Control of low-frequency square-wave electronic ballast with resonant ignition using a dsPIC30F2010

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Abstract—The paper proposes the use of a dsPIC30F2010 for implementing the ignition sequence, a double loop control and selecting the operation modes of a low frequency square wave (LFSW) electronic ballast for two-stage HID lamps. The first stage is the power factor correction (PFC) stage, and the second is a single full-bridge (FB) inverter with LC filter. The latter stage performs function of a resonant circuit for lamp ignition and a current controlled low frequency square wave inverter for normal operation. Ignition is achieved at the resonant frequency of the LC filter when the FB converter is working as a resonant inverter. After ignition the converter operates as a LFSW 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 \( f_R = 25\text{kHz} \), the Buck converter switching frequency \( f_{sw} = 200\text{kHz} \) is selected significantly higher to attenuate high frequency harmonics and avoid exciting acoustic resonance. Lamp stability is achieved by controlling the inductor current of the LC filter, and power mode control is achieved adjusting average current supplied by the PFC stage. The microcontroller 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. Particularly, Metal Halide Lamps (MHL) are sensitive to it. This kind of lamp has become very popular as a practical light source for general and specific applications for its high efficacy (measured in lumens per watt) and for the creation of compact lamps with superior color rendering properties. 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 source light requires a more complex composition of the lamp gas that then presents more resonant modes. Square-wave electronic ballast is an alternative to resonant converters and, theoretically, is a 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 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 [5] the efficient start-up of MHL with a half bridge resonant inverter was proved.

In this paper the digital control of a simple square-wave electronic ballast in which, stages (2), (3) and the ignition circuit integrated in a single stage is proposed. The power converter is depicted in Fig. 1. The use of a microcontroller with a suitable digital control allows the FB to work as a resonant inverter during the start-up and warm-up of the lamp and as a LFSW inverter during the steady state of the lamp.

The signals that control the power switches are generated by a dsPIC30F2010 microcontroller according to the schematic presented in Fig. 2. System stability in short term and long term power control, including dimming capability, are achieved with a double control: an inner control that provides stability to the lamp, controlling the average inductor current, \( i \), and adjusts the duty cycle. The second, outer control is a power mode control that regulates the average current, \( \langle i \rangle \) supplied by PFC stage. Both control loops use the same sensor resistor \( R_g \), see Fig. 1.

![Fig. 1. Square-wave electronic ballast. Stage 1 is a Power Factor Corrector stage, and stage 2 is a full bridge that works as a resonant inverter or low frequency square wave inverter according to the state of the lamp.](image-url)
output current $i_g$ during $DT$, sensed by $R_g$ (see Fig. 1). On the other hand, power mode control is achieved by using the PFC average output current, $<i_g>$. Power mode control regulates $P_{lamp}$ and provides the reference inductor current $i_{ref}$ to achieve the correct control parameter, $D$, in the current mode control.

Since the Buck converter input current, $i_{in}$, is discontinuous, the sampled inductor current is multiplied by $D$ to obtain the average value. A two-loop control strategy is chosen. The inner high-speed current loop provides the ballast action since it stabilizes the lamp current resulting in high output impedance for the inverter. The outer power loop provides the current reference to the inner loop to achieve the desired lamp power. The power mode control does not necessitate high-speed performance, therefore the power sample data can be obtained by using the same current sample as in the current loop, and multiplying it by the duty cycle and the PFC front-end output current.

To simplify the current loop design with no influence on the measured current conditioner, the generation of open-loop transitions from positive to negative current through the lamp and vice-versa is proposed. If the duty cycle does not change during the transition from positive to negative output voltage and vice versa, the transition response depends on the quality factor of the filter loaded with the lamp. The control circuit should detect the end of the transition to retake control of the current and power mode.

### III. LC FILTER DESIGN

The full bridge starts working as a resonant inverter during the soft start-up sequence and the microcontroller generates the
transistor drive signals of Fig. 5. The lamp ignition occurs close to the unloaded resonant frequency of the LC circuit

\[ f_o = \frac{1}{2\pi}\sqrt{LC} \]  

(1)

Fig. 5. Full bridge transistor drive signals during start-up sequence and warm-up of the lamp.

After ignition and warm-up, the Buck converter operation generates an ac component whose amplitude in steady state at the fundamental (switching) frequency is

\[ \hat{V}_{ac,sw} = \frac{2V}{\pi}\sin(D\pi) \]  

(2)

The low pass filter gain at the switching frequency, (4), 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

\[ P_{lamp} = \frac{[DV_g + \hat{V}_{ac,sw}\sin(\omega_{sw}t)]\|G(jf_{sw})\|^2}{R_{lamp}} \]  

(3)

Where the magnitude of the Buck filter response to the switching is given by

\[ \|G(jf_{sw})\| = \frac{1}{\sqrt{1 - \left(\frac{f}{f_{sw}}\right)^2} + \frac{1}{Q^2}\left(\frac{f}{f_{sw}}\right)^2}} \]  

(4)

Therefore, the resulting amplitude of the ac component of the lamp power at the switching frequency is

\[ \hat{P}_{ac,sw} = \frac{2DV_g\hat{V}_{ac,sw}\|G(jf_{sw})\|}{R_{lamp}} \]  

(5)

To prevent the excitation of the resonant modes, let us impose the limitation

\[ \hat{P}_{ac,sw} < kP_{lamp} \]  

(6)

This limitation is based on the acoustic resonance studies in [7], where a typical value for the parameter \( k \) to prevent the excitation of acoustic resonance under any condition is \( k < 5\% \). The requirement for the low pass filter is then derived from

\[ \frac{2DV_g\hat{V}_{ac,sw}\|G(jf_{sw})\|^2}{R_{lamp}} < k \left(\frac{DV_g}{2}\right)^2 \approx k \left(\frac{DV_g}{2}\right)^2 \]  

(7)

And, using (2), results in

\[ \|G(jf_{sw})\| < \frac{k\pi D}{4\sin(D\pi)} \]  

(8)

This is the condition to prevent the excitation of acoustic resonance. In our practical case of 150W-HID lamps, in which the \( R_{lamp} \) is around 50 Ω for a new lamp, the output voltage of the front-end PFC is \( V_g = 400 \) V, the initial \( D = 0.22 \). The value of \( R_{lamp} \) increases up to 150 Ω during its lamp lifetime, so that the final duty cycle is \( D = 0.38 \).

A good solution in (8) to prevent the excitation of acoustic resonance during the lamp lifetime [8] is (9), and results in \( f_{sw} = 10f_o \).

\[ \|G(jf_{sw})\| \equiv -40dB \]  

(9)

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. The practical limit of \( Q = R_{lamp}/(L\omega_o) \) is 1.2 is selected in order to limit the low-frequency transient oscillations.

A practical upper limit for the switching frequency is \( f_{sw} = 200 \) kHz based on switching losses, which results in \( f_o = 20 \) kHz. The worst case \( Q \) occurs at the end of lamp life (150 Ω), resulting in \( L = 995 \) μH and \( C = 63.6 \) nF for \( Q_{max} = 1.2 \).

IV. CURRENT MODE AND POWER MODE CONTROL

The implementation of a suitable regulation of the lamp, should result in a double actuation. On the one hand, the ballast needs to stabilize the lamp in the short term, and on the other hand, it needs to have the capacity to fix the power of the lamp in medium and long term.

Here it is proposed use a fast current mode control that stabilizes the lamp and an outer power mode control that fixes the normal lamp power and implements dimming control.
Fig. 6. represents a typical buck converter scheme where \( i_g \) is the input current, and \( i \) is the inductor current that is the variable to be controlled by the inner loop.

Since, \( i = i_g \) during \( DT \) (on time), the inductor current can be captured at \( R_g \) (see Fig. 1) for inner loop. The outer power mode control is implemented taking \( V_g \) constant. Therefore, the power control variable is the input current, \( <i_g> \), that is calculated by multiplying the sampled inductor current, \( i \), by the duty cycle, \( D \).

The small-signal model is derived for controller design and to consider stability [8]. If it is assumed that the ripple of the PFC output voltage is negligible. The average inductor current \( <i> \) is the variable to be controlled. The resulting transfer function is given by

\[
G_{id}(s) = \frac{I(s)}{D(s)} = \frac{V_g}{Z_{lamp} + \frac{1 + sZ_{lamp}C}{L + s^2LC}}
\]  

(10)

where \( Z_{lamp} \) is the lamp incremental impedance. The gain and phase of the transfer function given in (10) is represented by a continuous line in Fig. 7, and a discontinuous line represents the discrete transfer function of (10). To obtain the plots in Fig. 7 of \( Z_{lamp} \) at the frequency of interest, e.g. cross-over frequency at 10 kHz, a positive resistance of around 30 \( \Omega \) [9-10] has been considered. From this result, a suitable compensator can be designed.

Power mode control is easily implemented as

\[
P_{lamp} = \eta_{fsw} \cdot V_g \cdot <i_g>
\]

(11)

\[
<i_g> = d \cdot \langle i \rangle
\]

(12)

where \( \eta_{fsw} \) is the efficiency of the LFSW converter.

\[\text{Fig. 7. Estimated Bode plot of } G_{id}(s) \text{ in a continuous line, and } G_{id}(z) \text{ in a discontinuous line. Above: gain in dB. Below: phase.}\]

In Fig. 8, the transistor drive signal, \( V_{gs} \), the input current, \( i_g \), and a sample of this, \( V_{sen} \), are shown. The buck converter is working in continuous conduction mode and \( \Delta i \) is small, so that the sample of input current, \( V_{sen} \), is similar to the graph in Fig. 8. In this way, the sample of input current is directly, the sample of inductor current which is the variable to control.

\[\text{Fig. 8. Top signal shows the sample of input current, } V_{sen}=i_g \cdot R_g \text{, that is the sample of the inductor current during DT. Middle signal shows } i_g \text{, and bottom shows the transistor driven signal.}\]

V. PRACTICAL IMPLEMENTATION

The converter operation is summarized in the flowchart of the dsPIC30F2010 program in Fig. 9. Four operation modes may be distinguished to provide the corresponding drive signals for the switches:
A. Soft start-up ignition sequence mode

The 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.

B. 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.

C. LFSW: Positive and negative Buck mode.

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 safe operation. Positive Buck mode is achieved when 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. Negative Buck mode is achieved, analogously, when the control circuit turns S3 on, S4 off and S1 off and the corresponding drive signal of S2 is generated. In this mode, an inductor current control and power mode control is activated.

Continuous time equivalent for the digital inductor current control loop is given by

\[ T_j(s) = G_c(e^{sT})H(s)G_{op}(s)e^{-\alpha_s} \]

(13)

where \( G_c \) is the discrete-time compensator with \( z = e^{sT} \), \( T = 1/f_{\text{sw}} \), is the sampling period, and \( e^{-\alpha_s} \) represents the total delay of the controller from the input current sampling instant to the time of the corresponding duty cycle actuation. The sample constant is \( H(s) = 0.25 \). 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, and zero order hold (ZOH) is selected. Based on \( G_{op}(s) \) given in (10), and its discrete transfer function, \( G_{op}(z) \), shown in Fig. 7., inductor current loop requires a simple proportional compensator to stabilize the electrical discharge, at suitable bandwidth. Assuming a total controller delay of one sample, \( t_d = T \), and a desired loop gain crossover frequency of \( f_c = 5 \text{ kHz} \), a suitable discrete-time compensator is given by \( G_c(z) = 0.5 \). On the other hand, power loop requires a simple integrator, \( G_p(z) = 0.2/(z-1) \), to achieve zero error lamp power control. Fig. 10. shows the discrete system, and Fig. 11. shows the stability loop using a SISO design tool of MATLAB®.

Fig. 9. Control program flowchart.

VI. RESULTS

Next, experimental and simulation results are shown as verification of the proposal. The digital control circuit is implemented in a dsPIC30F201. The FB, shown in Fig. 1., that controls the target power has been built using IGBT’s HGTP12N60A4D that includes an ultra fast diode to implement the switches S1-D1 and S3-D3. The switches S2 and S4 are MOSFETs IRFP340. Gate 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. 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 maximal. On the other hand, the copper section is calculated according to the inductor current in steady-state.

Experimental results include the resonant lamp ignition sequence (Fig. 12.) and square wave operation (Fig. 13.). Fig. 14. shows the situation when the dsPIC takes the sample of input current. The time to obtain correct data is controlled, and the ADC only needs 154 ns for a good sample, Fig. 14. The possibility of controlling the time of the sampling allows noisy zones to be avoided, i.e. during the transistor switching.
Fig. 12. Start-up for 150W HPS lamp. Above: lamp voltage. Below: lamp current.

Fig. 13. Experimental waveforms steady-state operation. Upper: lamp voltage, middle: lamp power, lower: lamp current.

Fig. 14. Experimental waveforms steady-state operation. Upper: sample of \( i_L \). middle: Sample and converter time, lower: transistor driven signal.

VII. CONCLUSIONS

A new digital control for a low-frequency square-wave ballast using a single FB circuit for both low-frequency drive and resonant lamp ignition has been presented. One of the benefits of this electronic ballast is the reduction of components due to use of digital control and fewer power stages than similar ballast. 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 high output impedance, as is required to stabilize the lamp operation in the short term (ballast action) and also to provide constant lamp power during the lamp lifetime (long term). Transitions from positive to negative Buck converter and vice versa are performed in 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.

ACKNOWLEDGMENT

This work is co-sponsored by the Spanish Government through the project CICYT TEC 2004-02607/MIC "Systems for discharge lamps and electrical discharge machining", and the National Science Foundation (under Grant No. 0348772).

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