

Isolated Three-Phase Rectifier With High Power Factor Using the Zeta Converter in Continuous Conduction Mode

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Abstract This paper presents the analysis of an Isolated Three-Phase Rectifier with high power factor using a Zeta converter operating in continuous conduction mode (CCM). The structure is particularly simple and robust. Its main features are: one power processing stage, lower harmonic distortion in the current and natural isolation. Principle of operation, design procedure and experimental results are presented.

1. Introduction

Over the past twenty years, Power Electronics has advanced to a surprising degree, both in relation to the topological variations and in the drive and control strategies of static converters.

Due to this enormous development, static converters are currently utilized in the most widely varying industrial applications, and today constitute one of the major themes for study in Electrical Engineering.

Despite this extraordinary development, the input stage of the main static converters, which utilizes a rectifying bridge to diodes, coupled with an extremely powerful capacitive filter, sometimes creates a serious problem. The association of these components generates a non-linear load that, when connected to the commercial power system, causes a serious distortion in the input current, resulting in the injection of a high harmonic content into electric power system current.

In the high power industrial applications (over 1KW), the three-phase feeding systems are generally the most recommended, where the AC-DC conversion has been dominated by conventional diode rectifiers and thyristor controlled rectifiers. The non-linear characteristic of the input current of these rectifiers, according to the above-mentioned, generate problems for the commercial network of electrical energy, among then we can point out:

- Injection of a high harmonic content into the output current;
- Distortion of the input voltage, due to the high peaks of the input current;
- Increases of losses in the transmission lines;
- Reduction of the power factor;
- Need to generate of large quantity of reactive power;

- Decrease in the structure efficiency, due to the high RMS value of the input current.

Due to those drawbacks many works has been presented by the scientific community of power electronics, with the aim of utilizing AC-DC converters with high power factor and low harmonic content of the input current [1,2,3].

Normally the power factor correction is obtained using converters that have a input current source [1], [2]. For the three-phase system this technique is used considerably. Taking for example the Boost converter [3], which at present is one of the converters used for power factor correction, we can observe the following characteristics: high output voltage, no natural isolating and one Boost inductor for each phase. Trying to overcome these basic limitations, this paper proposes the application of an isolated three-phase rectifier with high power factor using the Zeta converter in continuous conduction mode. The structure is particularly simple and robust. Its main features are: one power processing stage, lower harmonic distortion in the line current, input current system which does not require there input inductors, natural isolation and current source characteristic in the output, facilitating parallel connection, and very simple circuit drive.

Besides these features, it is important to emphasize that the proposed structure permits, with little difficulty and at relatively low cost, to increase the power factor of the conventional circuits that have been used in industry, with a simple inclusion of the Zeta module between the conventional three-phase rectifier and the load, as we can see in Fig.1.

2. Principle of Operation

2.1. Proposed Circuit

The proposed circuit is shown in Fig. 1. Fig. 2 and Fig. 3 present the equivalent circuit with all the parameters reflected to the primary side of the transformer, where:

$$V_o = (N_1/N_2)V_o' ; C_o = (N_2/N_1)^2 C_o' ; C_1 = (N_2/N_1)^2 C_1' \\ L_o = (N_1/N_2)^2 L_o' ; R_o = (N_1/N_2)^2 R_o'$$

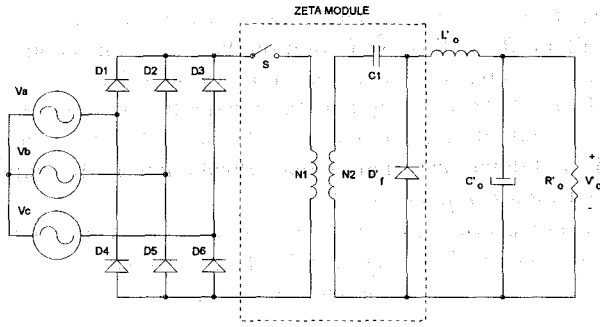


Fig 1. Proposed Circuit

2.2. Operation Stages

To simplify the analysis, the following assumptions are made:

- The operation of the circuit is steady-state;
- The semiconductors are considered ideals;
- The transformer is represented by its magnetizing inductance reflected to the primary side;
- The capacitor C_o is large enough to make its voltage constant and equal to V_o ;
- The line voltage is considered constant during a switching period.

The Zeta converter operating in continuous conduction mode has two operation stages:

1st Stage (t_0, t_1) Fig. 2: Switch S is conducting, the line source supply energy to the L_m inductance and the capacitor C_1 supplies energy to L_o . Current i_{Lm} and i_{Lo} increase linearly. Voltages V_{C0} and V_{C1} are considered constant and equal to V_o . During this stage the diode D_f continue blocked with a reverse voltage equal to $-(V_{in} + V_o)$.

2nd Stage (t_1, t_2) Fig. 3: By the time t_1 , switch S is turned off and the diode D_f starts to conduct, allowing L_m and L_o to transfer their energy to C_1 and C_o respectively. The currents i_{Lm} and i_{Co} decrease linearly. In second stage there is no energy circulation in the line, which assures us there is no harmonic distortion in the current line. The voltage across the switch S is equal to $(V_{in} + V_o)$.

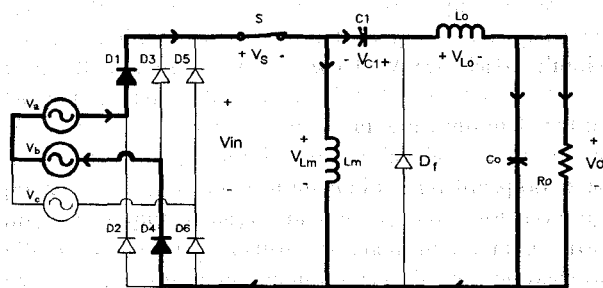


Fig 2. First Stage

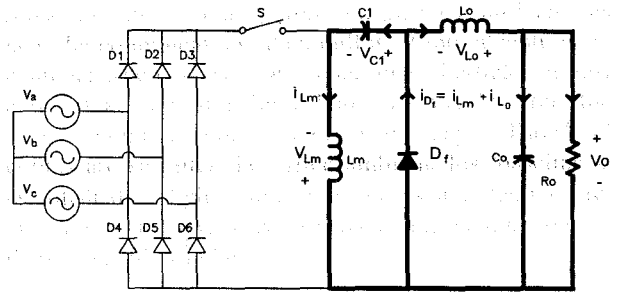


Fig 3. Second Stage

The main waveforms are shown in Fig. 4.

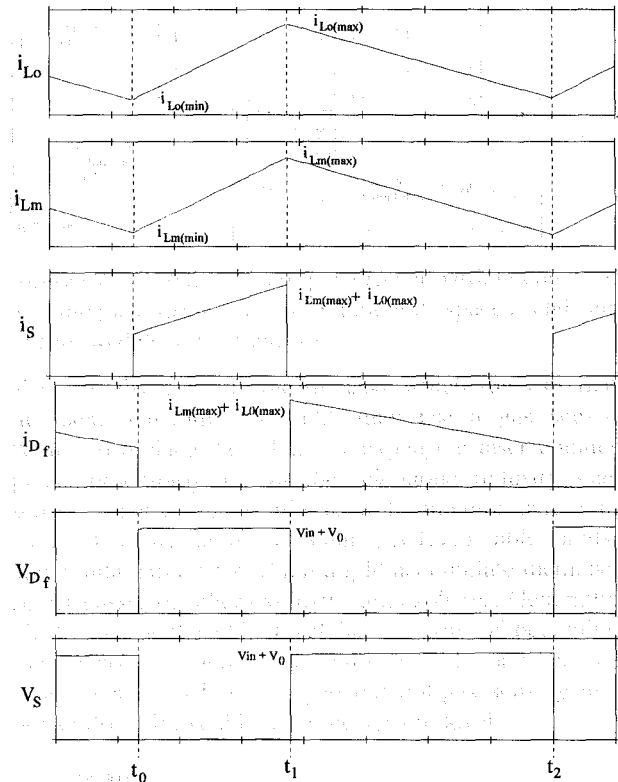


Fig 4. Main Waveforms

2.3. Quantitative Analysis

From Fig. 4 the following parameters are defined:

$$t_r = t_1 - t_0$$

$$t_a = t_2 - t_1 = T_s - t_r$$

$$T_s = 1 / f_s$$

$$D = t_r / T_s$$

$$V_{in} = \sqrt{3} \cdot V_p \cdot \sin(\omega \cdot t); \text{ for } (\omega \cdot t) \text{ varying from } \pi/3 \text{ until } 2\pi/3.$$

2.3.1) 1st Stage ($0 \leq t \leq t_1$)

Initial values: $i_{Lm}(t = 0) = i_{Lm}(min)$

$$i_{Lo}(t = 0) = i_{Lo}(min)$$

$$V_{Lm}(t) = V_{Lo}(t) = V_{in}$$

From the equivalent circuit referring to this stage we can obtain the following equations:

$$V_{L_m(t)} = L_m \cdot \frac{di_{L_m(t)}}{dt} \quad (1)$$

$$V_{L_m(t)} = V_{in} = \sqrt{3} \cdot V_p \cdot \sin(\omega \cdot t) \quad (2)$$

From Eqs. (1) and (2) we can obtain the current through the magnetizing inductance:

$$i_{L_m(t)} = \sqrt{3} \cdot \frac{V_p}{L_m} \cdot t \cdot \sin(\omega \cdot t) + i_{L_m(\min)} \quad (3)$$

For the output inductance L_0 we have:

$$V_{L_0(t)} = L_0 \cdot \frac{di_{L_0(t)}}{dt} \quad (4)$$

$$V_{L_0(t)} = V_{in} \quad (5)$$

Thus, following the same procedure we obtain:

$$i_{L_0(t)} = \sqrt{3} \cdot \frac{V_p}{L_0} \cdot t \cdot \sin(\omega \cdot t) + i_{L_0(\min)} \quad (6)$$

The current through the switch S is given by the sum of the currents in the inductances L_m e L_0 , resulting in:

$$i_{S(t)} = i_{L_m(t)} + i_{L_0(t)} = \frac{1}{L_{eq}} \cdot \sqrt{3} \cdot V_p \cdot t \cdot \sin(\omega \cdot t) + i_{L_m(\min)} + i_{L_0(\min)} \quad (7)$$

Where:
$$\frac{1}{L_{eq}} = \frac{1}{L_m} + \frac{1}{L_0} \quad (8)$$

2.3.2) 2nd Stage ($0 \leq t \leq t_a$)

Initial values: $i_{L_m(t=0)} = i_{L_m(\max)}$

$$i_{L_0(t=0)} = i_{L_0(\max)}$$

$$V_{L_m(t)} = V_{L_0(t)} = V_0$$

Taking pattern for the 1st Stage, from the equivalent circuit we find the following equations:

$$V_{L_m(t)} = L_m \cdot \frac{di_{L_m(t)}}{dt} \quad (9)$$

$$V_{L_m(t)} = -V_0 \quad (10)$$

From the Eqs. (9) and (10) we obtain the magnetizing inductance current:

$$i_{L_m(t)} = i_{L_m(\max)} - \frac{V_0}{L_m} \cdot t \quad (11)$$

Where, from Eq.3 taking $t = t_p$, we have:

$$i_{L_m(\max)} = \sqrt{3} \cdot \frac{V_p}{L_m} \cdot t_f \cdot \sin(\omega \cdot t) + i_{L_m(\min)} \quad (12)$$

By substituting Eq. (12) into Eq. (11):

$$i_{L_m(t)} = \frac{\sqrt{3} \cdot V_p \cdot t_f}{L_m} \cdot \sin(\omega \cdot t) - \frac{V_0}{L_m} \cdot t + i_{L_m(\min)} \quad (13)$$

The expression of the output inductance L_0 is given by:

$$V_{L_0(t)} = L_0 \cdot \frac{di_{L_0(t)}}{dt} \quad (14)$$

$$V_{L_0(t)} = -V_0 \quad (15)$$

By substituting Eq. (15) into (14) and solving the differential equation, we obtain:

$$i_{L_0(t)} = i_{L_0(\max)} - \frac{V_0}{L_0} \cdot t \quad (16)$$

were, from Eq. 6 taking $t = t_p$, we have:

$$i_{L_0(\max)} = \frac{\sqrt{3} \cdot V_p}{L_0} \cdot t_f \cdot \sin(\omega \cdot t) + i_{L_0(\min)} \quad (17)$$

Thus, $i_{L_0(t)}$ current is given by:

$$i_{L_0(t)} = \frac{\sqrt{3} \cdot V_p}{L_0} \cdot t_f \cdot \sin(\omega \cdot t) - \frac{V_0}{L_0} \cdot t + i_{L_0(\min)} \quad (18)$$

The D_f diode current is given by sum of the currents through the inductances L_m and L_0 . Consequently:

$$i_{D_f(t)} = i_{L_m(t)} + i_{L_0(t)} = \frac{\sqrt{3} \cdot V_p \cdot t_f}{L_{eq}} \cdot \sin(\omega \cdot t) - \frac{V_0}{L_{eq}} \cdot t + i_{L_m(\min)} + i_{L_0(\min)} \quad (19)$$

From Eq. (18), for $t = t_a \Rightarrow i_{L_0} = i_{L_0(\min)}$. Thus, the time t_a can be obtained as follow:

$$t_a = \frac{\sqrt{3} \cdot V_p \cdot t_f}{L_0} \cdot \sin(\omega \cdot t) \quad (20)$$

Where:

$$\sqrt{3} \cdot V_p \cdot \sin(\omega \cdot t) = V_{in} \quad (21)$$

We know that the parameter " t_a " is constant; so, the V_{in} voltage can be replaced by its average value given by:

$$V_{in_{av}} = \frac{3 \cdot \sqrt{3}}{\pi} \cdot V_p \quad (22)$$

Finally, substituting Eq. (22) into Eq. (21), and put the results obtained into Eq. (20), the expression (23) is find:

$$t_a = \frac{3 \cdot \sqrt{3} \cdot V_p}{\pi \cdot V_0} \cdot t_f \quad (23)$$

Defining:

$$a = \frac{\sqrt{3} \cdot V_p}{V_0} \quad (24)$$

We have:

$$t_a = \frac{3 \cdot \alpha}{\pi} \cdot t_f \quad (25)$$

3. Design Procedure and Example [4].

3.1. Input Data

$V_{RMS} = 220V$ (rms line voltage);
 $P_o = 3.0$ kW (output power);
 $V_o = 60V$ (average output voltage);
 $f_s = 20kHz$ (switching frequency);
 $f_L = 60$ Hz (line frequency).

3.2. Transformer Ratio (a)

$$a = N_1/N_2 = 5 \quad ; \quad V_o = 300V$$

3.3. Static Gain (G) and Duty Cycle (D)

$$G = \frac{V_o}{\sqrt{3} \cdot V_p} \Rightarrow G = 0.557 \quad (26)$$

$$\text{Where: } V_p = \sqrt{2} \cdot V_{RMS} = 311V$$

$$V_{in} = \sqrt{3} \cdot V_p \cdot \sin(\omega \cdot t)$$

By definition $\alpha = 1/G$; consequently:

$$\alpha = 1.796$$

In the Eq. (25) replacing t_a by $T_S - t_f$, we obtain the expression of the duty cycle (D):

$$D = \frac{1}{1 + \frac{3}{\pi} \alpha} \quad ; \quad D = 0,368$$

3.4. Output Current (I_o) and Load Resistance (R_o) reflected into the primary side.

$$I_o = \frac{P_o}{V_o} = 10A \quad ; \quad R_o = \frac{V_o}{I_o} = 30\Omega$$

3.5. Equivalent Inductance (L_{eq})

$$\frac{1}{L_{eq}} = \frac{1}{L_m} + \frac{1}{L_o}$$

From the equationing presented in [4], we have:

$$L_{eq} \geq \frac{R_{o(max)} \cdot (1-D)^2}{2 \cdot f_s} \quad (27)$$

We desire continuous conduction operation up to 10% of the load. Therefore:

$$R_{o(max)} = \frac{V_o}{10\% \cdot I_o} \quad \therefore R_{o(max)} = 300\Omega \Rightarrow L_{eq} \geq 3.13 \text{ mH}$$

Thus, the L_{eq} chosen was \Rightarrow $L = 3,20 \text{ mH}$

3.6. Output Inductance (L_o) and Magnetizing Inductance (L_m)

For that purpose, we adopt an output current ripple of 10%. Thus:

$$L_o = \frac{\sqrt{3} V_p D}{f_s \Delta i_{L_o}} \Rightarrow L_o = 9,9 \text{ mH}$$

From the relation:

$$\frac{1}{L_{eq}} = \frac{1}{L_m} + \frac{1}{L_o} \Rightarrow L_m = 4,7 \text{ mH}$$

3.7. Capacitor C_1 and Output Capacitor (C_o)

For both capacitors we adopt a 10% voltage ripple ($\Delta V_{C_1} = \Delta V_{C_o} = 30V$). Therefore, from the equationing developed in [4] we have:

$$C_1 = \frac{\pi \cdot I_o \cdot D}{3 \cdot \Delta V_{C_1} \cdot f_s} = \frac{\pi \cdot 10 \cdot 0.368}{3 \cdot 30 \cdot 20K} \Rightarrow C_1 = 6,42 \mu F$$

$$C_o = \frac{I_o \cdot (2 - \sqrt{3})}{72 \cdot f_L \cdot \Delta V_{C_o}} = \frac{10 \cdot (2 - \sqrt{3})}{72 \cdot 60 \cdot 30} \Rightarrow C_o = 20,7 \mu F$$

Note: For the input filter the conventional design was used. So, the following values were obtained:

$$C_F = 365 \text{ nF (start configuration)}$$

$$L_F = 17,3 \text{ mH}$$

Although the input filter inductances result in a high value, they can be build of silicon iron plate, obtaining cheap and low volume inductors.

4. Simulation Results

Figures below show the main results obtained from numerical simulation, using the calculated data in the previous item.

Fig.5 presents the input voltage and current waveforms for the full load condition. We can observe a little displacement, about $-7,862$, between the input voltage and the fundamental current component. This

displacement is caused by the input filter inductances. From the simulation results, for the full load condition, we have:
 TDH = 7,7 % and PF = 0,984

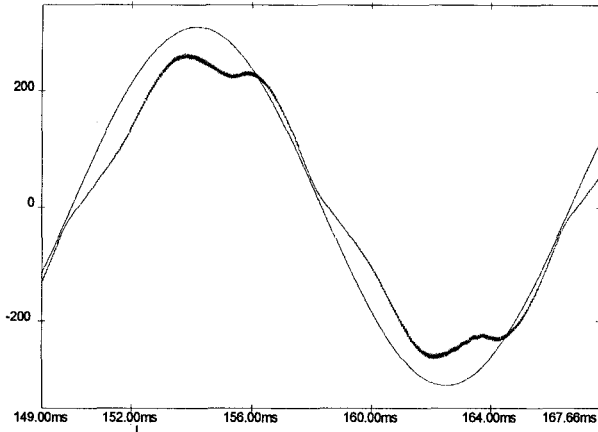


Fig 5. Input voltage and input current (*40) per phase.

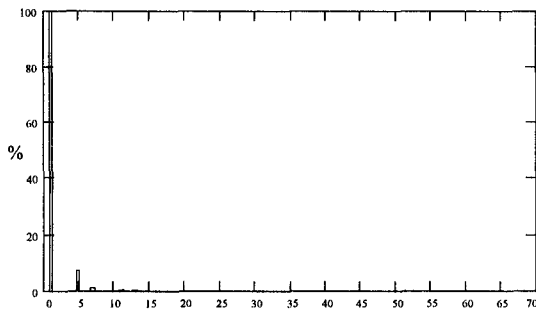


Fig 6. Current harmonic analysis per phase

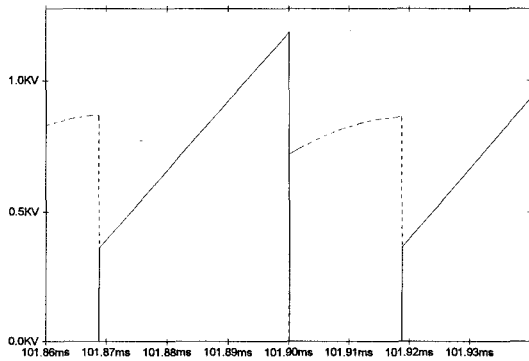


Fig 7. Switch voltage → full line
 Switch current (*30) → dotted line

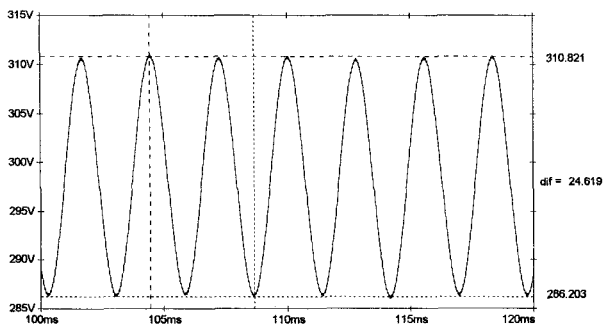


Fig 8. Output voltage

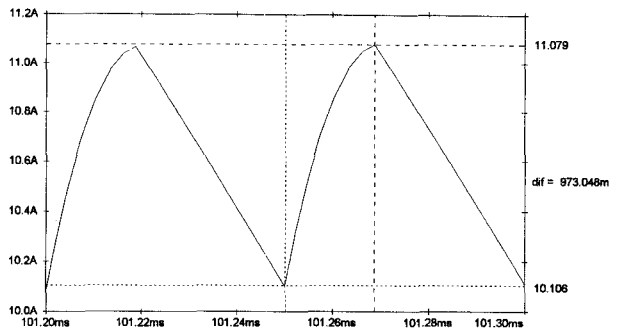


Fig 9. Output current

5. Experimental Results

A prototype rated 600W was built to evaluate the proposed circuit [4]. The prototype was designed to operate in CCM from 20% of the load. IGBT was used for the main switch. Other characteristics of the prototype were:

$$V_{in} = 73,3 \sin \omega t \text{ (V)} ; f_s = 20\text{KHz} ;$$

$$V_o = 60\text{V} ; N_1/N_2 = 2$$

All the results presented in this paper were obtained for full load conditions and the output voltage were kept constant, equal to 60V. Fig.10 presents the voltage and current in phase "A". The harmonic analysis is shown in Fig.11, from which the harmonic distortions of 2,9% and 6% were obtained respectively. In this case the power factor was 0,984.

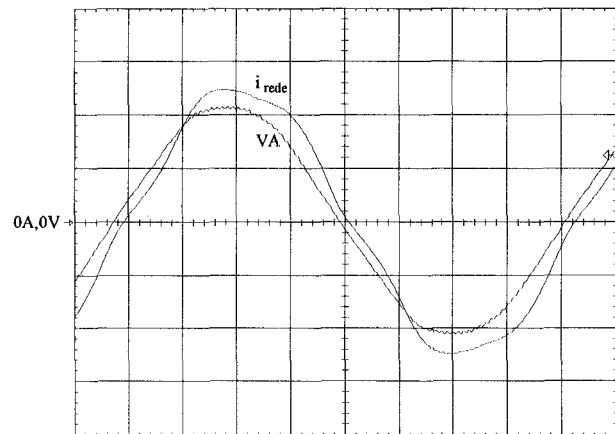


Fig 10. Voltage and current in phase A
 Scale: 50V/div ; 2A/div ; 2ms/div

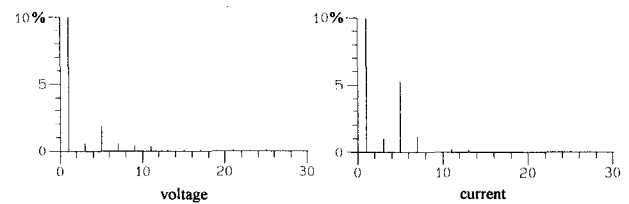


Fig 11. Harmonic analysis.

Table I presents many acquisitions of voltage and current harmonic analysis in phase A, for different load conditions. From these data Figs 12 and 13 were elaborated. In this experiment the output voltage was kept constant equal to 60V, and the converter was operating in a closed-loop control. We can observe that for 20% of the load, we have the worst results relating to Total Harmonic Distortion and Power Factor (THD about 33% and PF about 0,947).

load (%)	voltage THD (%)	current THD (%)	displacement of the current (degree)
5	4	29	-0.98
10	4	30	-3.62
20	4	33	-4.45
30	4	27	-5.44
40	4	22	-6.14
50	4	17	-6.33
60	5	14	-6.76
70	5	10	-8.48
80	5	9	-8.54
90	5	7	-8.61
100	6	6	-9.69

Table 1. THD of the voltage and current of phase A

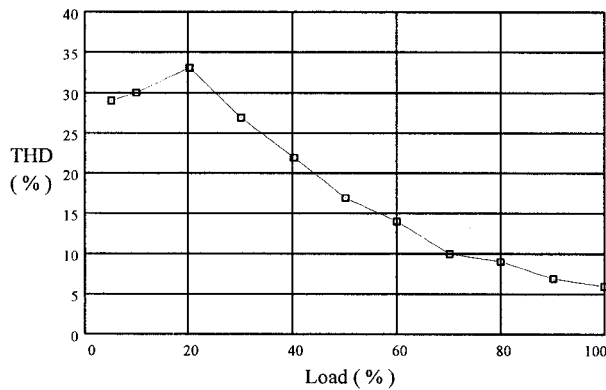


Fig 12. THD vs load

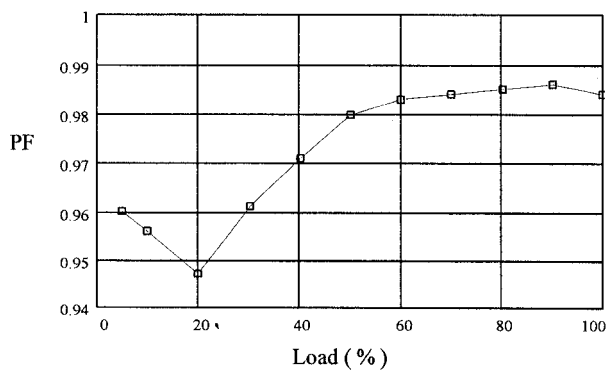


Fig 13. Power Factor vs Load

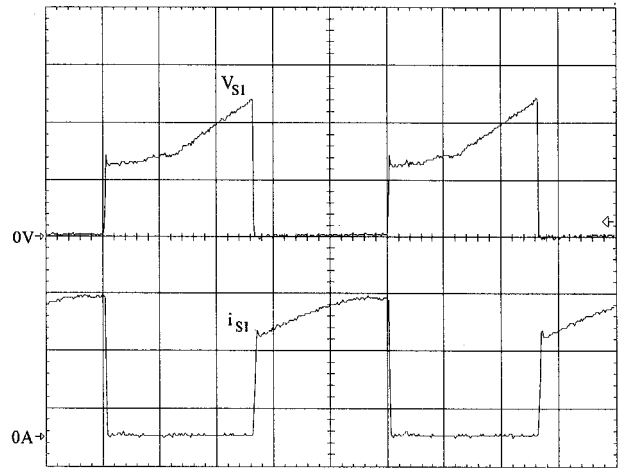


Fig 14. Voltage and current in the IGBT
Scale: 200V/div ; 5A/div ; 10µs/div

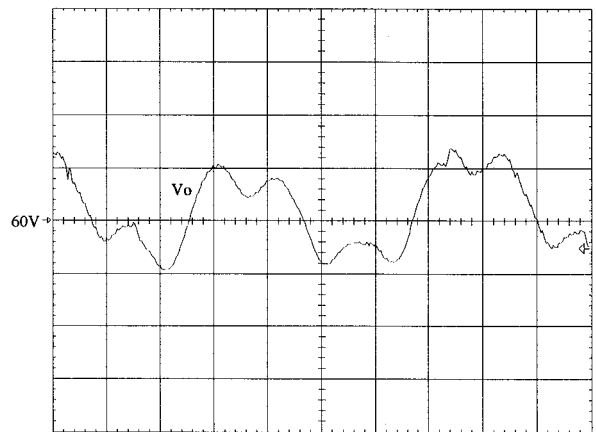


Fig 15. Output voltage
Scale: 500mV/div ; 2ms/div

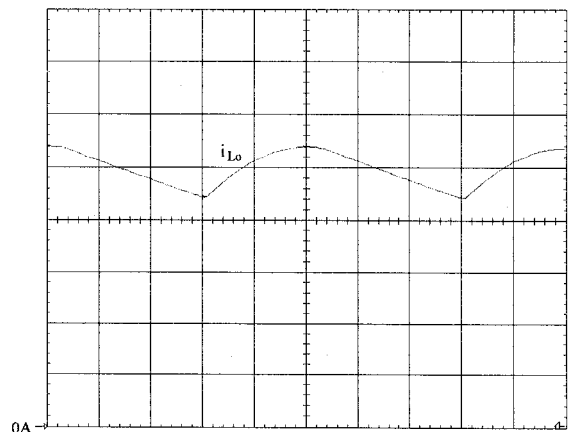


Fig 16. Output current
Scale: 2A/div ; 10µs/div

Fig.14 presents the voltage and current in the IGBT.

The output voltage and current are shown in Figs 15 and 16 respectively, and Fig.17 presents the behavior of the voltage across the capacitor C_1 . Figs 18 and 19 show the dynamic behavior of the converter for a 50% load increase.

In Fig.18 the voltage originating in the controller, implemented from a simple integrator, is presented. This voltage is compared with a sawtooth signal, generating a PWM pulse. Fig.19 shows that, for a 50% variations of the load there is neither oscillations nor overvoltage in the output voltage of the converter.

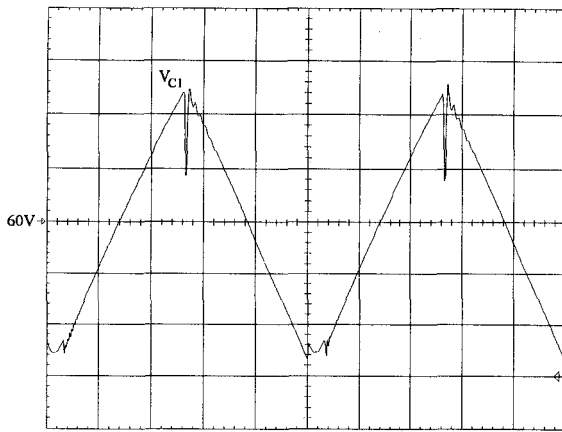


Fig 17. Capacitor C_1 voltage
Scale: 2V/div ; 10 μ s/div

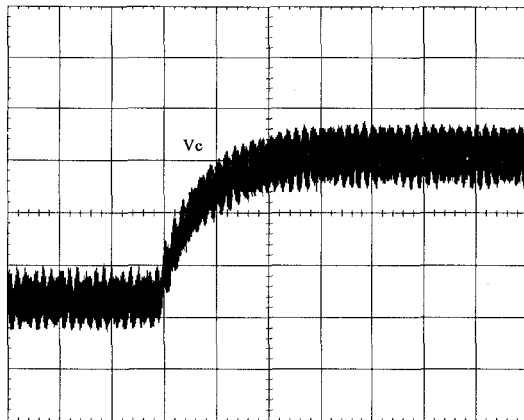


Fig 18. Controller voltage
Scale: 10mV/div ; 100ms/div

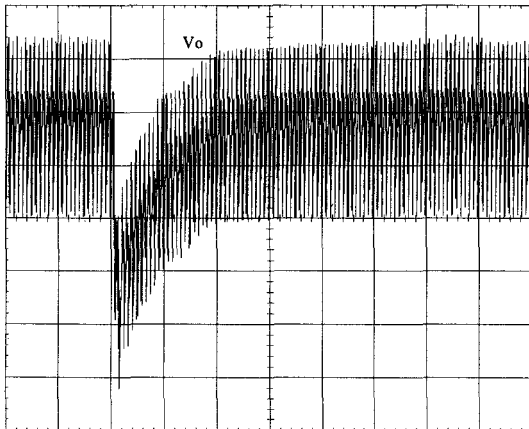


Fig 19. Output voltage for a variation load
Scale: 500mV/div ; 100ms/div

6. Conclusion

This paper has presented the analysis of an isolated three-phase rectifier with high power factor using a DC-DC Zeta converter operations in continuous conduction mode. The use of the Zeta converter in CCM permits to reduce the peak and the rms current value through the switch and the magnetic elements. Consequently, the

conduction losses in the structure are decreased. Furthermore, this converter has an input characteristic such as voltage source. Thus, it does not need an input inductor in each phase, increasing the efficiency and decreasing the volume and the weight of the system.

Therefore, according to the results obtained we have an AC-DC converter with the following features:

- It is particularly simple and robust;
- It provides power factor correction operating in continuous conduction mode and is therefore more adequate for high power application;
- It is naturally isolated;
- It has only one controlled switch;
- It operates either as step-up or step-down voltage;
- It can allow a regulated output voltage with only one power processing stage;
- In case of failure, the switch can open, protecting the structure;
- Its current source characteristic readily offers the possibility of parallel association.

The high peak voltage across the switch represents the great drawback of this structure. However, considering the many features previously presented, the authors believe that this structure can be very useful for some industrial applications.

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