practical, easy implementation.

The studied topology can find a large field of application in the generation of energy from DC voltage sources (photovoltaic panels, for example). In the operation as rectifier, it offers a solution to the connection of loads that demand DC voltages to the three-phase system, without degrading the power factor. This is especially true if the levels of the required DC voltages at the output are lower than the peak value of the available sinusoidal mains voltage.

V. ACKNOWLEDGEMENT

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VI. REFERENCES

[1] Cáceres Agelviz R. O.; Barbi, I. (1995). “A boost DC-AC converter: operation, analysis, control, and experimentation,” In: Intern. Conf. on Ind. Electron., Control, and Instrumentation – IECON (1995), Orlando, EUA. Anais. Piscataway, 1995. v.1,p.546-551.

[2] Cáceres Agelviz R. O.; Barbi, I. (1999). “A boost DC-AC converter: operation, analysis, control, and experimentation,” IEEE Trans. On Power Electron., New York, EUA. 1999. v.14 ,p.134-141.

[3] Cáceres Agelviz. R. O (1997). “CA-CC família de conversores, derivado de conversores básicos CC-CC,” Tese (Doutorado em Engenharia Elétrica). Instituto de Eletrônica de Potência – Universidade Federal de Santa Catarina.

[4] Romaneli E. (1998). “Inversor trifásico Boost. Teoria, equações, simulação, e experimentação.” Disertação de Engenharia Elétrica. Instituto de Eletrônica de Potência – Universidade Federal de Santa Catarina.

[5] Colling I.E.; Barbi, I. “A reversible step-up voltage-source inverter controlled by sliding mode,” IEEE Power Electronics Specialists Conference – PESC (30. June – July 1999: Charleston, Carolina do Sul, EUA). Anais. Piscataway, 1999. v.1, p.538-543.

[6] Colling I.E.; Barbi, I. “Conversor CA-CC trifásico reversível com elevado fator de potência e controle por regime de deslizamento,” Congresso Brasileiro de Automática – CBA (13. Set. 2000: Foz do Iguazu). Anais. Curitiba:CBA. p.711-716.

[7] Colling I.E.; Barbi, I. (2001). “Reversible Unity Power Factor Step-Up/Step-Down AC–DC Converter Controlled by Sliding Mode,” IEEE Trans. On Power Electron., 2001. v.16, No.2 ,p.223-230.

[8] Colling I.E. (2000). “ Conversores CA-CC monofásicos e Trifásicos reversíveis com elevado fator de potência”. Tese (Doutorado em Engenharia Elétrica). Instituto de Eletrônica de Potência – Universidade Federal de Santa Catarina.

[9] Fuentealba E., Barbi I. “Conversor CA-CC buck bidirecional monofásico com elevado fator de potência,” Congresso Brasileiro de Automática – CBA (3. Out. 2006: Salvador-BH). Anais. Curitiba:CBA. p.162-167.

[10] Fuentealba E., Barbi I. “AC-DC Bidirectional single-phase step-down converter with high power factor”. 32th Annual Conference of the IEEE Industrial Electronics Society (IECON’06). ISBN: 1-4244-0136-4. November 6. Paris 2006.

Three-Phase Series-Buck Rectifier with Split DC-bus Based on the Scott Transformer

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Abstract: **In this paper, a new unity power factor isolated three-phase buck rectifier is presented. Based on the Scott transformer, this rectifier is simple and it has the capability to obtain low output voltage. Besides, it protects against short circuit and it needs no auxiliary circuit for inrush current. Using only two active switches, it is able to generate symmetrical currents in the line and output voltage regulated. The modulation is used conventional SPWM. The control has only one voltage control loop. Theoretical analysis, design procedure, complete simulation results with closed loop operation are given, as well as results of an experimental verification.**

I. INTRODUCTION

Looking for improve the energy quality in the distribution networks and efficiency, it were created strict current harmonic limitation imposed on power supplies [1]. Therefore, rectifier with power factor correction are researched and used for the developments of power supplies in recent years.

In this paper, the unity power factor three-phase rectifier buck with a simplified control loop technique, based on the Scott transformer is presented.

The single-phase PWM buck pre-regulator in fig 1 has some important characteristics such as the absence of inrush current, low DC output voltage, protection against short circuit, among others.

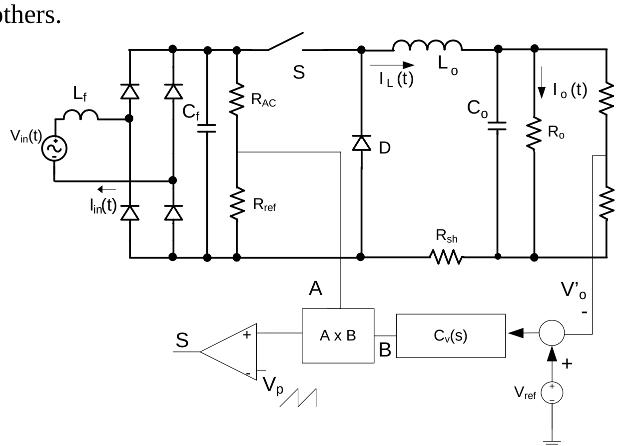

Fig. 1: Single-phase buck pre-regulator.

In continuous conduction mode, with a low frequency output inductor designed in suck a way that if behaves as a constant current source. Therefore the size and weight of the output inductor, in this case, is much bigger.

In order to optimizer the size and weight, the inductance Lo may be decreased, so it no longer behaves as a constant current source. Although, increasing the output inductor current ripple distorts the input current, with a significant third harmonic component.

In [2] proposes a control technique to eliminate the distortion on the input current even when the output inductor current presented large ripple. In reference [3] is also presented control techniques more simple to eliminate the distortion on the input current.

In [4]-[5] the unity power factor three-phase rectifier with a split DC-bus based on the Scott transformer is presented.

The proposed topology is show in the Fig. 2. In this application use two single-phase buck rectifiers in continuous conduction mode. There is a output voltage with a split DC-bus and the output inductors are coupled. It is reduced de size and cost of rectifier.

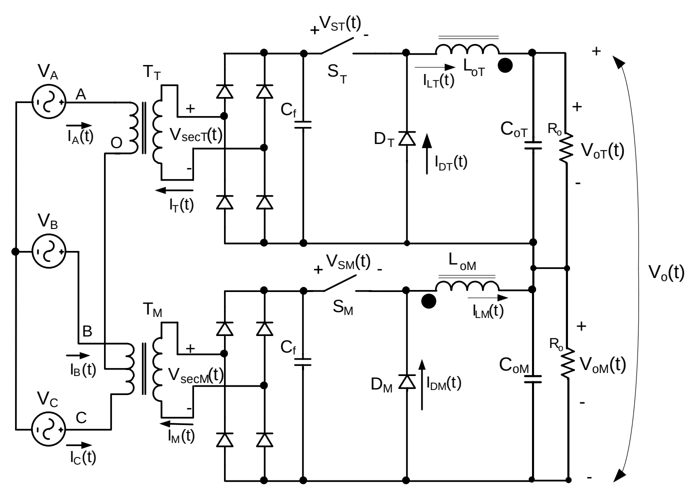

Fig. 2: Unity power factor isolated three-phase rectifier buck series.