OPTIMUM DESIGN OF A DC-DC HB ZVS ASYMETRIC PWM CONVERTER

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***Abstract* – This work deals with the study and development of a project methodology, for optimum design of the losses and magnetic volume of a DC-DC half-bridge ZVS asymmetric PWM converter. Theoretical, simulation and experimental results are presented based on a prototype with 500 Watts output power.**

KEYWORDS

ZVS comutation, Half bridge converter assimectric, optimization of losses.

I. INTRODUCTION

The converter DC-DC HB ZVS-PWM presented in Fig.1 it is a good solution for power supplies of telecommunications, for presenting the following characteristics:

- Commutation under null voltage;
- Few components for being a structure in half bridge;
- Simple control strategy;
- Low losses for conduction;
- Low interference of EMI;
- Voltage stress in the switches is equal the input voltage;

This converter is derived from the isolated half bridge dc-dc converter with the inclusion of the resonant inductor L_r, resonant capacitors C_1 and C_2. These two capacitors are the switches capacitances. D_1 and D_2 are the body drain diodes. D_{C1} and D_{C2} are clamping diodes with the purpose to reduce overshooting and ringing voltages on the output rectifier diodes.

II. ASYMETRIC CC-CC ZVS-HB CONVERTER

The asymmetric ZVS-HB CC-CC converter, is a converter that possesses a resonant stage during the commutation period through a circuit series LC that allows the so much when the switches goes into conduction as the blockade under null voltage. The prototype developed in this article it possesses the following specifications:

Input voltage : $V_i = 400$ V
Output voltage: $V_o = 54$ V
Commutation frequency: fs = 100 kHz
Output power: 500 Watts.
The structure of the converter can be seen below in Fig.1.

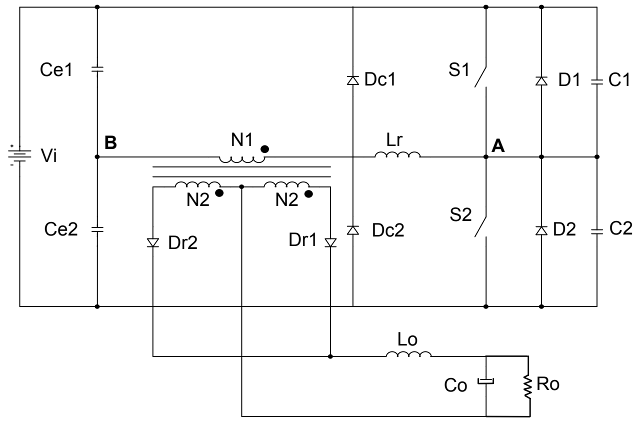

Fig. 1 - DC-DC HB ZVS-PWM

OPTIMUM DESIGN.

The optimum design proposed by this work it felt in several stages. Firstly decreased her the volume of the magnetic elements to begin for the resonant inductor. They were still made studies to minimize the value of the resonant inductor in way to this to assist the commutation specifications ZVS under critical load and to guarantee the cycle of smaller work than the maximum stipulated by the planner. They were lifted up equations and traced the way curves to find acceptable values of Lr (resonant inductor) in function of the transformer ratio for several situations of operation of output voltage and load.

RESONANT INDUCTOR L_R AND TRANSFORMER TURN RATIO

The necessary condition for commutation ZVS is had it is that the energy stored in the inductor L_r it is enough to discharge the capacitor completely C_1 loaded with the voltage $(1\text{-}D)\cdot V_i$, in the most critical case. This condition is represented by the equation (1).

$$\frac{1}{2}\cdot L_r\cdot\left[\left(2\cdot D\cdot I_o'\right)\right]^2>\frac{1}{2}\cdot C_{eq}\cdot\left[\left(1-D\right)\cdot Vi\right]^2 \tag{1}$$

Where, D is duty cycle, C_{eq} is equivalent capacitance of mosfets and I_o' is nominal load current reflected to the primary side of transformer. Then :

$$L_r > \left[\frac{(1-D_{crit})\cdot Vi}{2\cdot D_{crit}\cdot I_o{}'_{crit}}\right]^2\cdot C_{eq} \tag{2}$$

Therefore in the worst case the converter operates with D = D_{crit} when $I_o' = I_o'_{crit}$. Dcrit is determined by the equation (3).

$$D_{crit} = \frac{1}{2}-\frac{1}{2}\cdot\sqrt{1-2\cdot\Delta D_{crit}-2\cdot q} \tag{3}$$

The duty cycle loss and the static earnings are certain for (4) and (5) respectively.

$$\Delta D_{crit} = \frac{4\cdot f_s\cdot I_o{}'_{crit}\cdot L_r}{Vi} \tag{4}$$

$$q = \frac{n\cdot(V_o+Vd_f)}{Vi} \tag{5}$$

Where, V_{df} is the voltage fall on the diodes of the exit rectifier and 'n' is the transformer ratio.

The current of load critical referenced to the primary of the transformer is given for (113).

$$I_o{}'_{crit} = \frac{I_o\cdot ZVS\%}{n} \tag{6}$$

Where, ZVS% is the minimum percent of load where the converter have a commutation ZVS (zero voltage switching). Substituting (4), (5) and (6) in the equation (3):

$$D_{crit} = \frac{1}{2}-\frac{1}{2}\cdot\sqrt{1-2\cdot\frac{4\cdot f_s\cdot\frac{I_o\cdot ZVS\%}{n}\cdot L_r}{Vi}-2\cdot\frac{n\cdot(V_o+Vd_f)}{Vi}} \tag{7}$$

substituting (6) and (7) in (2), we obtain (8).

$$L_r > \left[\frac{\left[1-\left[\frac{1}{2}-\frac{1}{2}\cdot\sqrt{1-2\cdot\frac{4\cdot f_s\cdot\frac{I_o\cdot ZVS\%}{n}\cdot L_r}{Vi}-2\cdot\frac{n\cdot(V_o+Vd_f)}{Vi}}\right]\right]\cdot Vi}{2\cdot\left[\frac{1}{2}-\frac{1}{2}\cdot\sqrt{1-2\cdot\frac{4\cdot f_s\cdot\frac{I_o\cdot ZVS\%}{n}\cdot L_r}{Vi}-2\cdot\frac{n\cdot(V_o+Vd_f)}{Vi}}\right]\cdot\left(\frac{I_o\cdot ZVS\%}{n}\right)}\right]^2\cdot C_{eq} \tag{8}$$

Starting from the equation (7):

$$D_{\max} = \frac{1}{2}-\frac{1}{2}\cdot\sqrt{1-2\cdot\frac{4\cdot f_s\cdot\frac{I_o}{n}\cdot L_r}{Vi}-2\cdot\frac{n\cdot(V_o+Vd_f)}{Vi}} \tag{9}$$

Solving (9) as function of Lr obtains:

$$L_{r\_D\max}(n) \ge \frac{1}{8}\cdot\frac{\left[1-2\cdot\frac{n}{Vi}\cdot Vs-2\cdot\frac{n}{Vi}\cdot Vd_f-(1-2\cdot D_{\max})^2\right]}{f_s\cdot I_o}\cdot n\cdot Vi \tag{10}$$

Solving numerical the equations (8) and (10) for several load values and output voltage we obtain them it curves presented in Fig.2. Where the curves ' a', ' b', ' c', ' of, ' e' and ' f' are regarding equation (10) and the curves ' g', ' h' and ' i' are referring the equation (8). Para the curves traced by the equation (10) the acceptable values are below the curve while for the equation (8) the values of resonant inductance acceptable are above the curve.

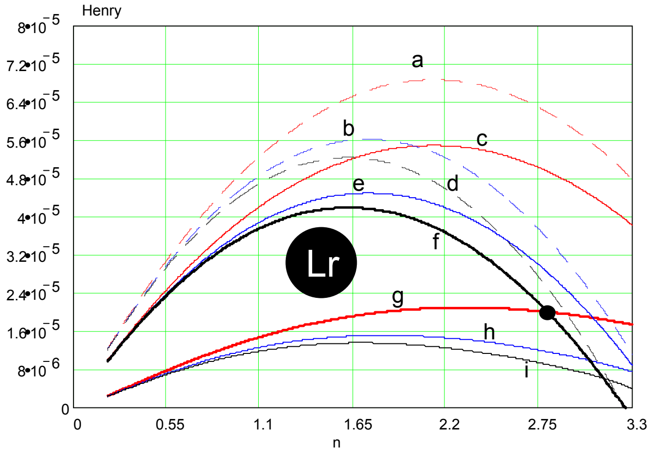

Fig. 2 - Resonant inductor Lr x n - I Saw = 400V.

In order to assist the commutation specifications ZVS under critical load and maximum value of the work cycle the value of the resonant inductor should be in the area among the curves marked by the letters ' f' and ' g'.

SEMICONDUCTORS STRESS AND LOSSES

SWITCHES CONDUCTION LOSSES

The effective currents in the switches S1 and S2 are given below by the equations, where, Io'_nom is the load current contemplated for the primary of the transformer.

$$I_{S1\_eficaz} = \left[Io'\_nom\cdot 2\cdot(1-D)\cdot\sqrt{D}\right] \tag{11}$$

$$I_{S2\_eficaz} = \left[Io'\_nom\cdot 2\cdot(D)\cdot\sqrt{1-D}\right] \tag{12}$$

The losses for conduction in the switches S1 and S2 are presented for (13).

$$P = Rds_{on}\cdot\left[I_{S1\_eficaz}{}^2+I_{S2\_eficaz}{}^2\right] \tag{13}$$

The duty cycle is defined for (14).

$$D = \frac{1}{2}-\frac{1}{2}\cdot\sqrt{1-2\cdot\frac{4\cdot f_s\cdot\frac{I_o}{n}\cdot L_r}{Vi}-2\cdot\frac{n\cdot(V_o+Vd_f)}{Vi}} \tag{14}$$

Rated current I given by equation (15)

$$I_o{}'_{nom} = \frac{Po}{n\cdot(V_o+Vd_f)} \tag{15}$$

Substituting (14) and (15) in (13) and tracing the result of the losses in the switches in function of the transformer ratio obtains Fig.3.

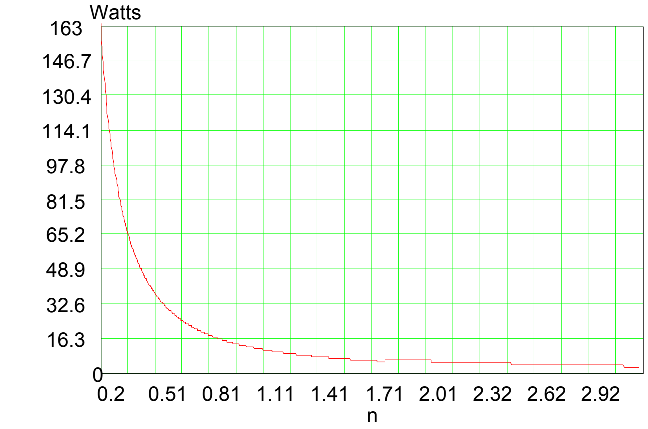

Fig.3 Losses x Transformer ratio

Therefore, the conclusion is that the value of the transformer ratio should be the possible largest for it is had minimum losses in the switches and therefore the chosen value for the resonant inductor, as well as, the value of the transformer ratio should be in the point of intersection of the curves 'f ' and 'g ' of Fig.2.

BRIDGE'S DIODES VOLTAGE STRESS

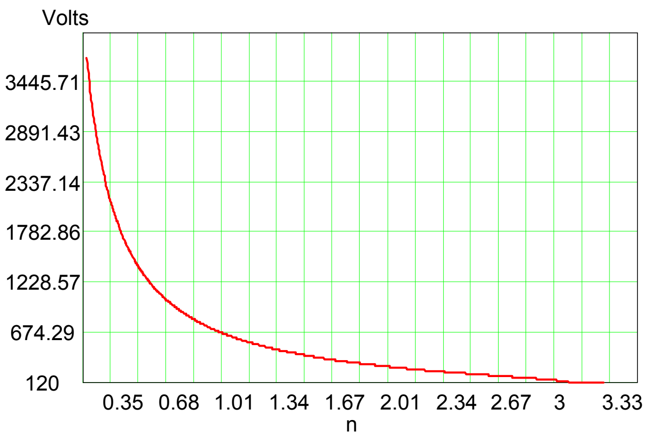

Fig.4 Bridge's diodes stress x ' n'.

As we can see in Fig.4 as well as the stress current in the switches, the voltage stress of the diodes of the rectifier bridge also decrease with the increase of the transformer ratio.

OPTIMUM DESIGN FOR THE LOSSES IN THE TRANSFORMER

COPPER LOSSES

Through a study of the current densities or even of the area of occupation of the copper of the rolling up of the transformer, and starting from the effective currents of each rolling up, it is arrived to occupation indexes that minimize the values of the losses in the copper given a density of maximum current stipulated by the planner.

$$P_{Cooper} = \frac{\rho \cdot MLT}{A_w \cdot K_w} \cdot \left(\frac{I_1^{\,2}}{\alpha_1} + \frac{I_2^{\,2}}{\alpha_2} + \frac{I_3^{\,2}}{\alpha_3} \right) \tag{16}$$

$$\alpha_1 + \alpha_2 + \alpha_3 = 1 \tag{17}$$

Using the method of Multipliers of Lagrange to meet the inflection point (minimum) of the system of equations described above we arrived to the values of the occupation coefficients wanted.

$$\alpha_m = \frac{n_m \cdot I_m}{\sum_1^3 n_j \cdot I_j} \tag{18}$$

Where Im is the effective current of each rolling up.

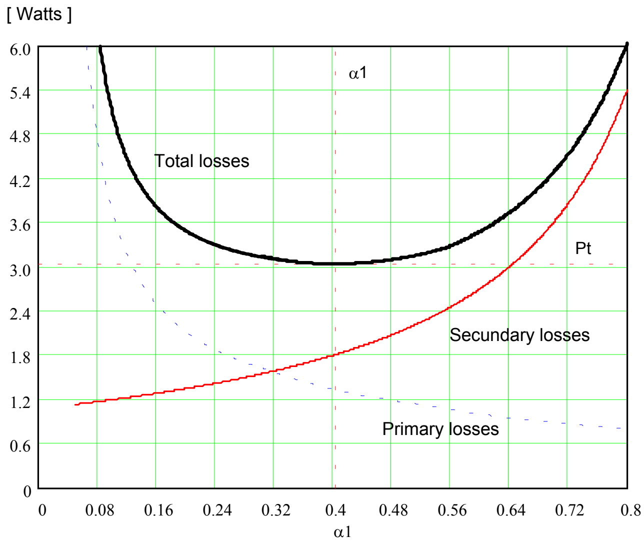

Fig.5 Copper losses for α1 values

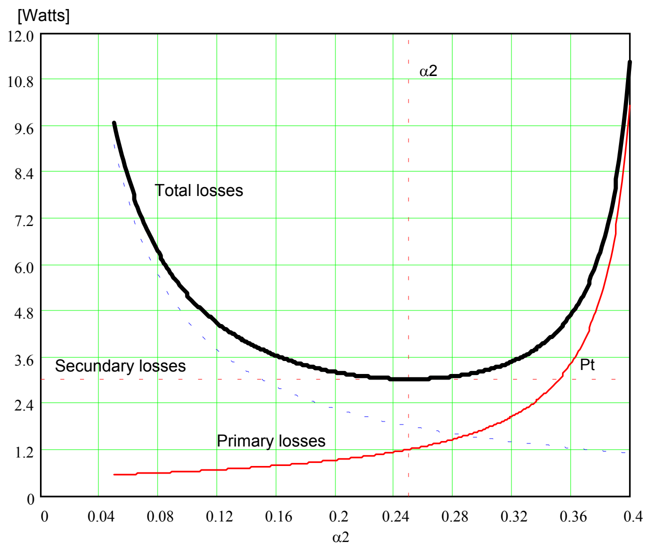

Fig.6 Copper losses for α2 values

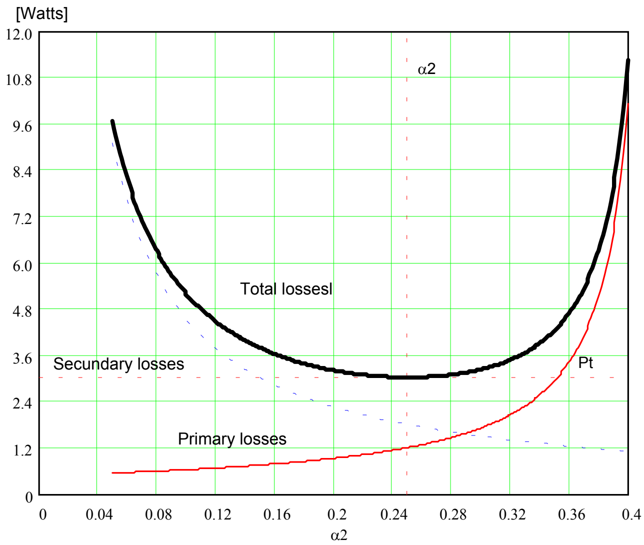

Fig.7 Copper losses for α3 values

CORE LOSSES

To the we write the losses in the core and losses in the copper in function of flux density we found the following equations:

$$P_{Cooper} = \frac{\rho \cdot MLT \cdot \lambda^2}{A_e^{\,2} \cdot A_w \cdot K_w} \cdot I_{Total}^{\;2} \cdot \frac{1}{\Delta B^2} \cdot 10^8 \tag{19}$$

$$P_{núcleo} = C_m \cdot V_e \cdot f_S^{\;X} \cdot \Delta B^Y \cdot 10^{-3} \tag{20}$$

Where Y is the coefficient of Steinmetz and X and Cm they are the constants of loss of the material lifted at laboratory.

$$\begin{aligned} Cm &= 7.9292 \cdot 10^{-3} \\ x &= 1.4017 \\ y &= 2.3294 \end{aligned} \tag{21}$$

Being added (19) and (20) deriving with regard ΔB arrived the (22). Solving the equation (22) to find the value of field density that minimizes the total losses in the transformer we found (23).

$$\frac{\partial P_{Total}}{\partial \Delta B} = \frac{\partial P_{Cooper}}{\partial \Delta B} + \frac{\partial P_{core}}{\partial \Delta B} = 0 \tag{22}$$

Solving the equation (129) we arrived the (130).

$$\Delta B_{Optimum} = \left(2 \cdot 10^{11} \cdot \frac{\rho \cdot MLT \cdot \lambda^2}{y \cdot Cm \cdot V_e \cdot f_s^{\,X} \cdot A_e^2 \cdot A_w \cdot K_w} \cdot I_{Total}^{\;2} \right)^{\frac{1}{y+2}} \tag{23}$$

THE CORE

To choose the core the following conditions for the geometric constants Kg and Kgfe they should be assisted.

$$K_g = \left(\frac{A_w \cdot A_e^2}{V_e^{\frac{2}{y}} \cdot MLT} \right)^{\frac{y}{y+2}} \tag{24}$$

$$K_g \geq \frac{10^8 \cdot \left(\dfrac{y \cdot C_m \cdot f_s^{\,x}}{2 \cdot 10^{11}} \right)^{\frac{2}{y+2}} \cdot \left(1 + \dfrac{2}{y} \right)}{P_{otimo} \cdot \left(\dfrac{\rho \cdot \lambda^2 \cdot I_{total}^{\;2}}{K_w} \right)^{\frac{-y}{y+2}}} \tag{25}$$

$$K_{gfe} = \left(\frac{MLT \cdot A_e^{\,y} \cdot A_w^{\,y+1}}{V_e} \right)^{\frac{1}{y+2}} \tag{26}$$

$$K_{gfe} \geq \frac{\lambda \cdot I_1 \cdot 10^4}{\left(2 \cdot 10^{11} \cdot \dfrac{\rho \cdot \lambda^2}{y \cdot Cm \cdot f_s^{\,X} \cdot K_w} \cdot I_{Total}^{\;2} \right)^{\frac{1}{y+2}} \cdot K_w \cdot J_{1\max} \cdot \alpha_i} \tag{27}$$

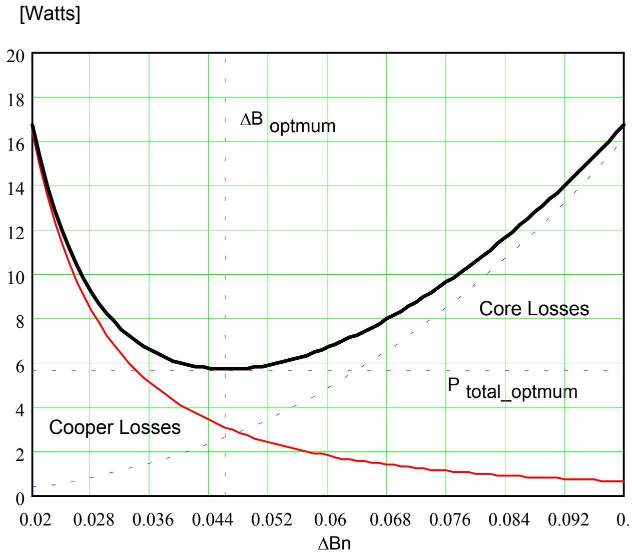

Fig.8 Transformer losses

EXPERIMENTAL RESULTS

Starting from the presented calculations it was arrived to the following values for the resonant inductor and transformer.

Lr = 20 μH (the own linkage inductance of the transformer was used);

Transformer ratio : n= 2.7 ; (core EE-55 IP12)

N1= 50 turns (33 x 33AWG);

N21 = 18 turns (56 x 33AWG);

N22 = 18 turns (77 x 33AWG).

With base in the harmonic content of the current in the primary of the transformer, shown in the Fig.9, in order to avoid the increase of the losses in the copper due to the effect skin, the wire 33 AWG was chosen for the construction of the transformer with ferrite core.

The dotted line indicates the frequency of the harmonic component below which effect skin won't be had due to the use of the wire 33AWG.

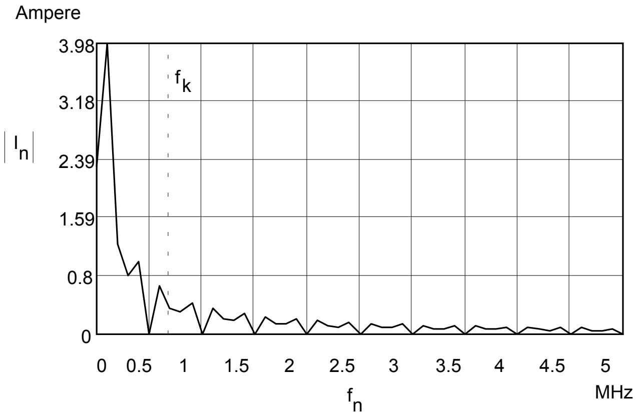

Fig.9 harmonic content for primary current.

Fig.10 presents the forms of wave of the voltage drain-source and drain current during the commutation period. It can be noticed the turn on of switches under null voltage.

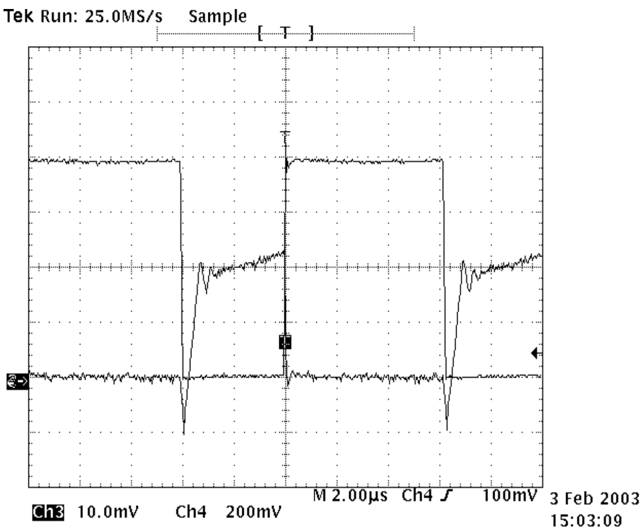

Fig.10 Commutation

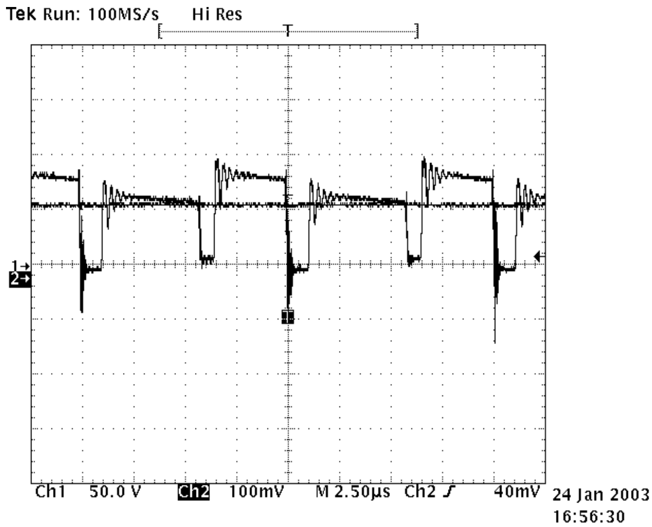

Fig.11 output voltage and voltage before output filter.

The theoretical revenue was of 96%. THE prototype got to reach 93% for nominal load and a pick of 95% to 38% of the nominal load. Fig.9 presents the experimental efficiency of the converter

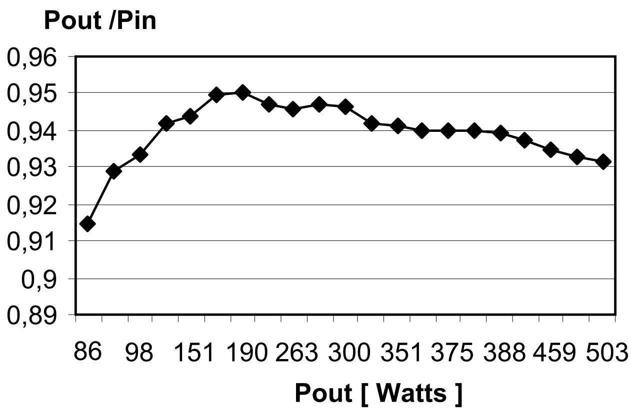

Fig.12 Eficience

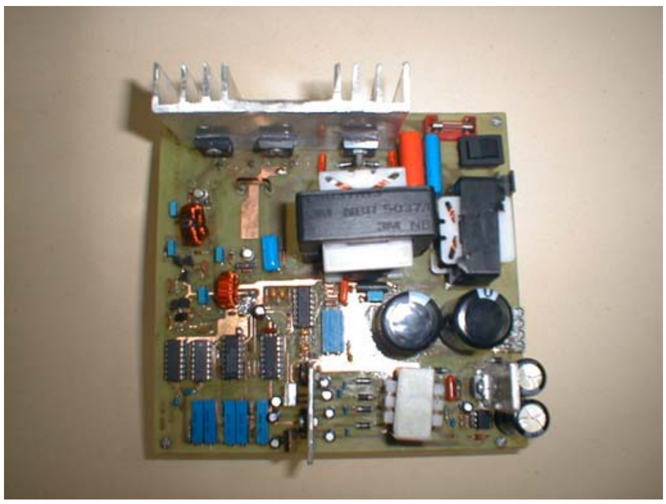

Fig 13. Board On-Top

III. CONCLUSION

The used methodology demonstrated to be quite efficient and the results obtained quite satisfactory.

The use of the linkage inductance of the transformer as resonance inductance and the intrinsic capacitors of the mosfets turned the less voluminous converter and they helped to reduce the losses.

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