Low-frequency square-wave electronic ballast with resonant ignition using digital mode and power control

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Abstract— The paper proposes the use of a single full-bridge (FB) inverter and LC filter in HID electronic ballast, realizing both the functions of a resonant circuit for lamp ignition and a current controlled low frequency square wave inverter for normal operation. Ignition is achieved by resonance when the converter is controlled as a FB inverter at the resonant frequency of the LC filter. After ignition the converter operates as a LF square wave inverter by controlling the FB to act alternately as a Buck converter supplying positive or negative current. While ignition occurs at the LC filter resonance, the Buck converter switching frequency is selected significantly higher to attenuate high frequency harmonics and avoid exciting acoustic resonances. System stability is achieved by controlling the input current to the FB (output current of the PFC front-end), resulting in a power mode control provided that the voltage supplied by the PFC stage is constant. A digital control circuit selects the different operation modes of the FB converter according to the lamp requirements.

I. INTRODUCTION

The primary motivation for using low frequency square wave drive in electronic ballasts is to avoid excitation of acoustic resonances. The frequencies at which acoustic resonance appears depend on the size of the arc tube, gas pressure and its composition. The elements that compose the gas enclosed in the vessel determine the lamp chromatic rendering; in this way, a more complete spectrum of the gas enclosed in the vessel determine the lamp chromatic pressure and its composition. The elements that compose the resonance appears depend on the size of the arc tube, gas that then presents more resonant modes. Square-wave electronic ballast is an alternative to resonant converters and, theoretically, is a definite solution to prevent the acoustic resonance provided that the lamp power has no ac component. Standard solutions for square-wave high-intensity discharge (HID) lamp drivers over 100 W require three stages: (1) a power factor correction (PFC) stage, (2) a current mode controlled dc-dc converter and (3) a full bridge (FB) inverter, in addition to an external lamp ignition circuit [1]. In [2] a two-stage solution with a specific igniter circuit is proposed where the lamp current control provides the required system stability. Resonant ignition has been proved to be efficient for fluorescent lamps and high pressure sodium (HPS) lamps [3]. Recently, [4] also proposes the resonant ignition for metal halide lamps (MHL) to reduce the electrode sputtering and the consequent tungsten deposition. In this case, the resonant igniter is an external circuit that is coupled to the main power stage by a transformer.

In this paper we propose the design of a square-wave ballast in which stages (2), (3) and the ignition circuit are integrated in a single stage, as is depicted in Fig. 1. The signals that control the power switches are generated by a digital circuit according to the schematic presented in Fig. 2. System stability in short term and long term power control, including dimming capability, are achieved by controlling the dc current to be constant for a given power level.

II. DESCRIPTION OF THE LF SQUARE-WAVE BALLAST

A front-end PFC stage supplies constant dc voltage to the bridge converter. A digital control circuit imposes different operation modes to the bridge converter to operate initially as a resonant igniter, followed by a positive and negative Buck converter. The bridge inverter, the inductor and capacitor form a parallel resonant inverter before the lamp ignition, where the control circuit defines a soft start-up ignition sequence by performing either a frequency or phase
controlled sweep towards the parallel resonance. A phase-controlled soft start-up sequence provides the advantage of self-tuning to resonance and has no dependence on the absolute resonant frequency [5].

During the ignition sequence, resonant inverter operation is obtained since the transition from positive to negative and vice versa in the converter output voltage is carried out at high frequency. Let us call this frequency the inverter switching frequency, \( f_{sw} \), that, in this mode, is higher than the resonant frequency, \( f_o \), of the LC filter (see Fig. 1) when the lamp is off and it gradually approaches \( f_o \), where the voltage gain is high enough to produce the discharge. One benefit of sweeping the frequency for ignition is the resulting capacitive coupling that induces current through the lamp gas and reduces the required over voltage to achieve ignition [6]. The lamp warm up is also carried out using the high frequency inverter mode. During ignition and warm-up the duty cycle control is disabled and the converter operates as a traditional full-bridge resonant inverter.

After lamp ignition, the control circuit defines the bridge converter alternately as a positive and negative Buck converter. The square wave oscillator imposes 50% of the period \( T_i = 1/f_{sw} \) for each Buck converter mode, positive and negative (see Figs. 3 and 4). Control signals of the FB transistors for the Buck operation modes are defined in Fig. 3. In the low frequency square-wave (LFSW) mode, \( D \) is regulated to control the lamp power.

The ignition occurs close to the unloaded resonant frequency of the LC circuit, \( f_o = 1/2\pi\sqrt{LC} \). After ignition, the Buck converter operation generates an ac component whose amplitude in steady state at the fundamental (switching) frequency is

\[
\hat{V}_{ac, fsw} = \frac{2V_s}{\pi} \sin(D\pi) \cdot \left(\mathbf{1}\right)
\]

The low pass filter gain at the switching frequency, \( \left| \mathbf{1}\hat{f}_{ac, fsw} \right| \), reduces the ac component of the lamp voltage, which generates ac power amplitude supplied to the lamp at twice the switching frequency. Using the fundamental approximation

\[
\hat{P}_{lamp} = \frac{\left| D\hat{V}_s + \hat{V}_{ac, fsw} \sin(\omega_{out}) \mathbf{1}\hat{f}_{ac, fsw} \right|^2}{R_{lamp}} \cdot \left(\mathbf{2}\right)
\]

Therefore, the resulting amplitude of the ac component of the lamp power at the switching frequency is
\[
\hat{P}_{ac, fev} = \frac{2DV_s\hat{I}_{ac, fev}}{R_{lamp}} \mathcal{G}(j\omega_s).
\] (3)

To prevent the excitation of the resonant modes, let us impose a limitation to \(\hat{P}_{ac, fev}\), considering the studies on acoustic resonance in [7-9]. If \(\hat{P}_{ac, fev} < kP_{lamp}\), then the requirement for the low pass filter is derived from

\[
\frac{2DV_s\hat{I}_{ac, fev}}{R_{lamp}} \mathcal{G}(j\omega_s) < k \left( \frac{DV_s}{2} \right) \mathcal{G}(j\omega_s) = k \left( \frac{DV_s}{R_{lamp}} \right) \mathcal{G}(j\omega_s).
\] (4)

and, using (1), results

\[
\mathcal{G}(j\omega_s) < \frac{k\omega_s}{4\sin(D\pi)} \] (5)

Considering the practical values of a 150W-HID lamp, in which the \(R_{lamp}\) is around 50 \(\Omega\) for a new lamp, then if the output voltage of the front-end PFC is 400 V, which is a common value for a boost based PFC, the initial \(D = 0.22\). If \(k = 5\%\), then \(\mathcal{G}(j\omega_s) \approx -37.4\) dB. The value of \(R_{lamp}\) increases up to 150 \(\Omega\) during its lamp lifetime, so that the final duty cycle is \(D = 0.38\), which requires \(\mathcal{G}(j\omega_s) \approx -35.9\) dB, where \(\mathcal{G}(j\omega_s)\) is given by

\[
\mathcal{G}(j\omega_s) = \frac{1}{\sqrt{1 - \left( \frac{f_s}{f_o} \right)^2 + \frac{1}{Q^2(\frac{f_s}{f_o})}}}.
\] (6)

To increase the filter action, \(f_s\) is located above \(f_o\), e.g. \(f_s = 10f_o\) where \(\mathcal{G}(j\omega_s) \approx -40\) dB for large \(Q\) range.

Inductor and capacitor values are chosen to obtain the characteristic impedance \(Z = \sqrt{L/C}\) which limits the resonant current during the ignition sequence to a desired value and a quality factor \(Q\) that results in a fast and non-oscillatory transient from positive to negative Buck converter and vice versa. Practical limit of \(Q = R_{lamp}/(L\omega_o) = 1.2\) is selected in order to limit the low frequency transient oscillations.

As an illustration let us consider a design for a typical \(R_{lamp}\) variation from 50 \(\Omega\) to 150 \(\Omega\). A practical upper limit for the switching frequency is \(f_s = 200\) kHz based on switching losses, which results in \(f_o = 20\) kHz. Worst case \(Q\) occurs at the end of lamp life (150 \(\Omega\)), resulting in \(L = 995\) \(\mu\)H and \(C = 63.6\) nF for \(Q_{max} = 1.2\). Simulation of this case is represented in Fig. 6.

![Fig. 6. Gain (dB) of the double pole low-pass filter around \(f_s\) for different \(R_{lamp}\) (50 \(\Omega\), 75 \(\Omega\), 100 \(\Omega\) and 150 \(\Omega\)).](image)

IV. POWER-MODE CONTROL AND LOW-FREQUENCY TRANSITIONS

Power mode control is achieved by regulating the FB input current, which results in constant lamp power, neglecting variations in the PFC stage output voltage and FB losses. Here, we derive the small-signal model for controller design and consider stability.

If it is assumed that the ripple of the PFC output voltage is negligible, then the averaged model equation is described by

\[
\frac{dV_s}{sL} = \frac{\langle i_s \rangle_{1/4}}{d},
\] (7)

where the average input current \(\langle i_s \rangle\) is the variable to control. The resulting small signal model is given by

\[
\hat{d}(2DV_s - V) - \hat{v}D = \hat{i}_s.
\] (8)

Eq. (8) can be used to analyze the effects of independent perturbations of the duty cycle and output voltage on the input current, or the duty cycle command to input current transfer function when neglecting variations at the output voltage. However, for this application the output capacitance is generally small enough that it and the load must be included in the loop gain analysis. A modified result for (7) including output capacitance and lamp incremental impedance, \(Z_{lamp}\), is given by

\[
\hat{d}V_s = \langle i_s \rangle \left[ sL + \frac{Z_{lamp}}{1 + sCZ_{lamp}} \right].
\] (9)

The desired small-signal control transfer function from duty cycle to input current can then be derived. Given the relationship between averaged input and inductor current

\[
\langle i_s \rangle = d\langle i \rangle,
\] (10)

the resulting small-signal model is given by
where $R_{\text{lamp}}$ is the equivalent lamp resistance and $Z_{\text{lamp}}$ the lamp incremental impedance. The gain and phase of the transfer function given in (11) is represented in Fig. 7. To obtain the plots of Fig. 7 we have considered $Z_{\text{lamp}}$ at the frequency of interest, e.g., cross-over frequency at 10 kHz, which is a positive resistance of around 30 $\Omega$ [10]. From this result, a suitable compensator can be designed.

Another consideration from the dynamic model is the closed-loop output impedance, particularly given the potential impact of the negative incremental impedance of the lamp. For a loop gain $T(s)$ the closed loop small signal equation results in

$$i_g = \frac{T(s)}{[1 + T(s)]H(s)} v_{\text{gref}} - \frac{D}{sL} \frac{1}{1 + T(s)} v_i$$

(12)

with $i_g H(s) = \hat{v}_g$, where $H(s)$ models the sensor that samples $i_g$ and any associated low-pass filter.

The objective is to solve (12) for the output impedance, defined as the output voltage over the inductor current perturbations (neglecting the output capacitance). Additional equations are needed, including the small-signal form of (10), given by

$$i_s = iD + \hat{i}d,$$

(13)

where $i$ is the inductor current; the closed loop duty-cycle perturbation

$$\hat{d} = \frac{1}{V_m} G_c(s)(v_{\text{gref}} - i_g H(s)),$$

(14)

where $1/V_m$ and $G_c(s)$ are the modulator and compensator gain respectively; and the loop gain, including the control transfer function of (8)

$$T(s) = \frac{G_c(s)H(s)}{V_m sL} \frac{2DV_g - V}{V_n}.$$  

(15)

Then, introducing (14) into (13) and using $i_g$ given in (12) we obtain

$$\frac{\dot{i}}{\dot{v}_{\text{gref}}} = -\frac{\hat{v}}{sL} + \frac{1 + T(s)}{sL} \frac{G(s)H(s)}{V_n}.$$  

(16)

If $T(s) >> 1$ and $IGV/V_m >> 1$, then the resulting system output impedance is

$$z_{\text{o}} = -\frac{1}{sL} \frac{1 + T(s)}{sL} \frac{G(s)H(s)}{V_n} R_{\text{lamp}},$$

(17)

which corresponds to the condition of the converter controlled as a power source. Eq. (17) neglects the output capacitance; therefore the total output impedance is $R_{\text{lamp}}$ in parallel with $C$, where $C$ is negligible over the middle frequency range of interest (where the lamp incremental impedance is negative). Since the absolute value of the lamp negative incremental impedance at middle frequencies is typically an order of magnitude less than $R_{\text{lamp}}$, the power mode control output impedance of (17) provides the ballast system with the necessary stability [11].

Controlled transition from positive to negative lamp current and vice versa may result in a slow transient that might produce flickering effect and low frequency harmonics. In [12] a PWM controller modulates the output current with a square wave whose rising time is several times that of the Buck converter switching period. Here, to achieve a fast transient, an open-loop transition is proposed.

As $L$ and $C$ form a low pass filter loaded with $R_{\text{lamp}}$, the step response depends on the quality factor. If the quality factor, $Q$, is between 0.4 and 1.2 (see Fig. 6), the desired output voltage step response from positive to negative and vice versa is obtained.

Simulated Fourier analysis of the lamp power is shown in Fig. 8 and 9.
V. PRACTICAL IMPLEMENTATION

A. Power stage

The power stage is implemented according to the schematic shown in Fig. 1. Switches S1 and S3 are realized using IGBTs and D1 and D3 using ultra fast diodes in order to achieve high efficiency at a switching frequency of 200 kHz. A dead-time of 1 µs is selected for the inverter to prevent cross-conduction between the upper and lower transistors. Note that the inverter switching frequency, $f_{sw}$, is quite low, except at the beginning of the ignition sequence, where soft turn-on transition (ZVS) occurs.

The number of turns of the inductor $L$ is calculated to prevent core saturation during the ignition sequence, where the flux density in the core is the maximum. On the other hand, the copper section is calculated according to the inductor current in steady-state.

B. Digital control circuit

The control circuit is implemented by means of a Complex Programmable Logic Device (CPLD), which allows easy modification of the design, clocked by a 10 MHz crystal oscillator. Four operation modes may be distinguished to provide the corresponding drive signals of the switches:

1. Soft start-up ignition sequence mode, in which a frequency sweep from 100 kHz to 20 kHz is performed for the FB resonant inverter. If needed, the sweep is repeated until the lamp ignition is achieved. It is also possible to perform a phase sweep toward resonance to remove dependence on resonant component tolerances [5].

2. Warm-up mode: when the lamp ignition is detected, the circuit provides a fixed switching frequency slightly above the resonant frequency during a specified period.

   After warm-up, the control circuit establishes an alternate operation of the FB inverter as a positive and negative Buck converter with a frequency of 200 Hz. A dead time is included to assure a safe operation. The generated drive signals switching are depicted in Fig. 3.

3. Positive Buck mode; the control circuit turns S1 on and S2 off. S3 is also turned off and the drive signal of S4 is provided at a switching frequency of 200 kHz with the corresponding duty cycle.

4. Negative Buck mode; analogously, the control circuit turns S3 on, S4 off and S1 off and the corresponding drive signal of S2 is generated.

The control circuit performs the following actions:

When the lamp is off, the FB switches $V_g$ at high frequency. The lamp voltage, $v$, is sinusoidal. The switching frequency, $f_{sw}$, is reduced to approach $f_o$ until $v$ reaches the ignition voltage. The lamp ignition generates a signal that keeps $f_{sw}$ equal to $f_o$ during the lamp warm-up. If the ignition fails or a significant decrease in the input current is detected, the ignition sequence is repeated. After the lamp warm-up mode, $f_{sw}$ is shifted to a low frequency and the lamp voltage becomes a square wave. Power mode control is then activated to regulate the average lamp power.

A continuous time equivalent for the digital control loop is given by

$$T_d(s) = G_c(e^{Ts})H(s)G_{ip}(s)e^{-ts},$$  

where $G_c$ is the discrete-time compensator with $z = e^{Ts}$, $T = 1/f_{sw}$ is the sampling period, and $e^{-ts}$ represents the total delay of the controller from the input current sampling instant to time of corresponding duty cycle actuation. The sample frequency is set equal to the Buck converter switching frequency, with sampling instants just prior to the end of the duty cycle. The gain of the input current A/D converter and DPWM are assumed to be unity.

Based on $G_{ip}(s)$ given in (11) and shown in Fig. 7, a simple integral compensator is sufficient for zero steady-state error and stable operation at suitable bandwidth. Assuming a total controller delay of one sample, $t_d = T$, and a desired loop gain crossover frequency of $f_c = 5$ kHz, a suitable discrete-time compensator is given by

$$G_c(z) = 0.023 \frac{z + 1}{z - 1}. \quad (19)$$

The corresponding digital control algorithm is given by
\[ d[n] = d[n-1] + 0.023 \cdot (v_e[n] + v_{e[n-1]}), \]  

where \( v_e[n] \) is the error signal in the power control loop. An additional benefit from the simple integral controller is that impact of coefficient rounding in hardware is minimal, facilitating low cost implementation. Additional hardware reduction can be achieved by removing the compensator zero at the expense of additional phase lag. The required resolution of the duty cycle depends on the desired tolerance in lamp power regulation. For an input current sampling zero-error bin resolution of 5% full current, an 8-bit DPWM is suitable to avoid limit cycle behavior. To reduce hardware requirements, this resolution can be achieved using a combination of hardware and dither bits.

A plot of the system loop gain based on (18) and the controller of (19) is shown in Fig. 11, assuming a sense resistor gain of \( H = 1 \). The resulting crossover frequency \( f_c = 5 \text{ kHz} \) with 55° of phase margin and 24 dB of gain margin.

Next, experimental and simulation results are shown as verification of the proposal. The digital control circuit is implemented in a CPLD Altera EPM7128S. A prototype that controls the target power has been built using IGBTs HGTP12N60A4D that include ultra fast diode to implement the switches S1-D1 and S3-D3. Each switch S2-D2 and S4-D4 are two IRFP340 in parallel. Gate IGBTs and MOSFETs drive signals are generated with two IR2110 drivers. The inductor (1.3 mH) of the LC filter uses an E42 core size, material N27 and the capacitor is a 1600V MKP, 47nF.

Experimental results include the resonant lamp ignition sequence (Fig. 12), square wave operation (Fig. 13). In Fig. 13, the measured maximum lamp power noise is 15dBW, i.e around 4% of output power.

![Fig. 12. Digitally controlled ignition sequence. Top: lamp voltage. Bottom: lamp current.](image)

![Fig. 13. Experimental waveforms steady-state operation. Upper: lamp voltage, middle: lamp power, lower: lamp current.](image)

VI. CONCLUSIONS

A low frequency square-wave ballast using a single FB circuit for both low frequency drive and resonant lamp ignition has been presented. In this way, the proposed converter is a universal solution for different lamps of the same wattage (HPS and MHL). The converter is digitally controlled to define three operation modes: as a resonant inverter close to the resonant frequency to achieve the lamp ignition and initiate the warm up, and as a positive and negative Buck converter. The Buck converter operation is input current controlled with a high output impedance, as required to stabilize the lamp operation in short term (ballast
action) and also to provide constant lamp power during the lamp life time (long term). Transitions from positive to negative Buck converter and vice versa are performed open-loop to obtain a fast response. The design of the LC filter assures a quality factor that limits the resonant current during the ignition and generates non-controlled transitions of the low frequency square-wave around the critically damped behavior.

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