A Line-Frequency Commutated Rectifier Complying with IEC 1000-3-2 Standards

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Abstract. Consumer and household appliances require cheap ac/dc power supplies complying with EMC standards. The commonly employed passive solutions are bulky and do not provide output voltage stabilization. Active solutions, based on PFC's with high-frequency switching, provide compactness and regulation capability, but are generally expensive due to the need for fast-recovery diodes and complex EMI filters. This paper presents a high power factor rectifier, based on a modified conventional rectifier with passive L-C filter, which improves both the harmonic content of the input current and the power factor, by means of a low frequency commutated switch and a small line-frequency transformer, and allows to comply with IEC 1000-3-2 standard with reduced overall inductive components' volume.

I. INTRODUCTION

Line-current harmonic standards, like IEC-1000-3-2 [1], have led to a great effort in developing front-end AC to DC converters absorbing lightly distorted currents. High frequency power factor correctors (PFC's), which draw from the grid a current nearly proportional to the input voltage, have already been extensively analyzed in the literature. Their typical performance is very good but, for some large volume applications, like household appliances and personal computers they imply an unacceptable increase of the cost and complexity of the conversion unit. These applications indeed require very cheap and reliable solutions; therefore, in many cases, passive filters are still used in conjunction with diode rectifiers. A classical diode bridge rectifier and filter capacitor with a series filter inductor (L-C rectifier), can actually achieve compliance with the standards, but bulky and heavy reactive components are needed [2].

Different passive configurations are analyzed in [3], which are derived from the classical L-C filter by adding another capacitor inside the rectifier or even another diode [4]. The result is a substantial improvement of the harmonic content of the absorbed current and power factor. However, such solutions are effective for an input power up to 300 W, even taking into account the Class A limits of IEC 1000-3-2 [1]. Moreover, being completely passive, these solutions do not provide any kind of output voltage stabilization.

This paper discusses a high power factor rectifier, based on a modified conventional rectifier with passive L-C filter, which includes, as the main additional elements, a low frequency commutated switch (twice the line frequency), two diodes and a small line frequency transformer. This approach improves both the harmonic content of the line current and the power factor and therefore allows compliance with the standards with a much smaller inductive components' volume as compared to fully passive rectifiers. Moreover, the boost action achieved by the switch operation allows the proposed rectifier to compensate for the input inductor voltage drop and to regulate the rectified output voltage in a wide load range. Finally, the rectifier exhibits limited di/dt and dv/dt, which imply reduced high frequency EMI generation, and very small switching losses, which allow to get a quite high overall efficiency.

II. LINE-FREQUENCY COMMUTATED RECTIFIER

The scheme of the proposed modified rectifier is shown in Fig. 1. The basic structure is that of the usual rectifier with an L-C filter, where an additional switching unit is inserted. Such unit consists of a low-frequency commutated switch, two diodes and a line-frequency transformer which is reset by the secondary side capacitor C_r. All the elements of the switching unit, with the exception of diode D (which can be a slow-recovery diode), are rated for only a small fraction of the output power. The switch is turned on only twice per line period, thus allowing reduced di/dt, dv/dt and losses.

The operation of the circuit depicted in Fig. 1, momentarily neglecting the transformer magnetizing current, can be explained as follows: the switch is turned on with a constant delay T_d after the zero crossing of the line voltage causing a fraction n_2/n_1 of the output voltage to appear in series to the inductor with the right polarity to cause the premature bridge diode turn-on (diode D is off in this interval). As a consequence, the inductor current starts to

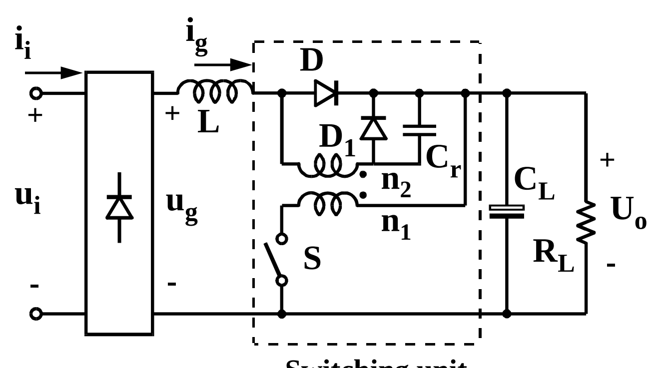

Fig. 1 - Scheme of the low frequency commutated rectifier

increase earlier with respect to the natural diode turn-on instant, as shown by Fig. 2. The duration of the switch on-time T_{ON} is controlled by an output voltage regulator and is limited to a maximum level to avoid the transformer saturation. Thus a simple current limiting protection of the switch is also inherently implemented. As the switch turns off, diode D starts to conduct and the filter inductor resonates with the output capacitor.

The input current waveform equations, neglecting the output voltage ripple, are given by the following expressions:

1. interval T_{ON}: $T_d \le t \le T_d + T_{ON}$

$$i_g(t) = \frac{\hat{U}_g}{\omega_g L}\left[\cos(\omega_g T_d) - \cos(\omega_g t) - \frac{U_o}{\hat{U}_g}\left(1 - \frac{1}{n}\right)\omega_g (t - T_d)\right] \tag{1}$$

where $n = n_1/n_2$ is the transformer turns ratio,

2. interval T_{OFF}: $T_d+T_{ON} \le t \le T_d+T_{ON}+T_{OFF}$

$$i_g(t) = I_{g0} + \frac{\hat{U}_g}{\omega_g L}\left[\cos(\omega_g (T_d + T_{ON})) - \cos(\omega_g t) - \frac{U_o}{\hat{U}_g}\omega_g (t - T_d - T_{ON})\right] \tag{2}$$

where I_{g0} is the input current value at the end of the T_{ON} interval. Then the inductor current goes to zero and diode D turns off together with the bridge diodes.

In the practical implementation a reset circuitry must be provided to make sure the magnetizing current is forced to zero at the end of each period. This is achieved by capacitor C_r. To explain the reset process let us refer to Fig. 3a which reports the transformer winding currents i_{pri} and i_{sec} and the voltage across the reset capacitor C_r. During the switch on-time, the output voltage is applied to the transformer primary winding and its magnetizing current i_μ increases linearly according to the relation:

$$i_\mu(t) = \frac{U_o}{L_\mu} t \tag{3}$$

where L_μ is the primary magnetizing inductance. During the same interval the secondary winding current i_{sec} coincides with the input current. Note that due to a residual voltage across capacitor C_r at the beginning of the T_{ON} interval, the input current does not follow the behavior predicted by (1); instead, the input inductor initially resonates with C_r until it is completely discharged, causing the input current waveshape shown in Fig. 3a to differ slightly from the ideal case of Fig. 2. When the switch turns off, i_μ transfers to the secondary winding and charges capacitor C_r flowing through diode D, which now conducts the sum of the input current and the magnetizing current reflected to the secondary side. Thus, a resonant oscillation between capacitor C_r and

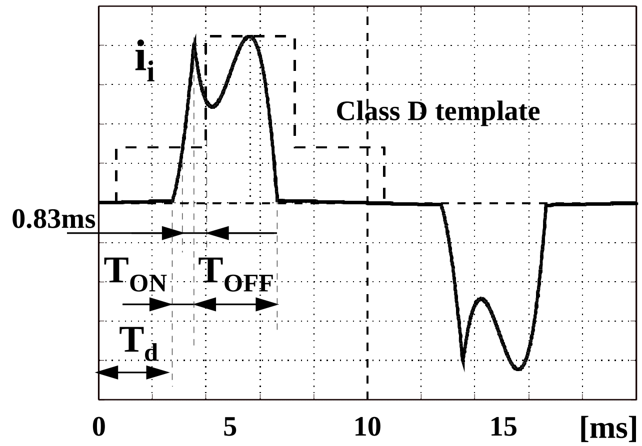

Fig. 2 - Input current waveform (1 A/div) of low-frequency PFC and Class D template (U_i = 220 V_{rms}, P_o=300 W, T_d = 3 ms, T_{ON} = 0.7 ms).

magnetizing inductance L_μ takes place. Note that the resonant frequency can be selected low enough to cause the input current to go to zero when i_μ is still greater than zero, thus minimizing the switch voltage stress. In fact, the magnetizing current and capacitor C_r voltage during this interval are given by:

$$\begin{aligned} i_\mu(t) &= \hat{I}_\mu \cos(\omega_r t) \\ u_{Cr}(t) &= Z_r n \hat{I}_\mu \sin(\omega_r t) \end{aligned} \tag{4}$$

where $\hat{I}_\mu = \frac{U_o}{L_\mu} T_{ON}$, $\omega_r = \frac{n}{\sqrt{L_\mu C_r}}$, and $Z_r = \frac{1}{n}\sqrt{\frac{L_\mu}{C_r}}$.

Thus the highest C_r value should be chosen which ensures the transformer reset, and this goal is achieved when the resonance period is chosen to be four times the minimum available reset time, i.e.:

$$T_r = 4\left(\frac{T_g}{2} - T_{ON\,max}\right), \qquad T_r = \frac{2\pi}{\omega_r} \tag{5}$$

where T_g is line period. In this way, assuming the reset capacitor C_r is completely discharged during the switch on-time, the switch voltage stress is minimized and is given by:

$$\hat{U}_S = U_o\left(1 + \pi \frac{T_{ON\,max}}{T_g - 2\,T_{ON\,max}}\right) \tag{6}$$

This result holds on the hypothesis that the switch on-time is long enough to completely discharge C_r during T_{ON}, as shown in Fig. 3a. If this is not the case, the voltage across C_r stabilizes around an average value which guarantees the transformer reset. At the limit of a constant voltage across it, the switch voltage stress becomes:

$$\hat{U}_S = U_o\left(1 + 2 \frac{T_{ON\,max}}{T_g - 2\,T_{ON\,max}}\right) \tag{7}$$

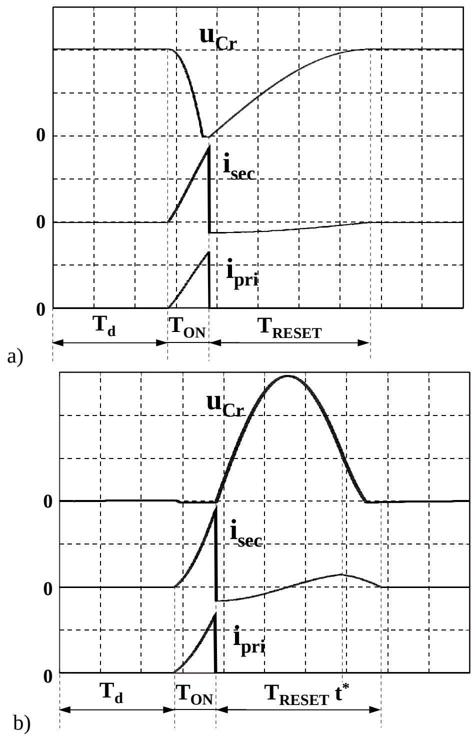

Fig. 3 - Key waveforms during the transformer reset. a) $T_r > 4T_{OFF}$; b) $T_r < 4T_{OFF}$

If the latter condition holds at nominal power, then, in theory, there is no need for diode D_1. However, a small diode should be used in order to prevent the reversal of the voltage across the electrolytic capacitor C_r during transient conditions.

In order to complete the analysis, Fig. 3b shows the circuit behavior with a reduced C_r value which causes interval $T_r / 4$ to be lower than interval T_{OFF}: in addition to the higher peak voltage across it as compared to the previous case, which reflects to the transformer primary winding, we can observe that i_μ can now reverse and at instant t^* it becomes equal in magnitude to the input current, but of opposite polarity. At that point diode D stops conducting, the input current and the secondary magnetizing current remain equal, and go to zero, thus completely the reset interval.

III. Modified Rectifier Applicability

As clearly demonstrated in [3] and [4], there is a wide variety of simple modifications of the conventional L-C diode rectifier which allow to achieve the compliance with the IEC 1000-3-2 standard for loads having a rated power lower than 300 W. The basic idea is to exploit the difference between the absolute harmonic limitations applied to class A loads and the relative limitations applied to class D loads [1]. As known, the difference can be remarkable especially for low power applications. Thus, the goal of these modified rectifiers is to change the shape of the input current so as to stay outside the Class D template, also shown in Fig. 2, for at least 5% of the line half period, i.e. 0.5 ms if the line frequency is 50 Hz. For the low-power range of applications these solutions are surely cost-effective.

If the required output power is in the range between 300 W and 600 W, the same basic idea can be applied, but the simple solutions proposed in [3] may be effectively replaced by a converter, such as the low frequency commutated boost presented in [5] and [6]. The same effect can be achieved also by the modified rectifier proposed in this paper. As a comparison, let us consider the case of a standard diode-capacitor rectifier with inductive filter. The scheme is the same of Fig. 1 without the switching unit. For an input voltage U_i of 220 V_{rms} (which is the minimum voltage considered by IEC 1000-3-2) and a rated power of 300 W, the minimum value of inductor L, which allows compliance with the standard, is 19 mH (C_L =2x470 μF). In this case, the output voltage at the rated current is 276 V, due to the inductor voltage drop. As well known, the resulting line current waveform classifies the rectifier as a Class D piece of equipment. The maximum power deliverable by the equipment is limited by the third harmonic as stated also in [2]. The switching unit added to the standard passive L-C filter shown in Fig. 1 can achieve class A current absorption. The corresponding current drawn by the line for the same operating conditions, i.e. U_i = 220 V_{rms} and P_o = 300 W, is shown in Fig. 2. The figure shows that the input current waveform stays outside the Class D template for more than 5% of the line half-period, thus the rectifier is now in Class A (it is important to remember that the Class D template must be centered to the highest current peak and scaled accordingly). As a consequence, the filter inductor needed to comply with the standard, at this power level, reduces to 4 mH. As it will be explained in the following, the transformer has both a stored energy and a global size which is considerably smaller than the inductor's. Therefore, the converter actually reduces the total magnetic material required to comply with the standard, with respect to the passive solution.

A consequence of the switching unit operation is that the maximum load power is limited by the high-order harmonics (in this case 15^{th} harmonic). The output voltage is stabilized at about 300 V, thanks to the lower inductor voltage drop and to the boost effect of the switching unit. Nevertheless, the solution proposed in [6], achieves the compliance almost with the same inductance value and being a little bit simpler, it is probably the preferred choice for this power range.

If the required output power is higher than 600 W, the load is considered in class A, no matter the current waveform. The modified rectifier has no longer the aim of modifying the input current to stay out of the class D template, but simply to improve the current harmonic content. The boost converter proposed in [5] and [6] requires inductor values in the range around 5 mH to achieve this goal. The solution we discuss here requires almost the same inductor. For instance, at P_o = 600W, 6 mH are enough to comply with the standard. The presence of the transformer makes the boost solution still preferable. Table I sums up all of these comparative considerations including also other relevant data. To derive the Table, for different power levels ranging from 300 up to 900 W, a passive L-C rectifier (P) is simulated together with the proposed active rectifier (A_1) and the boost rectifier (A_2). For each power level listed in the Table the following data were collected: average output voltage U_o, inductor current value ensuring compliance with the standard (Class D for the passive solution up to 600 W and Class A for the active ones and for higher output power), peak inductor current, peak energy E_L in the inductor ($E_L = 0.5\ L\ I^2_{gpeak}$), input current RMS value I_{grms}, distortion factor $DF = I_{g1rms}/I_{grms}$, displacement factor $\cos(\phi_1)$, power factor $PF = DF \cdot \cos(\phi_1)$, peak-to-peak output voltage ripple Δu_o. By comparing the results, and taking into account the previous remark on the transformer size, it is possible to conclude that the solution we discuss here can be effectively applied to reduce the size of the magnetic components necessary for the compliance with IEC standard 1000-3-2, especially in the power range from 600 W to at least 900 W.

IV. Design Considerations

A. Selection of reactive element values

To develop a fully-compliant rectifier, the first step is the selection of the L and C reactive element values. As far as the output capacitor value is concerned, a good guess is the value obtained by the approximate analysis of the classical diode-bridge+capacitive filter rectifier, i.e.:

$$C_L = \frac{P_o}{2\, f_{LINE}\, U_o\, \Delta U_{opp}} \tag{8}$$

where ΔU_{opp} is the maximum allowed output voltage ripple (peak-to-peak). Note that, due to the extended diode conduction angle, caused by the filter inductor, and the switching unit operation, the effective output voltage ripple will be lower than the theoretical one.

The choice of the filter inductor is more difficult and the design guidelines given here have to be verified by simulation. In case the desired output power is lower than 600 W, the goal is to modify the waveshape of the input current so as to take advantage of the less restrictive Class A limits. This single condition, normally allows compliance with the standard. Thus, a good starting point should be an inductor value which, without the help of the switching unit, achieves at least 60° of conduction angle, which is the width of the Class D template. Only in this case, in fact, the switching unit can increase the conduction angle so as the current waveform stays outside the Class D template for at least 5% of the line half-period without using high T_{ON} values which would cause an increase in the transformer size and of the high-frequency current harmonics. For power levels above 600 W no difference exists between Class D and Class A limits, thus the inductor value should be progressively increased as the power increases. In fact, the extension of the conduction angle and the reduction of the current rate of change during the switch on-time are mandatory in order to keep the current harmonics below the limits.

B. Transformer design

The objective of this work is to provide compliance with the standards with a reduced overall magnetic components' volume as compared to the passive solution. To this purpose, the transformer size should be minimized by choosing the minimum switch on-time which provides the desired current harmonic reduction. This, together with the desired turns ratio, determines the winding number of turns. Then, for a complete transformer size estimation, the winding RMS currents are calculated approximating with a linear rise the shape of the input current during T_{ON} (see Fig. 3). From this figure and from (3) and (4) we obtain:

$$I_{sec_{RMS}} = \sqrt{\left(I_{g0}\sqrt{\frac{2}{3}\frac{T_{ON}}{T_g}}\right)^2 + \left(\frac{n\hat{I}_\mu}{2}\sqrt{\frac{\omega_g}{\omega_r}}\right)^2} \tag{9}$$

$$I_{pri_{RMS}} = \left(\frac{I_{g0}}{n} + \hat{I}_\mu\right)\sqrt{\frac{2}{3}\frac{T_{ON}}{T_g}} \tag{10}$$

where I_{g0} is the input current value at the end of the T_{ON} interval, calculated from (1). The transformer volume is related to the product of iron and window areas, i.e.:

$$A_e A_w = \frac{U_o T_{ON}}{B_{max}}\left(I_{pri_{RMS}} + \frac{I_{sec_{RMS}}}{n}\right)\frac{1}{J\, k_R} \tag{11}$$

where B_{max} is the maximum flux density, J is the desired current density and k_R the window filling coefficient.

In order to give an idea of the transformer dimensions, let us consider a practical example:

Converter specifications:

Table I. Comparison between passive and active rectifiers at different power levels

P_o [W]	U_o [V]	L [mH]	I_{gpeak} [A]	E_L [mJ]	I_{grms} [A]	DF	$\cos(\phi_1)$	PF	Δu_o [V]
300 - P	*276*	*19*	*4.11*	*160*	*1.85*	*0.792*	*0.926*	*0.733*	*7.02*
300 - A_1	301	4	4.23	36	1.89	0.792	0.916	0.725	6.63
300 - A_2	*301*	*6*	*4.08*	*50*	*1.78*	*0.766*	*0.998*	*0.764*	*7.13*
600 - P	*284*	*7*	*8.9*	*277*	*3.81*	*0.757*	*0.946*	*0.716*	*14.46*
600 - A_1	300	6	7.06	150	3.32	0.810	0.999	0.810	13.00
600 - A_2	*293*	*5*	*8.8*	*194*	*3.69*	*0.755*	*0.975*	*0.736*	*16.40*
900 - P	*247*	*20*	*10*	*1000*	*5.30*	*0.893*	*0.862*	*0.770*	*18.76*
900 - A_1	300	8	8.34	278	4.52	0.872	0.999	0.871	17.17
900 - A_2	264	15	9.93	739	5.09	0.883	0.921	0.813	24.85

P = passive; A_1 = active proposed solution; A_2 = active boost rectifier; DF = Distortion Factor; $\cos(\phi_1)$ = displacement factor; PF = Power Factor

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$U_i = 220V_{rms}\pm20\%$, $P_o = 800W$, $L = 6$ mH, $T_d = 2.8$ms, $T_{ON} = 0.5$ms, $n = 4$.

The material used for both the inductor and the transformer has the following parameter values:

relative permeability:............................$\mu_r = 11674$
flux density:..$B = 1.35$ T

The utilized window filling coefficient k_R is 0.4, and the current density J is 3 A/mm^2. The transformer parameters are:

iron cross section:................................$A_e = 2.56\cdot10^{-4}$ m^2
window area: ....$A_w = 1.92\cdot10^{-4}$ m^2
mean magnetic path:............................$\ell_{avg} = 0.104$ m
primary number of turns: ....$N_1 = 435$
secondary number of turns:..................$N_1 = 109$
magnetizing inductance: ....$L_\mu = 0.56$ H
primary wire diameter: ....$\Phi_1 = 0.25$ mm
primary wire diameter: ....$\Phi_2 = 0.6$ mm
total winding area:...............................$A_{cu} = 1.3\cdot10^{-4}$ m^2.
external core volume:...........................$Vol = 3.07\cdot10^{-5}$ m^3

Note that the total winding area A_{cu} is well below the available window area A_w, meaning that the transformer size could be further reduced.

The inductor parameter, calculated for the maximum input current (i.e. minimum input voltage), are:

iron cross section:................................$A_e = 7.7\cdot10^{-4}$ m^2
window area: ....$A_w = 3.63\cdot10^{-4}$ m^2
mean magnetic path:............................$\ell_{avg} = 0.143$ m
number of turns:...................................$N = 67$
air gap: ....$t_{gap} = 0.36$ mm
wire diameter:......................................$\Phi = 1.6$ mm
total winding area:...............................$A_{cu} = 3.37\cdot10^{-4}$ m^2.
external core volume:...........................$Vol = 1.27\cdot10^{-4}$ m^3.

The rectifier output voltage at the minimum input voltage and nominal power is 222 V.

For the sake of comparison a similar design was carried out for the passive solution. The inductor value needed to comply with the standard for the same converter specification is 15 mH. The resulting inductor parameters are:

iron cross section:................................$A_e = 1.12\cdot10^{-3}$ m^2
window area: ....$A_w = 5.88\cdot10^{-4}$ m^2
mean magnetic path:............................$\ell_{avg} = 0.182$ m
number of turns:...................................$N = 111$
air gap: ....$t_{gap} = 0.58$ mm
wire diameter:......................................$\Phi = 1.6$ mm
total winding area:...............................$A_{cu} = 5.58\cdot10^{-4}$ m^2.
external core volume:...........................$Vol = 2.35\cdot10^{-4}$ m^3.

The rectifier output voltage at the minimum input voltage and nominal power is 198V. Comparing the resulting volumes, the reduction implied by the proposed solution is about 33%.

C. Selection of switching unit parameters

The design of the proposed converter and switching unit is characterized by several degrees of freedom. All the design parameters are somehow related to one another; therefore different design strategies can be identified. A possible procedure is to select the duration of the switch on-time, which directly determines the size of the transformer, to be as small as possible.

After this choice, which must be guided by simulations, the transformer turns ratio $n = n_1/n_2$ has to be selected. The effect of the variation of this parameter is illustrated by Fig. 4. As can be seen, by increasing the turns ratio it is possible to improve the high frequency harmonic content of the line current. This helps to limit the inductor value and/or the duration of the switch on-time needed to achieve compliance. The inevitable drawback is that, increasing the turns ratio, the converter boost action reduces and so the quality of the output voltage regulation worsens.

The effect of the turn-on delay T_d is described by Fig. 5. As can be seen, the increase of the delay initially reduces the harmonic content, but further increasing it implies an increase in the current peak value (Fig. 5b) and also in the harmonics.

A further effect of the variation of the described control parameters is the variation of the output voltage achieved by the converter in open loop conditions, which accounts for the boost capability of the rectifier. This is described by Fig. 4 and Fig. 5 too. As can be seen, both an increase of the delay and a reduction of the transformer turns ratio imply an increase in the boost action of the converter. This effect must be traded-off against the previously discussed drawbacks.

D. Output voltage regulation

As far as the output voltage regulation is concerned, we must consider separately the effects of load and input voltage variations, having in mind the constraint imposed by the maximum switch on-time, which strongly affects the transformer size. Thus, once we have selected the maximum T_{ON} in order to achieve compliance with the standard at nominal load and prescribed input voltage, the control can only reduce the switch on-time at load current decreasing (delay time T_d is simply kept constant). A standard PI regulator having a bandwidth well below the line frequency, like any other PFC regulator, is sufficient to do this. Clearly, a minimum power level exists, below which the output voltage regulation cannot be maintained. It corresponds to the value for which the passive L-C rectifier (without the switching unit) achieves the same output voltage. At lower power levels, the output voltage increases toward the input voltage peak, like in any standard rectifier. For this reason, a high output voltage reference is preferable, since it can be maintained for a broader load variation. To give an idea, the converter described in the experimental results' section can maintain the output voltage regulation approximately down to 30% of the nominal power.

Differently from the low-frequency boost converter presented in [5]-[6], the proposed topology does not achieve a high boost action unless a low transformer turns ratio is used (at the limit of a unity turns ratio the behavior of this structure becomes the same as [5]-[6]). As a consequence, regulation of the output voltage can be maintained only for a small input voltage increase (which requires reduction of the switch on-time), while, at low input voltage, T_{ON} is kept constant and equal to the maximum value allowed by the transformer design, causing the decrease of the output voltage too.

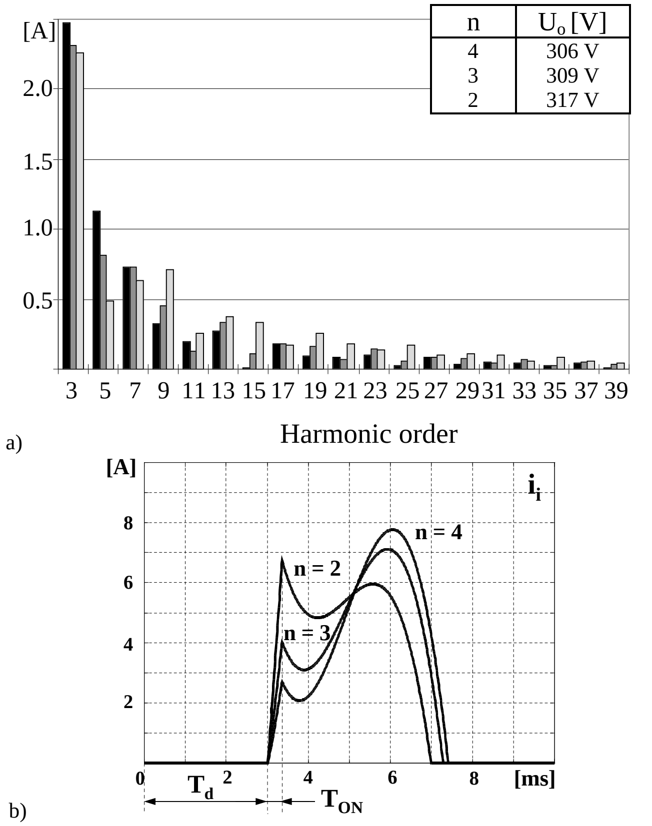

Fig. 4 - Line current as a function of the transformer turns ratio. a) Frequency domain; b) time domain (U_i = 230 V_{rms}, P_o = 600W, T_d = 3 ms, T_{ON} = 360 µs)

V. Experimental Measurements

In order to verify the results obtained by simulation a prototype was built having the following specifications:

U_i = 230 V_{rms}, U_o = 300V, P_o = 900W, L = 5.3 mH, C = 2x470µF

Initially the rectifier was tested without activating the switching unit. The line voltage and current measured in these conditions are shown in Fig. 6. It is important to notice that in all the performed measurements a controlled, low

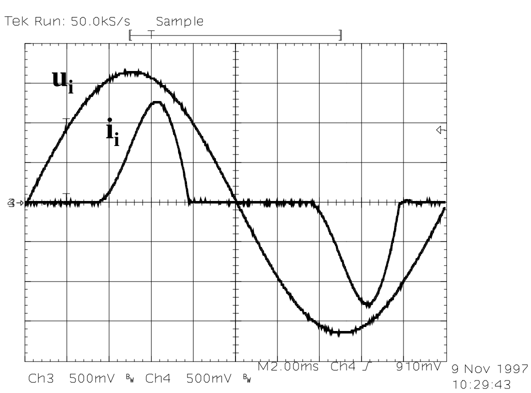

Fig. 6 - Input voltage U_i (100V/div) and current i_i (5A/div) (U_i = 230V_{rms}, P_o = 900W)

impedance, voltage source is used as the test power supply.

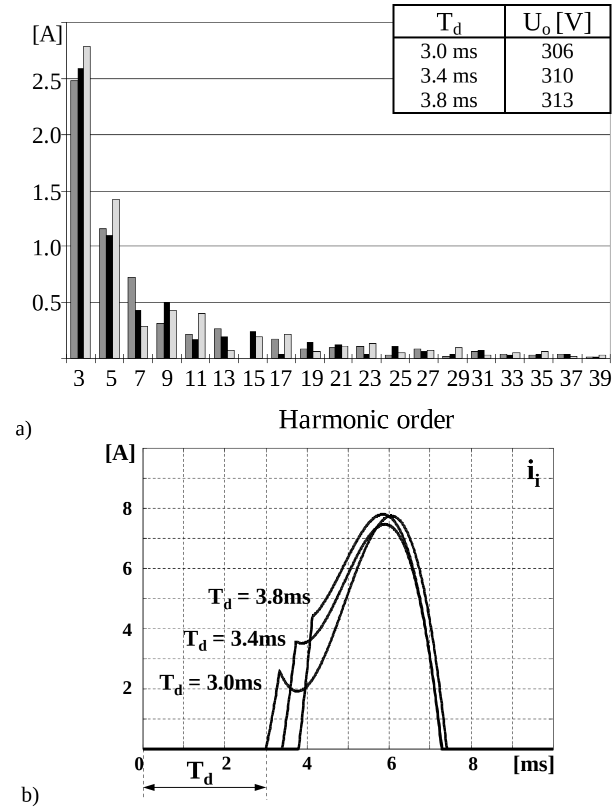

Fig. 5 - Line current as a function of turn-on delay T_d. a) Frequency domain; b) time domain (U_i = 230V_{rms}, P_o = 600W, T_{ON} = 360 µs, n = 4)

This allows to have an almost harmonic-free input voltage, as required by the IEC standards. For the passive L-C rectifier, the harmonic content of the current drawn from the utility grid is above the standard limits, especially in the third and fifth harmonic components, as predicted by the simulations.

When the switching unit is activated the current waveform modifies as shown in Fig. 7, where the main converter waveforms at nominal conditions are depicted. The turn-on delay T_d of the gate signal was set to 2.8 ms. As can be seen, the input current waveform well agrees with the simulation results. As a consequence, the compliance with the standard is achieved and only the high order components of the current spectrum get near to the allowed limit values. The harmonic components of the current spectrum can be seen in Table II again for different power levels. Harmonics from 19^{th} up to 25^{th} are normally the closest to the

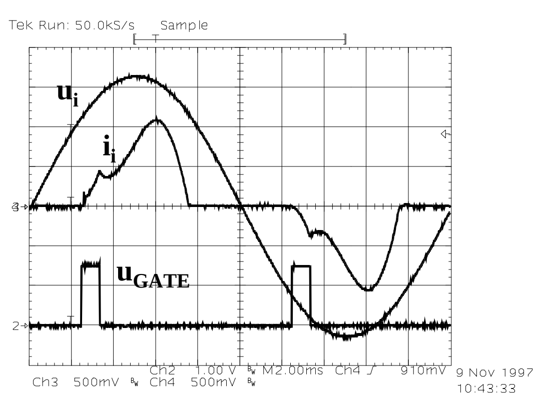

Fig. 7 - Input voltage U_i (100V/div), input current i_i (5A/div) and gate signal u_{gate} (10V/div) (U_i = 230V_{rms}, P_o = 900W)

Table II - Measured input current harmonics at different output power levels for the proposed rectifier

P_o [W]	600	700	800	900	Class A limits
I_n	Har. $[A_{rms}]$	Har. $[A_{rms}]$	Har. $[A_{rms}]$	Har. $[A_{rms}]$	$[A_{rms}]$
I_1	2.736	3.209	3.595	4.100	
I_3	1.739	2.038	**2.223**	**2.293**	**2.30**
I_5	0.825	0.932	**1.123**	1.016	**1.14**
I_7	0.448	0.487	0.618	0.708	0.77
I_9	0.318	0.326	0.173	0.137	0.40
I_{11}	0.041	0.058	0.235	0.273	0.33
I_{13}	0.182	0.200	0.137	0.128	0.21
I_{15}	0.103	0.081	0.103	0.114	0.15
I_{17}	0.067	0.095	0.126	**0.113**	**0.132**
I_{19}	**0.111**	**0.107**	0.035	0.030	**0.118**
I_{21}	0.027	0.017	**0.103**	0.085	**0.107**
I_{23}	**0.075**	**0.089**	0.013	0.031	**0.098**
I_{25}	0.059	0.041	**0.072**	0.052	0.09

corresponding limits, thus confirming the simulation results. However, for the higher power levels, the adopted inductor is probably undersized, since the margin on the low order harmonics tends to reduce. Anyway, as can be seen, the proposed solution allows to comply with the IEC-1000-3-2 standard with a pretty low inductor compared to fully passive solutions. The 5.3 mH inductor adopted in the laboratory prototype allows to increase the power level to about 900 W, without exceeding the standard harmonic limits. It is worth noting that, as explained in the previous section the size of the necessary transformer is a fraction of the inductor's size. Therefore, the overall size of the magnetic components of the power rectifier is greatly reduced as compared to the passive solutions. The measured efficiency of the modified rectifier is always above 96%, as shown by Table III. The Table also reports other measured data which allow to evaluate the different performance of the active and passive rectifier; the boost capability of the active solution, for instance, is indicated by the open loop output voltage achieved by the rectifier. As can be seen, the difference between the output voltage in the active and passive rectifier for a given output power is in the range of 7-8 V, thanks to the switching unit operation. It is also worth noting the negligible difference between active and passive solution efficiency. Finally, the behavior of the reset circuitry and of the adopted power switch (IGBT) snubber are shown by Fig. 8. The drain source

Table III - Experimental comparison of active and passive rectifiers at different power levels (P: passive; A_1: proposed solution)

P_{in} [W]	U_o [V]	I_o [A]	T_d [ms]	T_{ON} [μs]	η [%]
621 (P)	299	2.02			97.2
618 (A_1)	307	1.96	3.00	360	97.4
725 (P)	297	2.37			97.1
724 (A_1)	304	2.31	3.00	360	97.0
827 (P)	295	2.72			97.0
826 (A_1)	303	2.64	2.80	380	96.8
928 (P)	294	3.07			97.3
931 (A_1)	301	3.00	2.45	890	97.0

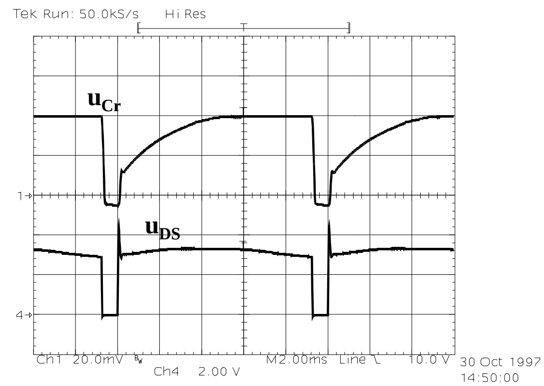

Fig. 8 - Reset capacitor C_r voltage u_{Cr} (4V/div) and drain-source voltage on the IGBT (200V/div)

voltage exhibits a controlled overshoot which is kept by the snubber (100nF, 390Ω (1W)) to an acceptable level for a 600 V switch. The voltage across the reset capacitor is quite low, being the reset time long as compared to the switch on-time.

VI. Conclusions

The proposed low-frequency switched PFC is a simple and cheap solution to achieve compliance with EMC standards together with output voltage stabilization in ac/dc power supplies for household and general-purpose applications. As compared to a passive rectifier, it allows substantial reduction of the inductive components' volume at the expense of a limited increase of circuit complexity.

The added switch allows regulation of the output voltage against load variations, without affecting the converter efficiency. The solution seems to be more effective in the power range above 600W.

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