Electronic Ballast for Fluorescent Lamps Using Current Bidirecional 'Cuk Converter in Continuous Conduction Mode.

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Abstract -This paper presents a high power factor and high efficiency electronic ballast for fluorescent lamps. The topology consistis in a single power conversion stage, using a current bidirecional resonant 'Cuk converter with PWM control. The ballast operates in the continuous conduction mode. High efficiency is obtained by soft-switching techniques without the need of additional circuitry. Circuit description and design guidelines are presented. Experimental results are obtained from a 40W prototype.

I. INTRODUCTION

High frequency electronic ballast for fluorescent lamps present several advantages over magnetic ballast: higher power factor and efficiency, reduction or elimination of flicker effect, etc. [1].

The traditional high power factor electronics ballast circuit uses two power conversion stages. The first one is a power factor corrector (PFC), producing an input current with the same waveform of the respective. At the same regulates a DC voltage at its output.

The second stage is half-bridge resonant converter which produces at the output a high frequency quasi-sinusoidal voltage. Under these conditions it is expected an increase in the lamp life and in its luminous efficiency. Due to the lamp tube phosphorous coverage inertia, the stroboscope effect is practically eliminated at these frequencies [2], [3]. Moreover, the high frequency minimizes the size of the circuit capacitors and inductors, reducing volume and weigh.

Fig. 1 shows the two stages ballast circuit proposed in [4]. This topology presents some drawbacks, like the circuit complexity, need of two control circuits, higher losses due to the use of three switches.

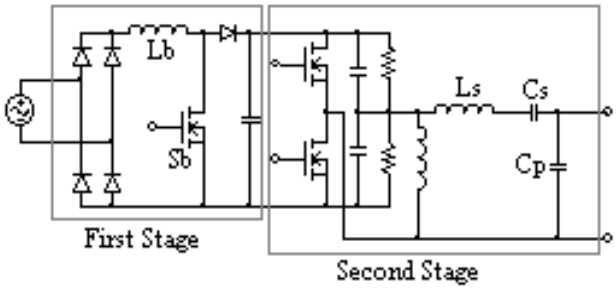

Fig. 1. Conventional two stages ballast.

The use of a single power stage conversion, maintaining both characteristics (high power factor and high frequency output) can provide better results in terms of circuit simplicity topology and control and also in its efficiency [5], [6], [7], [8], [9].

All of these circuits work in the discontinuous conduction mode (DCM). In fact, this mode of operation simplifies the control circuit once the switch responsible for controlling the input current can be driven by a constant duty-cycle signal. For many topologies this condition is sufficient to guarantee a high input power factor.

By the other hand, the operation in the conduction continuous mode (CCM) presents some advantages, especially the minimization of the necessary input electromagnetic interference filter (EMI). Also reduces the inductor and switch current stress. The main drawback seems to be control circuit complexity, but this problem can be solved by using commercial IC developed for PFC applications.

II. POWER FACTOR CORRECTION USING CUK CONVERTER

The basic circuit of a 'Cuk converter is shown in Fig. 2.

The capacitors C and Co values are sufficiency high so that their voltages can be considered constant. The voltage Vc is the sum of the voltages Vi and Vo, once the average voltage in L1 and L2, in steady-state, is zero. In the continuous conduction mode the switch M or the diode D are conducting.

Turning on the transistor M, the voltage applied to L1 and to L2 is Vi, and the respective currents increase. The diode is not conducting.

Turning off the switch, the continuity of the inductors currents is guaranteed by the diode conduction. At this situation, the voltage on both inductances is –Vo, and the currents start to decrease, charging C and Co.

The static output characteristic of this converter presents a buck-boost behavior. The output voltage is given by:

$$Vo = Vi \cdot \frac{\delta}{1-\delta} \tag{1}$$

Where δ is the transistor duty-cycle, which is the ratio between the turn-on time of the switch M and the switching period.

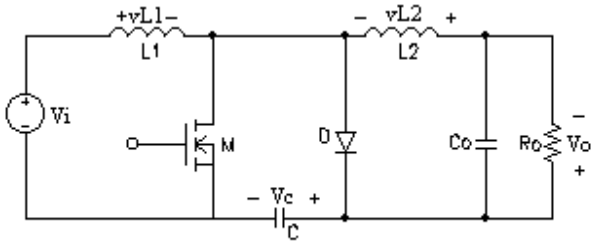

Fig. 2. Cuk Converter.

The 'Cuk converter can be used as a power factor pre-regulator working in both DCM or CCM. In the first case, like in the boost converter, using a simple PWM, constant duty-cycle control. In the CCM it is necessary to feedback the input current and change the duty-cycle according to the current error.

It is still possible to use eq. (1), but substituting the input voltage by the rectified line voltage.

$$vi(t) = Vm \cdot |sin(\omega t)| \tag{2}$$

III. THE RESONANT 'CUK CONVERTER

Let us substitute the output low-pass filter (L2 and Co in the Fig. 2) by a resonant filter, as shown in Fig. 3. If the resonance frequency is near the switching frequency, Fs, the voltage on capacitor Cp is practically sinusoidal. The chosen filter structure is a LCC connection due to its advantages for this kind of application [10].

With this circuit change, the currents through the switches M and D are no more unidirectional, then it is necessary to replace the diode by a current bidirectional switch, what means, a MOSFET.

As the converter operates in CCM, it is possible to use slow diodes in the rectifier bridge.

Ls, Cs e Cp form the LCC filter. Capacitor Cs blocks any DC current to flow through the lamp. The MOSFETs M1 and M2 allow the current bidirectionality due to its internal diodes.

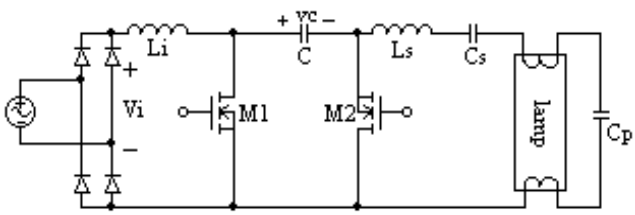

Fig. 3. Proposed Ballast.

The capacitor C voltage may be considered continuous and constant, because the input voltage variation now stays over the capacitor Cs (Cs<<C). In steady-stare the average voltage over the lamp is zero, once its average current is also null.

Considering the lamp resistive characteristic in high frequency, and neglecting the filaments resistance, it is possible to redraw Fig. 3 as shown in Fig. 4.

Before the lamp ignition, the resistance R is so high that the resonance frequency of the LCC circuit is simple given by:

$$Fr1 = \frac{1}{2\pi \cdot \sqrt{Ls \cdot Ceq}} \tag{3}$$

Where $Ceq = \dfrac{Cs \cdot Cp}{Cs + Cp}$.

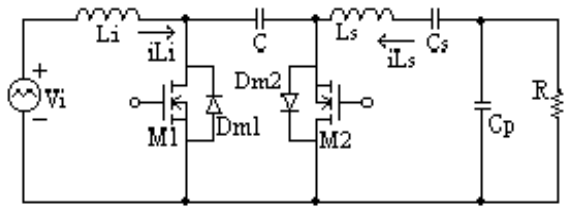

Fig. 4. Simplified ballast circuit.

This frequency must be slightly lower than the switching frequency, allowing to have the necessary lamp overvoltage for ignition, but still limiting the current.

After the lamp starting, the equivalent resistance, R, decreases to a value given by its power and voltage. This value must be significantly lower than the reactance of Cp, X_{Cp}, at the switching frequency, in way the new resonance frequency is:

$$Fr2 = \frac{1}{2\pi \cdot \sqrt{Ls \cdot Cs}} \tag{4}$$

This means that the resonance frequency decreases after the lamp starting, staying below the switching frequency and therefore decreasing the lamp voltage to its nominal value. Moreover, being the switching frequency higher than resonance frequency gives to the filter an inductive behavior, that is important to limit the lamp current and also to allow the soft-switching process, as will be shown.

IV. SOFT-SWITCHING

Fig. 5 we shows simulated waveforms of the switches voltage and current, and inductors current. Note that the input current is unidirectional, while the output one changes its polarity each switching cycle.

Due to the MOSFET drain-to-source capacitance, the transistor turn-off can be considered as a zero-voltage switching (ZVS).

At the beginning of t1 interval, M2 is turned-off. The current that flows through Dm1 is the difference between the inductors currents.

The input current starts to rise. The Ls current starts to decrease (in absolute value). The Dm1 current, therefore, decreases. At the moment these currents have the same value (with opposite direction) Dm1 turns-off, concluding interval t1. The M1 gate signal must have been applied during t1.

When this current difference inverts its polarity, M1 starts to conduct. The switch turn-on is a ZVS process., since the switch voltage is null. During interval t2, the Ls current continues to increase in a sinusoidal shape, and probably will invert its polarity. This will not happen only if the duty-cycle is very narrow. And this current stay with this direction until the t2 final. This interval finishes at the

moment the M1 gate signal is removed,. Then the diode Dm2 goes into conduction, beginning the t3 interval.

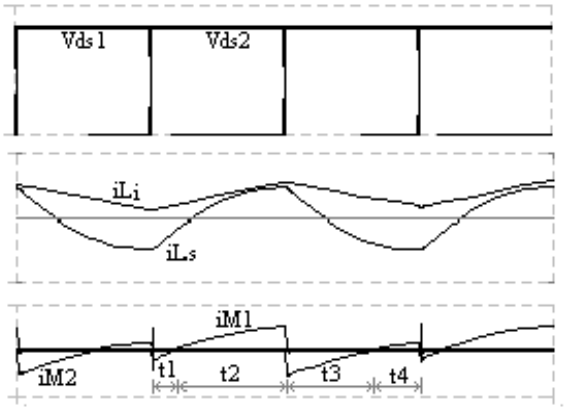

Fig. 5. Switches voltages and currents.

Intervals t3 and t4 are equivalent to t1 and t2, but considering M2 and its intrinsic diode Dm2. So, also for this switch there are soft-commutations.

The simplified circuit shown in Fig. 6 can be used to understand the beginning of intervals t1 and t3.

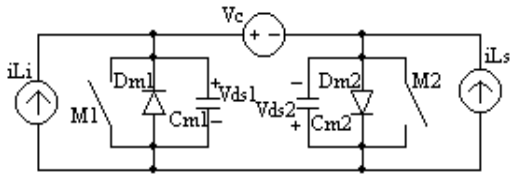

Figure 6- Equivalent circuit during the switch comutation.

After the t2 interval, M1 switch is turned-off. So, the capacitance Cm1, initially discharged, increase its voltage up to Vc and the capacitance Cm2, which initial voltage is Vc, discharges up to reach zero. The diode Dm2 begins to conduct. The Cm2 stored energy is transferred to Cm1, guaranteeing the Dm2 soft-commutation. At the same way, Dm1 turn-on is soft. At this situation, however, iLs is negative. So, the difference between iLi and iLs flows through the capacitances, charging Cm2 and discharging Cm1. Therefore, the M1 soft-switching process is more difficult than M2 one, especially when the duty-cycle tends to unity, because at this situation is possible that iLs does not invert its direction. In this case the soft-switching is lost.

This happens at a moment that corresponds to low input voltage and current values, what contributes to minimize the switching losses.

V. Control circuit

The CCM operation is implemented using the IC L4981A (SGS-THOMSON). This IC is a PWM controller, which internal block diagram is shown in Fig.7. The oscillation frequency is externally adjusted. The pulse width changes in accordance with the input current error. This error is generated by comparing the current reference, with the same waveform of the input voltage, and the actual current. The amplitude is determined by a PI controller that acts to maintain the capacitor C voltage constant. Input current is feedback at pin 9. The input voltage shape is feedback at pin 4. The capacitor C voltage is feedback at pin 14. The pin 12 allows to performer a soft-start.

As the switches commands are complimentary, the IC output and its complementary signal are used to drive the transistors.

The necessary isolation is provided by the driver IR2112.

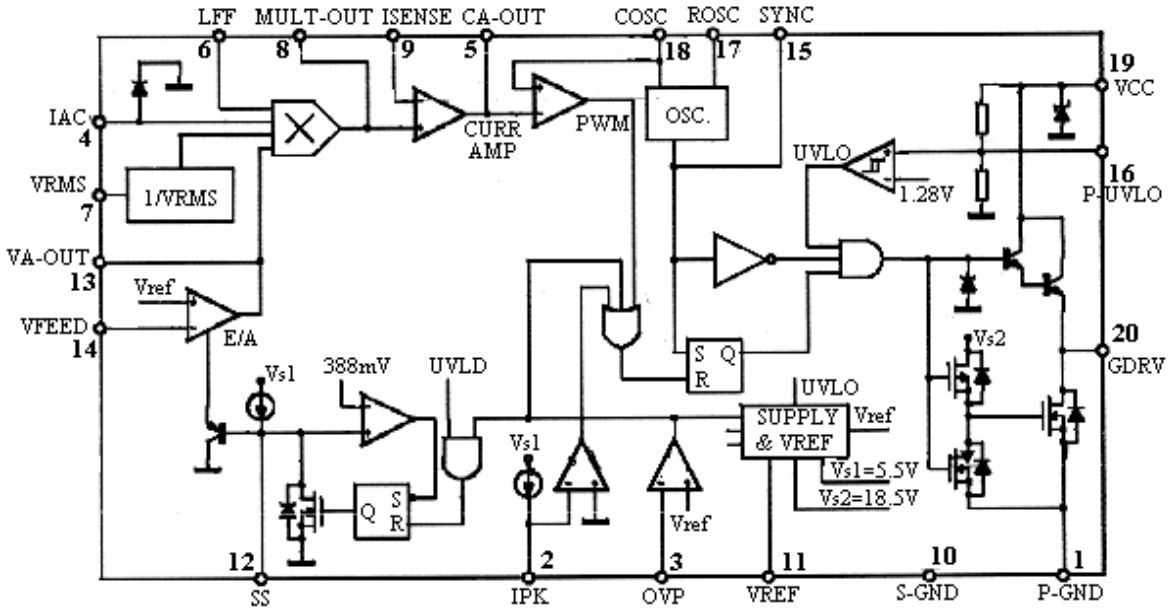

Fig. 7. Block diagram of L4981A

VI. Design Guidelines

The voltage conversion ratio is given by:

$$M = \frac{Vc}{Vm} \tag{5}$$

How higher is M value, lower is the δ variation, because:

$$\delta(t) = 1 - \frac{|sin(\omega \cdot t)|}{M} \tag{6}$$

So, chosen M, it is determined the minimum δ value:

$$\delta \min = 1 - \frac{1}{M} \tag{7}$$

As the duty-cycle varies along the line half-cycle, the rectangular voltage applied to the resonant filter changes its width. In a first approximation, it is possible to consider only the first harmonic component of the M2 voltage as producing current through the output.

The amplitude of this first harmonic component is given by:

$$a1 = \frac{2 \cdot Vm}{\pi \cdot (1 - \delta_{\min})} \cdot sin(\pi \cdot (1 - \delta_{mn}) \cdot |sin(\omega \cdot t)|) \tag{8}$$

Fig. 8 shows the variation of the first harmonic M2 voltage, for three values of δ_{min}.

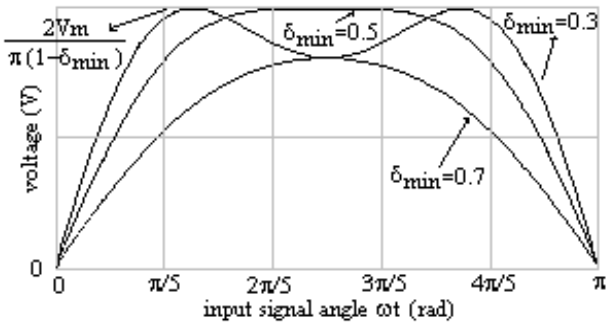

Fig. 8. Switch M2 voltage Fundamental Component variation.

This means that a low-frequency (120 Hz) lamp voltage modulation will be observed. For angles next the zero crossing the duty-cycle tends to 100% and the first harmonic of this signal goes to zero, reducing the lamp voltage and the current.

Nevertheless the IC4981A allows to limit the maximum duty-cycle. In the practical implementation this is done in order to maintain a sufficient high lamp voltage to not extinguish the arc.

The counterpart is that the input current will not be able to follow the input voltage waveform next the zero crossing. A compromise must be done, in which the input current distortion must be limited to the values given by the standard IEC 1000-3-2 [11].

As will be shown, the resulting lamp voltage variation will not produce the arc extinction, and does not deteriorate important quality indexes as the crest factor.

During the lamp starting process, the lamp filaments heating current flows through capacitor Cp.

This current is $if = \frac{Vf}{Rf}$, where Rf is the lamp filaments resistance and Vf is the lamp filaments voltage. Considering the lamp voltage is sinusoidal, with frequency Fs, we have:

$$Xcp = \frac{Vp}{if} \tag{9}$$

Where Vp is the lamp ignition voltage. Maintaining a narrow duty-cycle (using the soft-start pin of the L4981A), it is possible to implement a pre-heating stage.

Inductance Ls can be determined considering the resonant circuit impedance, in order to limit the current during the ignition process.

After the lamp starting, the resonance frequency decreases, being possible determine Cs value in order to adjust the desired lamp voltage.

The Li input inductor is dimensioned considering the desired input current ripple. Let us suppose a 100% efficiency and a sinusoidal input current, which peak value is $\mathrm{Im} = \frac{2 \cdot PL}{Vm}$, where PL is the nominal power lamp.

So, considering the maximum input current ripple, Δili%, that occurs at the input voltage peak, the input inductance is given by:

$$Li = \frac{Vm \cdot \delta \min \cdot Ts}{\Delta ili\% \cdot \mathrm{Im}} \tag{10}$$

VII. Experimental results

These results were obtained in a prototype with the following parameters:

Peak input voltage: Vm=180V,
Line frequency: F=60Hz,
Switching frequency: Fs=38kHz,
RMS lamp voltage: VL=127V,
Lamp power: PL=40W,
Filament resistance: Rf=4Ω,
Filament voltage: Vf=3V,
Ignition voltage: Vp=3VL,
Voltage conversion ratio: M=2,
Maximum current ripple: ΔiLi%=0,5.

The calculated values for the LCC filter, considering Cs=3Cp, are: Cp=7,5ηF, Cs=20ηF, Ls=2,7mH. The input inductance is Li=10mH.

Fig. 9 shows the input voltage, current and power. The distortion next the zero crossing is caused by the δ maximum value limitation, established by the control IC.

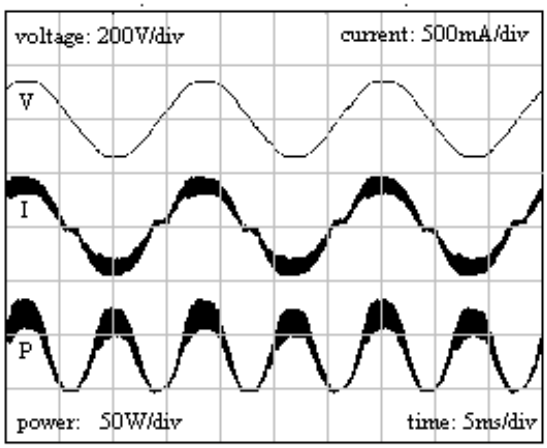

Fig. 9. Input voltage, current and power waveforms.

Fig. 10 shows the harmonic contents of the input current, which values are well below the limits of the standards.

Fig. 11 shows the lamp voltage, current and power. It can observed that does not happen voltage and current extinction along the line period, practically eliminating the stroboscopic effect. The amplitude modulation is consequence of the pulse width variation.

The measured current lamp crest factor was 1.6.

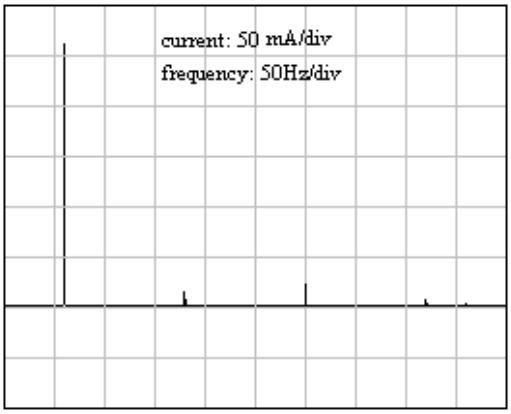

Fig. 10. Harmonic contents of input current.

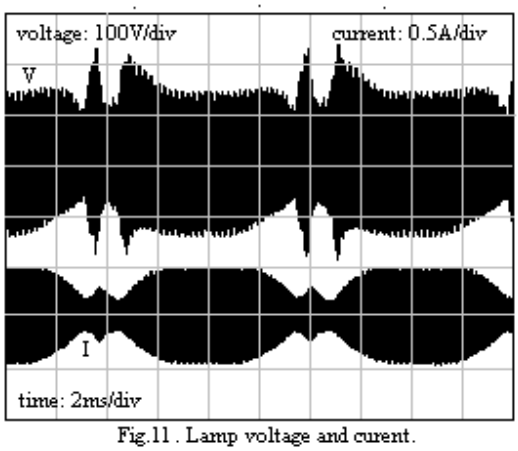

Fig.11 . Lamp voltage and curent.

Fig. 12 shows the high frequency lamp current and voltage waveforms, which present an almost sinusoidal shape.

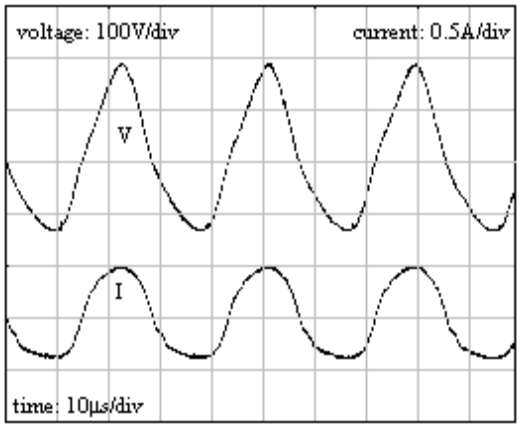

Fig. 12. High frequency lamp voltage and current.

Fig. 13 displays the voltage and current on switch M1, showing the ZVS commutations.

The measured power factor was 0.96 and the efficiency, 0.84.

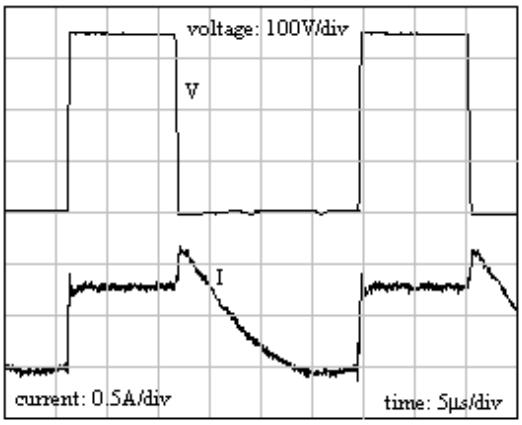

Fig. 13. M1 voltage and current.

VIII. Conclusions

The proposed electronic ballast presents several advantages especially due to its operation in the continuous conduction mode. The naturally filtered input current significantly reduces the necessary EMI filter. Soft-commutations are obtained without additional circuitry. The relatively complex control circuit is fully implemented using simple commercial IC used for PFC.

Even if the soft-switching properties are lost for wide duty-cycles, this fact does not compromise significantly the overall efficiency. To protect the switches it is convenient to use small passive snubbers.

IX. Acknowledgement

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X. REFERENCES

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