A ZCS Class-E Amplifier Used as a Single-Stage High-Power-Factor Electronic Ballast for Fluorescent Lamps

Mario Ponce, Rubén Vásquez, Jaime Arau
Centro Nacional de Investigación y Desarrollo Tecnológico (CENIDET)
P.O. Box 5-164, 62050.
Cuernavaca, México.
Telephone: +52-73-18-77-41, Fax: +52-73-12-24-34,
E-mail: ponce@cenidet.edu.mx

J. Marcos Alonso
Universidad de Oviedo
Tecnología Electrónica
Campus Viesques s/n, 33204
Gijón, Spain.
E-Mail: marcos@ate.uniovi.es

Abstract.- **In this paper a Zero-Current-Switching (ZCS) class-E amplifier is proposed as fluorescent lamp ballast. The ZCS class E amplifier features lower voltage stress across the switch than the Zero-Voltage-Switching (ZVS) class E amplifier. This is an important advantage which allow to supply the circuit from the American main voltage (Vac=120 V), maintaining only one switch for the ballast. The operation and characteristics of the ZCS class E amplifier as electronic ballast are analysed in this paper. The presented study includes selection of the more suitable resonant tank, power factor correction using the "modified valley fill" technique and parametric analysis of the ZCS class E amplifier.**

I. INTRODUCTION

The use of class E amplifiers to implement fluorescent lamp ballasts presents several interesting advantages as high power density, high efficiency and low component counts. However, the main disadvantage of these converters is the high voltage stress across the switch, which can reach a value as high as four times the input supply voltage. Until now, this disadvantage has limited the input voltage range of the class E amplifiers below 100 V, thus avoiding the very high voltage stress across the switch obtained when supplying from a high main voltage. For instance, if a class E amplifier is supplied from American main voltage (120 Vrms), the voltage stress across the switch would be V_{smax}=4*120*1.4142=648 V, making necessary the use of a switch with a voltage rate higher than 600 V.

In order to solve this problem, in [1] and [2] the use of two switches in series, a bipolar and a MOSFET, operating as one only switch has been proposed. In this way, the resultant switch could bear a voltage stress up to 2 kV. However, with this solution the main problem of the converter is not resolved, since a considerable high voltage stress across the switch is maintained. Moreover, an extra floating switch is needed, which is not ground referenced, making necessary the use of an isolated driver.

An alternative solution to this problem of high voltage stress across the switch is to operate the switch with zero current switching instead of zero voltage switching provided by a conventional class E amplifier. With this goal, the use of a class E amplifier with a shunt inductor has been proposed in [3], as shown in Fig. 1, which is the dual circuit of the conventional ZVS class E amplifier. Thus, the switch voltage and current stresses under optimum operation for the ZCS class E amplifier are V_{smax}= 2.89Vcc and Ismax=3.56Icc, whereas for the ZVS class E amplifier were V_{smax}= 3.56Vcc and Ismax=2.89Icc. A voltage stress decrease of 37.3 % is then obtained thanks to the zero current switching feature.

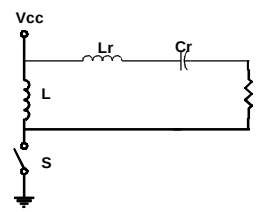


Fig. 1. ZCS class E amplifier.

In this paper the operation of the ZCS class E inverter as fluorescent lamp ballast is analysed. The presented study is focused on the selection of the resonant tank more appropriate to handle the fluorescent lamps on the effect of parameter variations in the circuit operation. A solution to incorporate power factor correction feature to the circuit is also proposed, this solution is based on the modified valley fill technique, thus maintaining the simplicity of the circuit [4].

II. PARAMETRIC ANALYSIS

The analysis presented in [3] considers only the optimum operation of the ZCS class E amplifier, this is, the operation with both zero current and slope switching. This is a very particular analysis, since there exists only one operation point satisfying this optimum switching. Thus, the class E amplifier could easily loss zero current switching operation if one or more circuit parameters are changed. The use of a diode in series with the switch maintains the ZCS operation within a circuit parameter range, but the zero slope switching is lost.

In this paper the circuit shown in Fig. 1, with a unidirectional current switch formed by a controlled switch with a series diode, is analysed under any operation conditions. A parametric analysis is performed to

determine the variation range of the different circuit parameters maintaining zero current switching operation.

Figure 2 shows the equivalent circuit of the ZCS class E amplifier used in this analysis. In this circuit the Cr capacitor has been split into two different capacitors: Cf which in tuned with Lf and C. The resonant tank Lf-Cf is tuned at the switching frequency F. In this analysis an ideal unidirectional switch S1 is considered, which can only handle current in one direction. In this way, when the switch is reverse biased it behaves as an open switch without regarding the control signal. For this reason, two different duty cycles are obtained; firstly, the duty cycle obtained from the control signal applied to the switch, which is considered equal to 50 % in this analysis, and secondly, the duty cycle given by the zero crossing of the current switch (D'). Figure 3 illustrates the two different duty cycles. In this figure the interval times t'_{on} and t'_{off} correspond to the unidirectional switch duty cycle and the interval times t_{on} and t_{off} correspond to the control signal duty cycle. The unidirectional switch will start conducting at instant ωt_c , referred to the instant time t_0, this angle can be expressed as follows:

$$\omega t_c = 2\pi(1-D') \tag{1}$$

A. Voltage and Current in L and in S1

By inspecting Fig. 2 the voltage and current in the switch are given by the following expressions:

$$i_c(\omega t) = i_L(\omega t) - i_R(\omega t) \tag{2}$$
$$v_c(\omega t) = Vcc - v_L(\omega t) \tag{3}$$

When the switch is in the off state, it behaves as an open circuit. Therefore:

$$i_c(\omega t) = 0, \qquad 0 \le \omega t \le \omega t_c \tag{4}$$
$$i_L(\omega t) = i_R(\omega t) = I_R \operatorname{sen}(\omega t + \phi) \qquad 0 \le \omega t \le \omega t_c \tag{5}$$

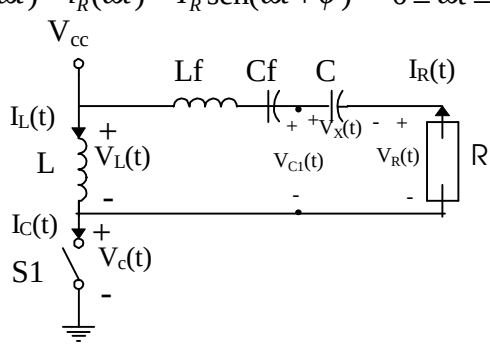


Fig. 2. ZCS Class E amplifier equivalent circuit.

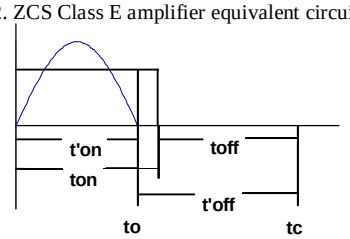


Fig. 3. Current waveform and signal control in the switch.

$$v_L(\omega t) = \omega L \frac{di_L(\omega t)}{d(\omega t)} = \omega L I_R \cos(\omega t + \phi) \quad 0 \le \omega t \le \omega t_c \tag{6}$$
$$v_c(\omega t) = Vcc - \omega L I_R \cos(\omega t + \phi) \quad 0 \le \omega t \le \omega t_c \tag{7}$$

When the switch is in the on state, it behaves as a short circuit, then:

$$v_c(\omega t) = 0, \quad \omega t_c \le \omega t \le 2\pi \tag{8}$$
$$v_L(\omega t) = Vcc \quad \omega t_c \le \omega t \le 2\pi \tag{9}$$

Current through inductor L can be calculated as follows:

$$i_L(\omega t) = \frac{1}{\omega L}\int_{\omega t_c}^{\omega t} v_L(\omega t)\, d(\omega t) + i_L(\omega t_c +) \tag{10}$$

where $\mathbf{i_L(\omega t_c+)}$ is the instantaneous inductor current during the turn-on transition of the switch, and its value is given by:

$$\begin{aligned} i_L(\omega t+) = i_L(\omega t_c -) &= I_R \operatorname{sen}[2\pi(1-D') + \phi] \\ &= I_R \operatorname{sen}(\phi - 2\pi D') \end{aligned} \tag{11}$$

By integrating (10) and using (11), the following expression is obtained:

$$\begin{aligned} i_L(\omega t) = \frac{Vcc}{\omega L}&[\omega t - 2\pi(1-D')] + \\ &I_R \operatorname{sen}(\phi - 2\pi D') \quad \omega t_c \le \omega t \le 2\pi \end{aligned} \tag{12}$$

$$\begin{aligned} i_c(\omega t) = \frac{Vcc}{\omega L}&[\omega t - 2\pi(1-D')] + \\ &I_R\{\operatorname{sen}(\phi - 2\pi D') - \operatorname{sen}(\omega t + \phi)\} \quad \omega t_c \le \omega t \le 2\pi \end{aligned} \tag{13}$$

B. Resonant tank voltages

Since the resonant tank Lf-Cf is tuned at the switching frequency, it behaves as an ideal filter, allowing only the current due to the fundamental component of the voltage $v_L(t)$ to pass through it. Therefore, the voltage shown as $v_{cl}(t)$ in fig. 2 will be equal to the fundamental component of $v_L(t)$. This fundamental component is a sinusoidal waveform given by the following equation:

$$v_{C1}(\omega t) = \rho V_R \operatorname{sen}(\omega t + \phi_1) \tag{14}$$

where $\mathbf{V_R}$ is the voltage amplitude across the load and:

$$\rho = \sqrt{1 + \frac{XC^2}{R^2}} \tag{15}$$

$$\phi_1 = \phi + \psi = \phi + tan^{-1}\left(\frac{XC}{R}\right) \tag{16}$$

Regarding the voltage $\mathbf{v_{C1}(t)}$ shown in Fig. 2, its fundamental component is calculated as follows:

$$\begin{aligned} V_{c1} &= \int_0^{2\pi(1-D')} [V_{cc} - \omega L I_{R1} \cos(\omega t + \phi)] \operatorname{sen}(\omega t + \phi_1)\, d(\omega t) \\ V_{c1} &= \frac{\omega L V_R}{4\pi R}[\cos(2D - \phi - \phi_1) - \cos(\phi + \phi_1) + (2D - 4\pi)\operatorname{sen}\psi] - \\ &\quad - \frac{V_{cc}}{\pi}[\cos(D - \phi_1) - \cos\phi_1] \end{aligned} \tag{17}$$

where D=2πD'. By equalling (17) to the amplitude given by (14) and solving for $\mathbf{V_R}$:

$$V_R = \frac{4V_{cc}R}{\omega L}\frac{[\cos(D-\phi_1)-Cos\phi_1]}{\left[\cos(2D-\phi-\phi_1)-\cos(\phi+\phi_1)+(2D-4\pi)\,\mathrm{sen}\,\psi-\frac{4\pi R\rho}{\omega L}\right]} \tag{18}$$

Equation (18) can also be expressed as follows:

$$V_R = \frac{4V_{cc}R}{\omega L}h(D,\phi,\phi_1,\psi,R,\rho,L) \tag{19}$$

where:

$$h(D,\phi,\psi,R,\rho,L) = \frac{\cos(D-\phi_1)-\cos\phi_1}{\cos(2D-\phi-\phi_1)-\cos(\phi+\phi_1)+(2D-4\pi)\,\mathrm{sen}(\psi)-\frac{4\pi R\rho}{\omega L}} \tag{20}$$

Since the fundamental component of voltage $\mathbf{v_{c1}(\omega t)}$ has only a sinusoidal term, the cosinusoidal component must be equal to zero, this is:

$$0 = \frac{1}{\pi}\int_0^{2\pi(1-D')} V_c(\omega t)\cos(\omega t+\phi_1)\,d(\omega t) \tag{21}$$

$$0 = \frac{1}{\pi}\int_0^{2\pi(1-D')}\left[V_{cc}-\frac{\omega L V_R}{R}\cos(\omega t+\phi)\right]\cos(\omega t+\phi_1)\,d(\omega t)$$

By integrating (21) and solving for V_R, it is finally obtained:

$$V_R = \frac{4RV_{cc}}{\omega L}\frac{[\mathrm{sen}\,\phi_1+\mathrm{sen}(D-\phi_1)]}{[(2D-4\pi)\cos\psi+\mathrm{sen}(2D-\phi-\phi_1)+\mathrm{sen}(\phi+\phi_1)]} \tag{22}$$

Equation (22) can also be expressed in the following way:

$$V_R = \frac{4RV_{cc}}{\omega L}g(D,\phi,\phi_1,\psi,R) \tag{23}$$

where:

$$g(D,\phi) = \frac{\mathrm{sen}(\phi_1)+\mathrm{sen}(D-\phi_1)}{(2D-4\pi)\cos\psi+\mathrm{sen}(2D-\phi-\phi_1)+\mathrm{sen}(\phi+\phi_1)} \tag{24}$$

By equalling (20) and (24), substituting $\phi_1=\phi+\psi$ and expanding, the numerators and denominators of the functions *g( )* and *h( )* can be expressed as follows:

$$\begin{aligned}
g_{nin} &= \mathrm{sen}\,\phi(\cos\psi-\mathrm{sen}\,D\,\mathrm{sen}\,\psi-\cos D\cos\psi)+\\
&\quad\cos\phi(\mathrm{sen}\,\psi+\mathrm{sen}\,D\cos\psi-\cos DSen\psi)\\
h_{num} &= \mathrm{sen}\,\phi(\mathrm{sen}\,\psi-\cos D\,\mathrm{sen}\,\psi+\mathrm{sen}\,D\cos\psi)+\\
&\quad\cos\phi(\cos D\cos\psi-\cos\psi+\mathrm{sen}\,D\,\mathrm{sen}\,\psi)\\
g_{den} &= \mathrm{sen}^2\phi(\mathrm{sen}\,\psi\cos 2D-\mathrm{sen}\,2D\cos\psi-\mathrm{sen}\,\psi+(2D-4\pi)\cos\psi)+\\
&\quad\mathrm{sen}\,\phi\cos\phi(\cos\psi+\cos\psi-2\,\mathrm{sen}\,2D\,\mathrm{sen}\,\psi-2\cos 2D\cos\psi)+\\
&\quad\cos^2\phi[\mathrm{sen}\,2D\cos\psi-\cos 2D\,\mathrm{sen}\,\psi+\mathrm{sen}\,\psi+(2d-4\pi)\cos\psi]\\
h_{den} &= \mathrm{sen}^2\phi\left(\cos\psi-\cos 2D\cos\psi-\mathrm{sen}\,2D\,\mathrm{sen}\,\psi+2D\,\mathrm{sen}\,\psi-4\pi\,\mathrm{sen}\,\psi-\frac{4\pi R\rho}{\omega L}\right)+\\
&\quad\cos^2\left(-\cos\psi+\mathrm{sen}\,2D\,\mathrm{sen}\,\psi+\cos 2D\cos\psi+2D\,\mathrm{sen}\,\psi-4\pi\,\mathrm{sen}\,\psi-\frac{4\pi R\rho}{\omega L}\right)+\\
&\quad\mathrm{sen}\,\phi\cos\phi(2\,\mathrm{sen}\,2D\cos\psi-2\cos 2D\,\mathrm{sen}\,\psi+2\,\mathrm{sen}\,\psi)
\end{aligned} \tag{25}$$

By assigning a new notation to each term multiplying to $\mathrm{Sen}^2\phi$, $\mathrm{Cos}^2\phi$, $\mathrm{Sen}\phi\mathrm{Cos}\phi$, as follows:

$$\begin{aligned}
q_1 &= \cos\psi-\mathrm{sen}\,DSen\psi-\cos D\cos\psi\\
q_0 &= \mathrm{sen}\,\psi+\mathrm{sen}\,DCos\psi-\cos D\,\mathrm{sen}\,\psi\\
r_2 &= \mathrm{sen}\,\psi\cos 2D-\mathrm{sen}\,2D\cos\psi-\mathrm{sen}\,\psi+(2D-4\pi)\cos\psi\\
r_1 &= \cos\psi+\cos\psi-2\,\mathrm{sen}\,2D\,\mathrm{sen}\,\psi-2\cos 2D\cos\psi\\
r_0 &= \mathrm{sen}\,2D\cos\psi-\cos 2D\,\mathrm{sen}\,\psi+\mathrm{sen}\,\psi+(2d-4\pi)\cos\psi\\
s_2 &= \cos\psi-\cos 2D\cos\psi-\mathrm{sen}\,2D\,\mathrm{sen}\,\psi+2D\,\mathrm{sen}\,\psi-4\pi\,\mathrm{sen}\,\psi-\frac{4\pi R\rho}{\omega L}\\
s_1 &= 2\,\mathrm{sen}\,2D\cos\psi-2\cos 2D\,\mathrm{sen}\,\psi+2\,\mathrm{sen}\,\psi\\
s_0 &= -\cos\psi+\mathrm{sen}\,2D\,\mathrm{sen}\,\psi+\cos 2D\cos\psi+2D\,\mathrm{sen}\,\psi-4\pi\,\mathrm{sen}\,\psi-\frac{4\pi R\rho}{\omega L}
\end{aligned} \tag{26}$$

Equations corresponding to the functions *g( )* and *h( )* can be expressed in the following way:

$$\begin{aligned}
g_{num} &= q_1\,\mathrm{sen}\,\phi+q_0\cos\phi\\
g_{den} &= r_2\,\mathrm{sen}^2\phi+r_1\,\mathrm{sen}\,\phi\cos\phi+r_0\cos^2\phi\\
h_{num} &= q_0\,\mathrm{sen}\,\phi-q_1\cos\phi\\
h_{den} &= S_2\,\mathrm{sen}^2\phi+S_1\,\mathrm{sen}\,\phi\cos\phi+S_0\cos^2\phi
\end{aligned} \tag{27}$$

Taking into account the following condition:

$$\frac{g_{num}}{g_{den}} = \frac{h_{num}}{h_{den}}$$

which is the same as:

$$g_{num}h_{den}-g_{den}h_{num} = 0$$

by substituting (27), multiplying and expanding:

$$\alpha_3\,\mathrm{sen}^3\phi+\alpha_2\,\mathrm{sen}^2\phi\cos\phi+\alpha_1\,\mathrm{sen}\,\phi\cos^2\phi+\alpha_0\cos^3\phi = 0 \tag{28}$$

where:

$$\begin{aligned}
\alpha_3 &= q_1s_2-r_2q_0\\
\alpha_2 &= q_1s_1+q_0s_2+r_2q_1-r_1q_0\\
\alpha_1 &= q_1s_0+q_0s_1+r_1q_1-r_0q_0\\
\alpha_0 &= q_0s_0+r_0q_1
\end{aligned} \tag{29}$$

Multiplying both sides of (28) by $\cos^{-3}\phi$, we obtain:

$$\alpha_3 tan^3\phi+\alpha_2 tan^2\phi+\alpha_1 tan\phi+\alpha_0 = 0 \tag{30}$$

Dividing (30) by $\boldsymbol{\alpha_3}$:

$$tan^3\phi+\frac{\alpha_2}{\alpha_3}tan^2\phi+\frac{\alpha_1}{\alpha_3}tan\phi+\frac{\alpha_0}{\alpha_3} = 0 \tag{31}$$

Since $\boldsymbol{\alpha_1 = \alpha_3}$ and $\boldsymbol{\alpha_0 = \alpha_2}$:

$$tan^3\phi+\frac{\alpha_2}{\alpha_3}tan^2\phi+tan\phi+\frac{\alpha_2}{\alpha_3} = 0 \tag{32}$$

Factorising (32), we obtain:

$$\left(tan\phi+\frac{\alpha_2}{\alpha_3}\right)\left(tan^2\phi+1\right) = 0$$

The first factor contains the real root of ϕ, therefore:

$$\phi = arctan\left(-\frac{\alpha_2}{\alpha_3}\right) = arctan\left(-\frac{\alpha_0}{\alpha_1}\right) \tag{33}$$

Once the ϕ angle has been obtained, it is possible to calculate any parameter of the class E amplifier under any operation condition.

C. Calculation of the D' duty cycle

The current through the switch is given by (13). The zero crossing of the switch-diode current occurs for an angle equal to 2π. Therefore:

$$0 = \frac{\mathrm{Vcc}}{2\mathrm{XL}\pi\mathrm{D}'}+I_R\left(\mathrm{sen}(\phi-2\pi D')-\mathrm{sen}(\phi)\right) \tag{34}$$

In this equation the only independent variable is the duty cycle D'. Consequently, with this equation and by using numerical methods the duty cycle D' can be calculated.

D. Voltage and current stresses in the switch

The peak current value is obtained by equalling to zero the derivative of (13) and solving for the angle. This angle will indicate the instant of maximum current and is given by the following equation:

$$\theta_{\mathrm{MI}} = -\cos^{-1}\left(\frac{V_{cc}}{I_R XL}\right) - \phi + 2\pi \tag{35}$$

By using (35) in (13) the maximum value of the switch current is obtained.

Regarding the peak voltage across the switch, the same procedure is followed with equation (7), in this case the peak voltage angle is the following:

$$\theta_{\mathrm{MV}} = \pi - \phi \tag{36}$$

E. Average input current

The average current delivered by the power supply to the ZCS class E amplifier is obtained by integrating (4) and (13) within a period. The result is given by the following equation:

$$\mathrm{Icc} = \frac{\mathrm{D}'^2 \pi V_{CC}}{XL} + \frac{I_R}{2\pi\left(\cos\phi - \cos(\phi - 2\pi D') + 2D'\pi\,\mathrm{sen}(\phi - 2D'\pi)\right)} \tag{37}$$

F. Input and output power and power capability

The input power is given by the following equation:

$$\mathrm{Pin} = \mathrm{V_{cc}} I_{cc} \tag{38}$$

and the output power can be calculated as follows:

$$Po = \frac{I_R^2}{2} R \tag{39}$$

The power capability is a quantity which evaluate the ratio of the output power to the voltage and current stresses and is given by the following expression:

$$Cp = \frac{Po}{I_{CM} V_{CM}} \tag{40}$$

$\mathbf{I_{CM}}$ being the maximum current through the switch and $\mathbf{V_{CM}}$ the maximum voltage across the switch.

Previous equations were introduced in a Mathematica program in order to evaluate the circuit behaviour under suboptimum operation and using a diode in series with the transistor. Using this program the operation of the class E amplifier under variations on the following parameters: R, L, equivalent series capacitance of the resonant tank C and switching frequency F was performed. In order to simplify the analysis, the different parameters were normalised using the optimum values [3]: R=1, F=1, Vcc=1, $L=(\pi^2+4)/4$ y $C=16/(2\pi^2(\pi^2+12))$.

As result of the performed analysis the following variations in the different elements of the class E amplifier were found to maintain ZCS operation: $0<R\leq 1$, $1\leq F\leq 1.15$, $1\leq L\leq 1.45$ and $1\leq C\leq 1.28$. Figures 4 and 5 illustrate the voltage and current waveforms in the switch obtained for variations in R. Similar waveforms can be obtained for the other parameters.

As conclusion of the performed analysis, the ZCS class E amplifier can accept quite well parameter variation, particularly in the load resistance R, without loosing ZCS operation, provided that a diode is used in series with the transistor. These results are similar to those obtained in [5] for the ZVS class E amplifier.

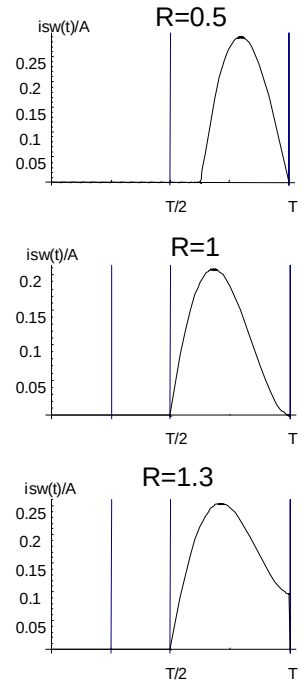


Fig. 4. Current waveform across the switch for variations of R.

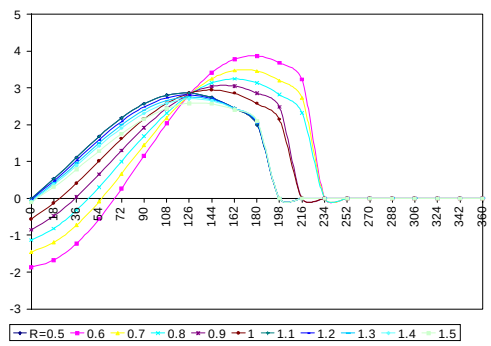


Fig. 5. Voltage waveform for variations of R.

III. SELECTION OF THE RESONANT TANK

The acceptance of high load variations is one of the important conditions in order to use an inverter as fluorescent lamp ballast. Thus, the inverter must tolerate load variations from a very high impedance load when the lamp is not ignited, to the nominal load impedance when the lamp is running. According to the performed parametric analysis for the ZCS class E amplifier, this topology can undergo load variations from the nominal load down to zero without loosing zero current switching, which is the opposite behaviour as necessary for a lamp ballast. In order to solve this problem an impedance inverter is used, thus reflecting a high load impedance as a low load impedance towards the ZCS class E amplifier. There exist two possibilities in order to implement the impedance inverter: the capacitive impedance inverter (Fig. 6a) and the LCC resonant tank (Fig. 6b). Both circuits are useful to perform the impedance inversion. However, to calculate the resonant tank elements two conditions must also be satisfied: a) to achieve lamp ignition and b) to supply the lamp at nominal power. For the LCC resonant tank, when trying to satisfy both conditions at constant switching frequency, the following constrain is obtained [6]:

$$V_1 = \sqrt{2P_L R_e} \tag{41}$$

Where V_1 is the fundamental component of the voltage applied to the resonant tank, P_L is the lamp power and R_e is the equivalent series resistance of the resonant tank and load.

For the following operation conditions: P_L=32 W, R_L=132 Ω, main voltage of 120V_{rms} and minimum lamp ignition voltage of 350 V, the constrain given by (41) can not be satisfied. Therefore, the LCC resonant tank can not be employed.

The capacitive impedance inverter avoid this constrain. Besides, it presents a higher voltage boost ratio than the LCC resonant tank, since the series capacitor is in series with the load instead of the input power supply as in the LCC resonant tank. Therefore, the capacitive impedance inverter is chosen as the more appropriate resonant tank for this topology.

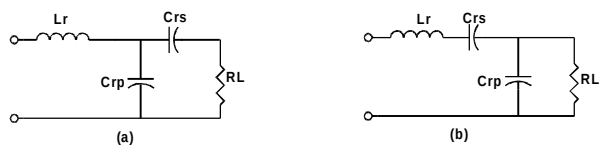


Fig. 6. Impedance inverters. (a) capacitive impedance inverter, (b) LCC resonant tank.

IV. POWER FACTOR CORRECTION

In order to simplify the structure of the presented ballast, the modified valley fill technique [4] is proposed to incorporate power factor correction feature to the ballast. Figure 7 illustrates the final schematic diagram of the proposed electronic ballast including the modified valley fill circuit. This technique consists of charging the filter capacitor (C_f) by using an extra winding in the resonant tank (Lr_2). The voltage across Lr_2 is rectified by means of the diode D_{vf}. In order to avoid capacitor C_f following line voltage, diode D_5 is used. In this manner, the main supply only deliver energy to the circuit when the main voltage is higher than voltage across capacitor C_f. With this arrangement a lower voltage across capacitor C_f will provide a higher input power factor, but also a higher lamp current crest factor. Consequently, a trade-off between power factor and lamp current crest factor is obtained. In order to attain a middle point for these two parameters a ratio between the peak line voltage and C_f voltage of V_{acp}/V_{cf}=0.3826, what corresponds to a line current conduction angle of 135 degrees within each line half cycle was selected.

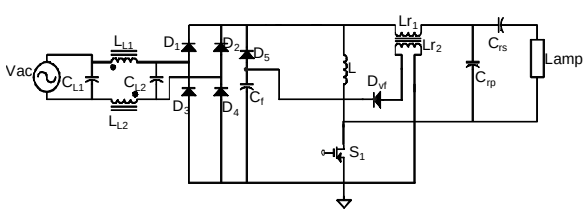


Fig. 7. Electronic ballast based on the class E amplifier.

V. SIMPLIFIED DESIGN OF A ZCS CLASS E AMPLIFIER

The procedure to calculate the different components of the class E amplifier is the following:

1. Calculation of the ZCS class E amplifier elements loaded a series resonant tank
2. Calculation of the capacitive impedance inverter elements
3. Calculations of the valley fill arrangement elements

The input specifications to calculate the different elements are the following: V_{ca}=127 Vrms, P_L=32 W, R_L=132, F=100 kHz and quality factor Q=10.

Using the equations given in [3] the following results were obtained: R_e=63.11 Ω, L=547 μH, Cr=2.521 nF y Lr=573.11 μH.

In order to calculate the elements of the capacitive impedance inverter, this circuit was simplified assuming an equivalent series circuit. Based on the circuit analysis, the relationship between the capacitive impedance inverter and a series resonant tank are obtained as follows:

$$R_e = \frac{R_L XC_{rp}^2}{R_L^2 + (XC_{rp} + XC_{rs})^2} \tag{42}$$

$$XCse = \frac{XC_{rp}\left(R_L^2 + XC_{rp}XC_{rs} + XC_{rs}^2\right)}{R_L^2 + (XC_{rp} + XC_{rs})^2} \tag{43}$$

Where Re and XCse are the equivalent series resistance and series capacitance and are known values, R_L is the equivalent lamp resistance and XC_{rp} and XC_{rs} are the reactances of C_{rp} and C_{rs} respectively.

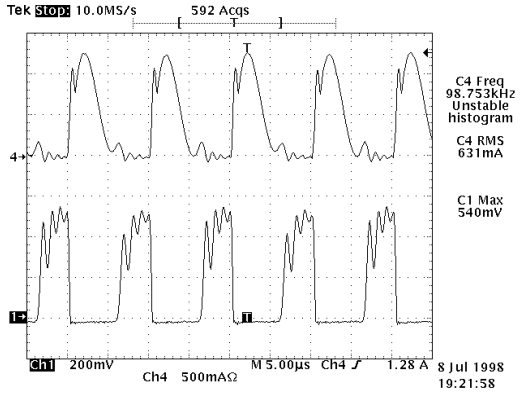
Solving (42) and (43) for XC_{rs} and XC_{rp}, the values of the capacitors can be obtained. In this example the following values are obtained: C_{rs}=766 pF and C_{rp}=1.77 nF.

The auxiliary winding in the resonant inductor used to improve the input power factor was calculated to provide a voltage across the capacitor Vcf=127*1.4142*0.3826=68.71 V. The obtained value was L_{r2}=30 uH.

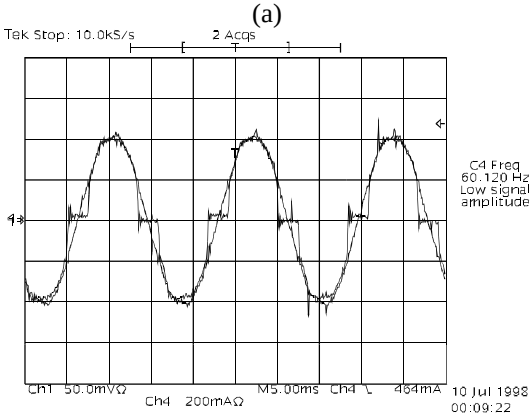
Finally, capacitor Cf was selected of 10 μF and the switch employed was and IGBT IRGBC30UD2 with a voltage rate of 600 V.

VI. EXPERIMENTAL RESULTS

Figure 8 shows the voltage and current stresses in the switch, and the main voltage and current obtained with the proposed electronic ballast. The maximum voltage stress across the switch is 540 V and the measured input power factor and total harmonic distortion were PF=97% and THD=25%.



(a)



(b)

Fig. 8. a) Switch voltage and current, b) Main voltage and current

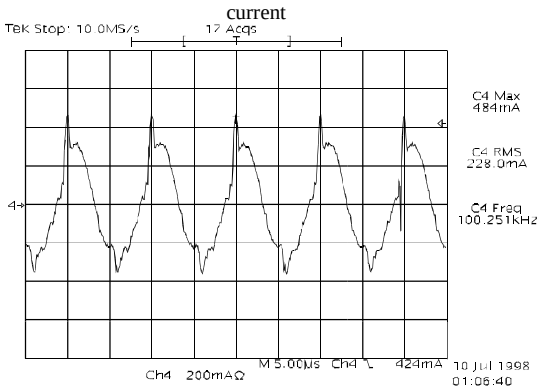


Figure 9. Lamp current

Figure 9 illustrates the lamp current waveform. The measured efficiency was 70 % and the measured lamp current crest factor was 1.8.

VII. CONCLUSIONS

In this paper the ZCS class E amplifier is proposed as fluorescent lamp electronic ballast featuring low voltage stress when compared to the ZVS class E amplifier. With this goal a parametric analysis of the ZCS class E amplifier was performed and the use of a capacitive impedance inverter as resonant tank was suggested. For the parametric analysis of the ZCS class E amplifier a diode in series with the switch was considered. With this analysis the variation range of the different circuit elements assuring zero current switching was determined and the important equations to evaluate the operation of the ZCS class E amplifier under any operation condition were derived. The power factor correction feature was implemented by using the modified valley fill technique.

Based on the presented study, electronic ballast for a 32W compact fluorescent lamp was designed and satisfactory experimental results were obtained. With the proposed circuit an electronic ballast able to be supplied from the American main voltage (Vcc=180 V) and using only one switch with a voltage stress below 600 V can be implemented.

VIII REFERENCES

[1] G. Lütteke, H. C. Raets. *"High Voltage High Frecuency Class-E Converter Suitable for Miniaturización"*. IEEE Transactions on Power Electronics, Vol. PE-1, No. 4, Oct. 1986, pp. 193-199.

[2] G. Lütteke, H. C. Raets. *"220-V Mains 500 kHz Class-E Converter Using a Bimos"*. IEEE Transactions on Power Electronics, Vol. PE-2, No. 3, Jul. 1987, pp. 186-193.

[3] M. Kazimierczuk. *"Class E Tuned Power Amplifier with Shunt Inductor"*, IEEE Journal of Solid-State Circuits, Vol. SC-16, No. 1, Feb 1981, pp. 2-7.

[4] M. H. Kheraluwala, S. A. El-Hamamsy, *"Modified Valley Fill High Power Factor Electronic Ballast for Compact Fluorescent Lamps"* IEEE Power Electronics Specialist Conference, 1995, pp. 10-14.

[5] F. H. Raab. *"Effects of Circuit Variations on the Class E Tuned Power Amplifier"*. IEEE Journal of Solid-State Circuits, Vol. SC-13, No 2, (Abril 1978), pp. 239-247.

[6] M. Ponce. *"Power Supply for Discharge Lamps based on Class E Amplifiers"*, Ph. D. Thesis, cenidet, 1999. In Spanish.