Comparison Among Line-Frequency Commutated Rectifiers Complying with IEC 1000-3-2 Standards

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***Abstract* - Line-frequency commutated rectifiers represent an interesting solution for large volume applications which do not need a precise output voltage regulation. They provide compliance with IEC 1000-3-2 standards with a smaller overall reactive component volume as compared to conventional rectifiers with passive L-C filter. Moreover, being the switch turned on and off only twice per line period, the associated losses are very small and the di/dt and dv/dt can be lowered compared to high-frequency commutated rectifiers, thus reducing the high-frequency noise emission and EMI filter requirements. This paper reviews the operating principles of different line-frequency commutated rectifier topologies highlighting their merits and limitations. A comparison, in terms of circuit complexity, overall reactive component size and performance, is made for different power levels, thus allowing selection of the most convenient topology for a given application.**

I. Introduction

High quality rectifiers (also called power factor correctors - PFCs) are rapidly substituting conventional front-end rectifiers due to the low-frequency harmonic limits imposed by international standards like IEC-1000-3-2 [1]. Such high-frequency commutated rectifiers provide very high power factors (much more than required by the standards) and a good output voltage regulation, at the expense of an increase of the overall ac-to-dc converter size and cost. For these reasons, some large volume applications still use standard low cost - high reliable rectifiers with passive filters in order to improve the quality of the current drawn from the line, even if the volume of the reactive components needed becomes rapidly prohibitive as the power increases [2].

Recently, different line-frequency commutated rectifier topologies were proposed as a trade-off between high-frequency PFCs and completely passive solutions [3-7]. Compliance with IEC-1000-3-2 can be achieved with a reduced overall reactive component size compared to passive solutions. All of them utilize a switch which is turned on and off only twice per line period, thus drastically reducing both conduction and switching losses. Moreover, the limited di/dt and dv/dt during the commutations allow reduction of the high-frequency noise emission and of the EMI filter requirements.

These different line-frequency commutated rectifier topologies are reviewed with the aim of providing the tools for the best choice for a given application. To this purpose, in addition to the performance in terms of input current distortion and output voltage regulation, also circuit complexity, switch current and voltage stresses and reactive component size are considered for a comprehensive comparison. Such evaluation is carried out for power level ranging from 300W up to 1.2kW.

A. Line-Frequency Commutated Rectifier Topologies

The schemes of the line-frequency commutated rectifier topologies we will consider are reported in Fig.1. Except for the last topology (T4), which is simply derived from the precedent one (T3), each of them contains a line frequency smoothing inductance L and an output filter capacitor C plus an auxiliary switching unit. Such unit consists of a line-frequency commutated switch plus some additional elements including another line-frequency magnetic component, all rated at a small fraction of the output power (except for diode D). All these rectifiers allow compliance with low-frequency harmonic standards, with a reduced smoothing inductor value compared to passive L-C filters, by increasing the input current conduction angle. The principles of operation are now briefly reviewed.

1) Topologies T1 and T2: The first one is identical to a conventional boost converter [3,4] while T2 behaves almost in the same way [5]: when the switch is turned on the input current starts to increase earlier respect to the natural diode bridge turn on instant (during this interval, the input inductor voltage is u_g for T_1 and u_g-$U_o(1-n_2/n_1)$ for T2) and when the switch is turned off, the conventional resonance between L and C occurs until the input current zeroes, remaining zero until the subsequent line half period. Actually, the second topology reduces to the first one if a unity transformer turns ratio is selected. However, a higher n_1/n_2 ratio allows a lower voltage to be applied to the smoothing inductor at the switch turn on which causes a lower initial input current rate of change with a consequent reduction of the high-frequency harmonic amplitude compared to solution T1. Moreover, a high n_1/n_2 ratio reduces the switch current stress at the expense of a higher voltage stress due to the transformer reset mechanism which involves the resonant capacitor C_r. An important topic is the size of the auxiliary transformer which must be kept as small as possible in order to make the overall magnetic component volume lower than that of T1.

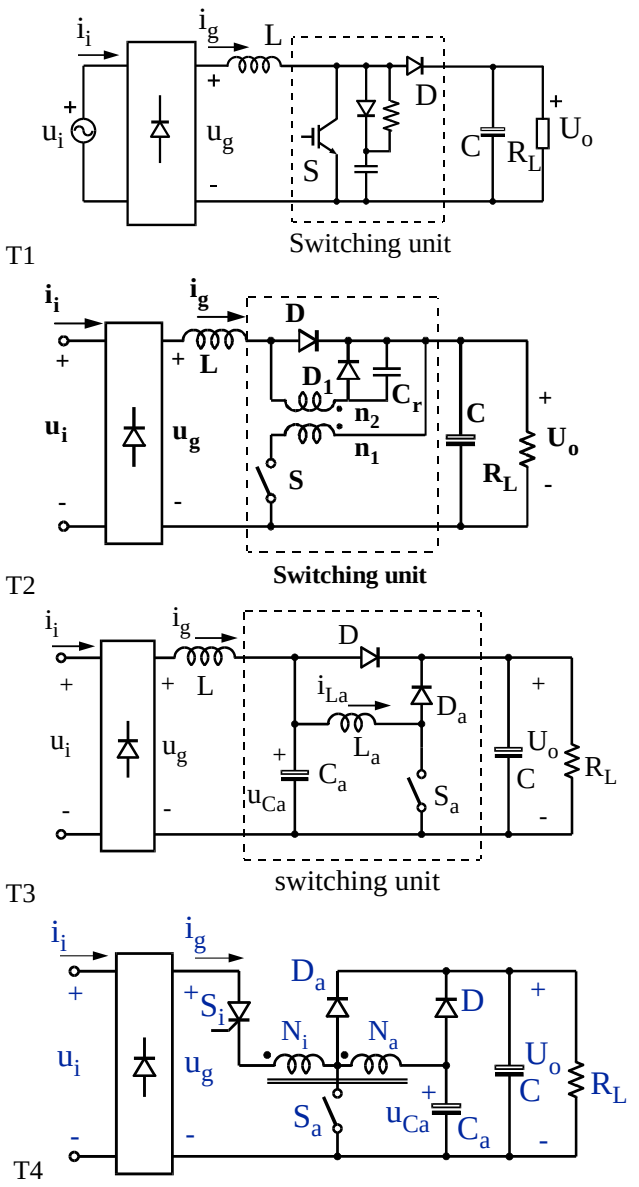


Fig. 1 - Different line-frequency commutated rectifiers topologies:
a) boost converter T1;
b) modified boost converter T2;
c) auxiliary boost converter T3;
d) modified auxiliary boost converter T4

2) Topologies T3 and T4: These converters operate in a slightly different way. Switch S_a is turned on around the input voltage zero crossing in an interval in which the input current is zero. This causes a partial discharge of the auxiliary capacitor C_a through L_a (associated to N_a in T4) [6,7]. When the switch is turned off this discharge phase continues through D_a until the current goes to zero. The energy absorbed from C_a is sent to the output capacitor, given a boost characteristic to these topologies. At the end of this interval, capacitor C_a is charged at a fraction of the output voltage which causes a premature diode bridge turn on. In T4 the series switch S_i, which can be substituted by two SCRs in the diode bridge, can be turned on immediately after the S_a turn off. The input current increases in a resonant way (inductance L in T3 or L_i+L_a+2.M in T4) charging C_a until diode D turns on. After that, the inductor resonates with (C_a+C) until the input current zeroes.

II. Topologies Comparison

The comparison among these low-frequency commutated rectifiers is carried out by taking into consideration the following aspects: the input current harmonic content, the switch current and voltage stresses, the output voltage regulation capability and the physical dimension of the magnetic components. As far as this aspect is concerned, let us consider the cross-section-window-area product which, for an inductor, is given by:

$$A_w A_e = \left(\frac{N}{k_R}\frac{I_{Lrms}}{J}\right)\left(\frac{L\, I_{Lpk}}{N B_{max}}\right) = \frac{K_L}{k_R\, J\, B_{max}} \qquad (1)$$

where B_{max} is the maximum flux density, J is the desired current density and k_R the window filling coefficient. Parameter K_L will be computed for each magnetic component of all the topologies of Fig. 1 as an estimation of their volumes. Note also that $K_L = L \cdot I_{Lrms} \cdot I_{Lpk}$ is also related to the inductor peak energy, i.e.:

$$K_L = \frac{2\,E_L}{CF} \text{ where } CF = I_{pk}/I_{rms} \text{ is the crest factor.}$$

A MATLAB program was used to generate the input current waveform of each circuit of Fig. 1 in order to find more quickly the parameter values which allow compliance with the standards. The analysis has been performed at $230V_{RMS}$ input voltage (value imposed by the standards) and with a constant output voltage. In the current harmonic calculation no margin was considered in comparing their amplitudes with the corresponding limits given by the IEC 1000-3-2.

The result of such analysis is reported in Table I for different power levels. The following informations are reported: output voltage (U_o), switch voltage stress (U_{spk}/U_o), input filter inductor value (L), auxiliary inductor value L_a used for the discharge phase in T3 (the value for T4 represents the fraction of the total inductance value corresponding to winding N_a in Fig. 1d), input peak and RMS current values (I_{gpk}, I_{grms}), switch peak and rms current values (I_{Spk}, I_{Srms}), auxiliary inductor peak and rms current values (I_{Lapk}, I_{Larms}), inductor coefficients K_L, auxiliary inductor coefficient K_{La} for T3 or transformer coefficient K_T for T2, total harmonic distortion (THD), displacement factor (cos(ϕ)) and power factor (PF).

A. Compliance with low-frequency harmonic standards.

As far as the input current harmonic content is concerned, the design criteria for the rectifiers shown in Fig. 1 change as a function of the rated power. For a power range below 600W, 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, as reported in the IEC 1000-3-2 standard [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 absorbed current so as to stay outside the Class D template, also

shown in Fig. 2, for at least 5% of the line half period.

If the required output power is higher than 600 W, the load is considered in class A, no matter the current waveform. The converters have no longer the aim of modifying the input current to stay out of the class D template, but to improve the current harmonic content. In particular, reduction of the third harmonic is accomplished at the expense of an increase of the high order harmonics. Note that the converters of Fig. 1 are progressively more efficient in the sense that topologies T3 and T4 produce less high frequency harmonic increase as compared to topologies T1 and T2.

B. Current and voltage stresses

The switch current stress is directly related to the switch on time. The latter should be kept as small as possible in order to maintain the size of the auxiliary magnetic component at a reasonable level, the only exception being represented by topology T4 in which the two magnetic components are combined in the same core. The current stress of topology T2 is much lower than the other's, thanks to the transformer's turns ratio. Instead, the voltage stress is equal to the output voltage for T1 and T3, while is greater for T2 and T4.

C. Output voltage regulation

This aspect is correlated to the maximum switch on time which is limited for all the topologies except T4 for the reasons explained above. As a consequence, a limited output voltage regulation is achieved at least up to a power level for which the passive L-C rectifier (without the switching unit) achieves an output voltage value below the reference value (only a boost action is provided), while practically no regulation is possible for an input voltage variation. This problem, is partially overcome for topology T4, since no auxiliary magnetic component is used.

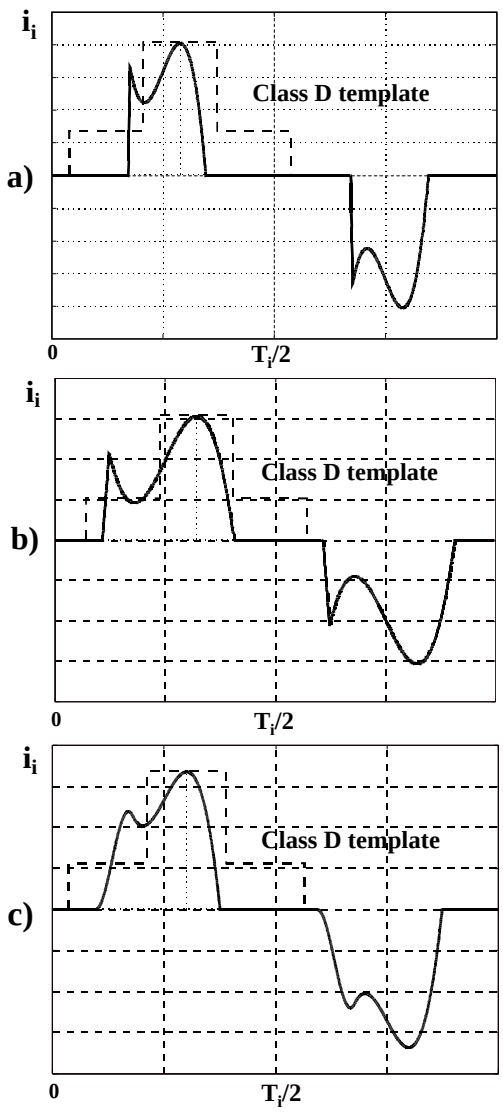


Fig. 2 - Typical input waveforms of the line-frequency commutated rectifiers. a) T1; b) T2; c) T3, T4

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

P_o [W]	U_o [V]	U_{Spk}/U_o [V]	L [mH]	L_a [mH]	I_{gpeak} [A]	I_{grms} [A]	I_{Speak} [A]	I_{Lapeak} [A]	I_{Larms} [A]	K_L [mJ]	K_{La} [mJ]	THD	$\cos(\phi_1)$	PF
300 - P	291.8		15.5		4.16	1.81				117		0.827	0.932	0.718
300 - T_1	314.9	1	4		4.59	1.81	2.95			33.3		0.945	0.995	0.723
300 - T_2	314.9	1.088	3.4		4.81	1.83	0.65	0.65	0.051	30	5.7	0.973	0.994	0.713
600 - P	294.83		6.5		8.62	3.68				206		0.869	0.938	0.708
600 - T_1	303.4	1	4.5		8.84	3.63	1.33			144		0.9	0.964	0.716
600 - T_2	306	1.059	4		8.86	3.61	0.53	0.53	0.031	128	1.9	0.908	0.971	0.719
600 - T_3	315.2	0	3	0.8	8.26	3.48	7.44	7.74	1.01	86.3	6.2	0.876	0.994	0.747
600 - T_4	326.6	<1	3.4	0.22	7.94	3.51	29.4	29.4	3.14	103		0.904	0.999	0.741
900 - P	258.96		18.5		9.8	5.12				929		0.523	0.861	0.763
900 - T_1	290	11	9		10.01	4.82	2.66			434		0.612	0.952	0.812
900 - T_2	301	1.294	7		9.8	4.72	1.24	1.24	0.158	324	46.6	0.625	0.977	0.829
900 - T_3	316.2		5.2	1	8.34	4.54	24.4	24.4	3.35	197	82	0.587	0.999	0.861
900 - T_4	325.8	<1	6.3	1	8.77	4.59	31.65	31.65	4.52	423		0.601	0.996	0.854
1200 - P	230.7		28		12.2	6.97				2381		0.314	0.799	0.762
1200 - T_1	273.6	1	16		11.21	6.16	3.53			1105		0.435	0.923	0.846
1200 - T_2	297.8	1.326	9.6		10.71	5.8	2.25	2.25	0.394	596	228	0.449	0.986	0.9
1200 - T_3	310.6		6.8	1.2	9.51	5.65	34.4	34.4	4.84	366	200	0.412	0.997	0.922
1200 - T_4	325.56	<1	8.8	1.9	10.24	5.78	32.5	32.5	5.48	904		0.468	0.994	0.9

P = passive; T_1 = boost; T_2 = modified boost; T_3 = auxiliary boost; T_4 = modified auxiliary boost

From the results of Table I, some comments can be done. In the low power range (up to 600W) the passive topology uses an inductance significantly higher than the active ones. For the four studied topologies, T3 presents the minimum overall magnetic volume (K_L+K_{La}) but the difference among them is not so relevant. On the other hand, in comparison to T1 and T2, the high current peak through the switch is an important drawback for T3 and T4.

As the power increases, the reduction in the magnetic volume of T3 is remarkable, and even considering the current stress, this topology seems to be indicated for this power range.

T4 present magnetic volume comparable to T1 and T2 and current stress comparable to T3. But its advantage over the others, as mentioned before, is the possibility of regulating the output voltage in a wider load range.

III. Experimental Results

In order to verify the results obtained by simulation a prototype of each converter was tested.

A. *Topology T1*

This circuit was built having the following specifications:

$U_g = 230V_{rms}$ $U_o = 284V$ $P_o = 600W$
$L = 14mH$ $C = 2x470\mu F$

The turn-on delay T_d of the gate signal was set to 2.2ms. Fig. 3 shows the main converter waveforms at nominal conditions: as we can see the input current waveform well agree with the simulation results. The input current spectrum is shown in the same figure and the first 25 harmonics are listed in Table II: harmonics 19th and 23th are the closest to the corresponding limits, thus confirming the simulation results.

Table II - Measured input current harmonics at nominal conditions

I_n	Harmonics [A_{rms}]	Class A limits [A_{rms}]
I_1	2.75	
I_3	0.856	2.30
I_5	0.612	1.14
I_7	0.134	0.77
I_9	0.320	0.40
I_{11}	0.112	0.33
I_{13}	0.182	0.21
I_{15}	0.114	0.15
I_{17}	0.106	0.132
I_{19}	0.105	0.118
I_{21}	0.0606	0.107
I_{23}	0.0804	0.098
I_{25}	0.0488	0.09

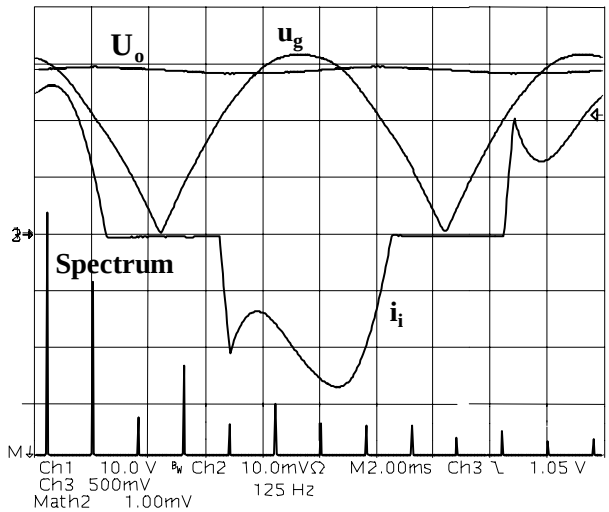


Fig. 3 - Rectified input voltage (100V/div), input current (2A/div) and its spectrum (0.4A_{rms}/div) at $U_i = 225V_{rms}$, $U_o = 284V$ and $P_o = 600W$

B. *Topology T2*

The prototype was built with the these specifications:
$U_i = 230\ V_{rms}$, $U_o = 300V$, $P_o = 900W$, $L = 5.3$ mH, $C = 2x470\mu F$

When the switching unit is activated, the current waveform modifies as shown in Fig. 4, 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 III

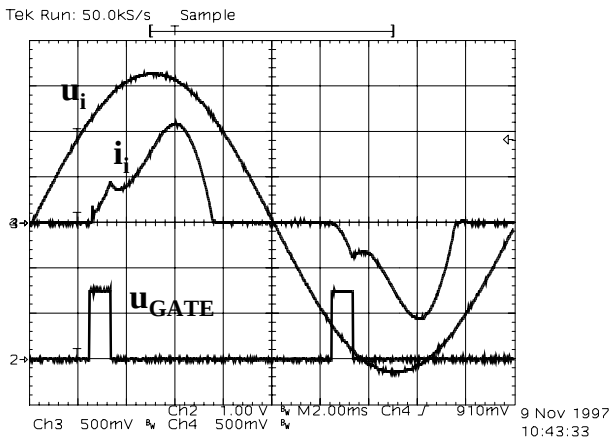


Fig. 4 - Input voltage U_i (100V/div), input current i_i (5A/div) and gate signal u_{gate} (10V/div) ($U_i = 230V_{rms}$, $P_o = 900W$)

The 5.3 mH inductor adopted in the prototype allows to increase the power level to 900 W, without exceeding the standard harmonic limits. It is worth noting that 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%.

Table III – Measured input current harmonics at different output power levels for T2 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

C. Topology T3

In order to verify the results, a prototype was built having the following specifications:

$U_i = 230V_{rms}$ $U_o = 292V$ $P_o = 900W$

$L = 6$ mH $L_a = 1$ mH $C_a = 44\mu F$ $C = 470\mu F$

The rectifier input current and voltage waveforms at nominal conditions are shown in Fig. 5 and the corresponding current harmonics amplitudes are reported in Table IV, together with Class A limits. As can be seen, the input current waveform well agrees with the simulation results and all the harmonics are well below their limits. The employed switch on-time is 70µs, as we can see from Fig. 6 which reports the auxiliary capacitor voltage u_{Ca}, the auxiliary inductor current i_{La} and the switch gate drive signal at the beginning of the line half period.

The switch current stress is about 20 A, while its voltage stress is limited by the output voltage. The measured efficiency is 96%, while the power factor is 0.85, and it is almost totally dominated by the distortion factor DF which results 0.86 (THD = 59%).

Table IV Harmonic content of T3 topology

I_n	Harm.	Class A limits
	[A_{rms}]	[A_{rms}]
I_3	2.18	2.30
I_5	0.73	1.14
I_7	0.46	0.77
I_9	0.028	0.40
I_{11}	0.14	0.33
I_{13}	0.05	0.21
I_{15}	0.058	0.15
I_{17}	0.048	0.132
I_{19}	0.022	0.118
I_{21}	0.026	0.107

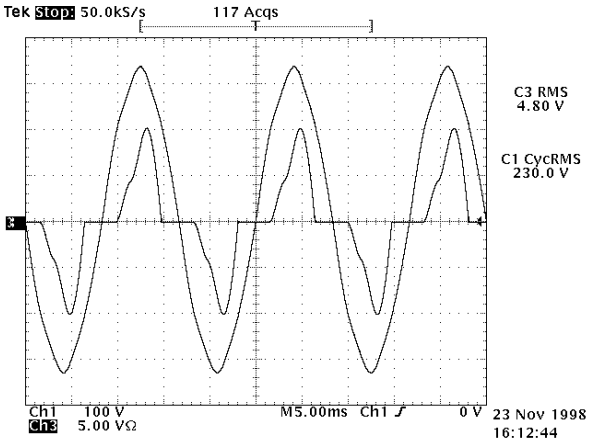


Fig. 5 – Input voltage U_i (100V/div), input current i_i (5^A/div) ($U_i = 230V_{rms}$, $P_o = 900W$)

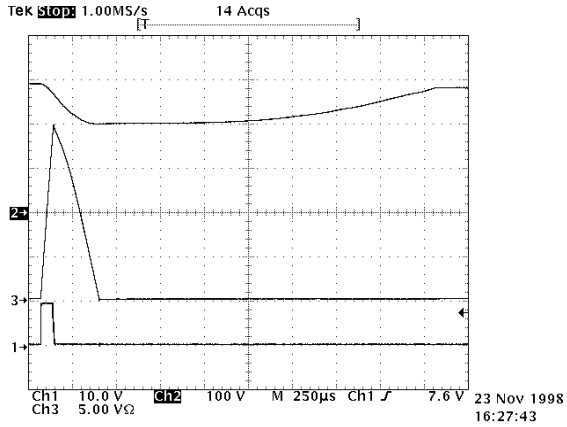


Fig. 6 – Auxiliary capacitor voltage u_{Ca} (100V/div), auxiliary inductor current i_{La} (5 A/div) and switch gate drive signal (10V/div) ($U_i = 230V_{rms}$, $P_o = 900W$)

D. Topology T4

Thus, in order to verify the results obtained, a prototype was built, according to this topology, having the following specifications:

$U_i = 230V_{rms}$ $U_o = 300V$ $P_o = 1200W$

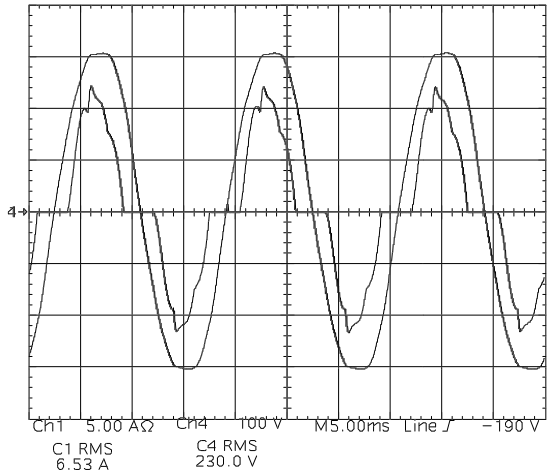
$L = 7.2$mH $L_a = 1.8$mH $C_a = 66\mu F$ $C = 1410\mu F$

The input current and voltage waveforms at nominal conditions are shown in Fig. 7. The distortion observed at the peak of the current happens if the output voltage becomes lower than the input voltage. This turns off diode D, turning on D_a. The effective inductance is reduced, changing the resonant frequency. This behavior stays while $U_o > u_g$. Nevertheless the current distortion complies with the standards as can be seen in Fig. 8 which shows the current harmonic spectrum.

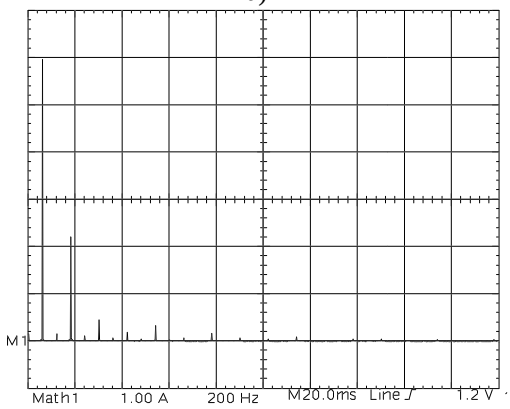
The peak current at the auxiliary switch reaches 35 A.

The converter efficiency is about 92%. This relatively low value is due to the losses in the coupled inductance used in the prototype which was not designed for such power. The measured power factor is 0.89.

As expected from the theoretical analysis, the output voltage can be maintained regulated down to no load condition and also for an increased input voltage up to $250V_{RMS}$, by simply reducing the switch on-time.



a)



b)

Fig. 7 - a) Input voltage U_i (100V/div), input current i_i (10A/div) and b) input current spectrum ($1A_{RMS}$/div) ($U_i = 230V_{rms}$, $P_o = 904W$).

IV. Conclusions

Low-frequency switched rectifiers are a simple and cheap solution to achieve compliance with EMC regulations in AC/DC power supplies for household and general-purpose applications.

A comprehensive comparison of different low-frequency switched rectifiers topologies allows the designer to select the right topology for a given application and power level.

In the power range below 600W, in which class D limits can be applied, it seems to be a good choice the topologies T1 and T2, once they modify the current waveform, allowing to apply class A limits, while using relatively small inductances.

Above this power, topology T3 seems to be the more convenient, considering the smaller overall inductance size. If the output voltage regulation is an important request, topology T4 presents the best behavior, allowing to stabilize the voltage down to no load.

V. References

1. IEC 1000-3-2, First Edition 1995-03, International Electrotechnical Commission, 3, Genève, Switzerland.
2. M. Jovanovic, D. E. Crow, "Merits and Limitations of Full-Bridge Rectifier with LC Filter in Meeting IEC 1000-3-2 Harmonic-Limit Specifications," IEEE Applied Power Electronics Conf. Proc. (APEC), March 1996, pp. 354-360.
3. I. Suga, M. Kimata, Y. Ohnishi, R. Uchida. “New Switching Method for Single-Phase AC to DC Converter,” Power Conversion Conference Proc. (PCC), 1993, pp. 93-98.
4. L. Rossetto, G.Spiazzi, P. Tenti, “Boost PFC with 100 Hz Switching Frequency Providing Output Voltage Stabilization and Compliance with EMC Standards”, Industry Applications Society Annual Meeting, St. Louis, 1998, pp. 1567-1573.
5. G. Spiazzi, S. Buso, "A Line-Frequency Commutated Rectifier Complying with IEC 1000-3-2 Standards," IEEE Applied Power Electronics Conf. (APEC), Dallas, USA, March 1999.
6. J. A. Pomilio, G. Spiazzi, "A Double-Line-Frequency Commutated Rectifier Complying with IEC 1000-3-2 Standards," IEEE Applied Power Electronics Conf. (APEC), Dallas, USA, March 1999.
7. J. A. Pomilio, G. Spiazzi, "A Low-Inductance Line-Frequency Commutated Rectifier Complying with IEC 1000-3-2 Standards," IEEE Power Electronics Specialists Conf. (PESC), Charleston, USA, June 1999.