A Self-Oscillating Drive for Single-Stage Charge-Pump Power-Factor-Correction Electronic Ballast with Frequency Modulation

Fengfeng Tao, Qun Zhao, and Fred C. Lee
Center for Power Electronics Systems
Virginia Polytechnic Institute and State University
Blacksburg, VA 24061-0111

Naoki Onishi
Matsushita Electric Works
R&D Laboratory, Inc
216 West Cummings Park
Woburn, MA 01801

Abstract- Single-stage current-source charge-pump power-factor-correction (SS CS-CPPFC) electronic ballast employs charge-pump concept to obtain high power factor with low cost. However, the circuit has a poor lamp current crest factor and large lamp output power variation, which deteriorates the lamp life. Feed-forward and/or feedback circuit are usually used to deal with these problems at the penalty of adding circuit complexity and cost. This paper introduces a novel self-oscillating drive circuit with a switch frequency modulation that avoids complicated control circuit. Self-oscillating principle and general structure are discussed. Current drive source selection and current injection concepts are originally developed to achieve constant lamp power operation and low-crest factor. It is shown that with this simple, low cost, passive drive system, low total harmonic distortion, low current stress, low-crest factor, and constant lamp power operation are achieved.

I. INTRODUCTION

High-frequency electronic ballasts for gas discharge lamps are used widely in lighting systems today because of their merits of small size, light weight, high light luminous efficacy, no flicker, no audible noise, and long life [1]. Currently, the electronic ballast has to meet the regulation of IEC 61000-3-2 class C to reduce the line rms current and line-current harmonic distortion. The power line then can be utilized more efficiently. Usually, a two-stage approach, i.e., a PFC stage followed by a DC/AC inverter stage, is used. The two-stage approach has good performances such as a near-unit power factor and wide range of line input voltage variation. Besides, the design procedure is relatively easy. The main problem of the two-stage approach is that it has more components and, thus, a high cost. This is not good for cost-sensitive products. Several single-stage PFC electronic ballasts, aimed at reducing the cost, have been proposed previously [2-8]. In the single-stage approach, the PFC stage is combined with the DC/AC inverter stage into one stage; thus one switch and its controller can be saved, and consequently the cost is reduced. The most commonly used method is to combine a boost converter with the DC/AC inverter by sharing a common switch. A high power factor can be achieved by deliberately operating the boost inductor in the discontinuous conduction mode (DCM) with constant duty-cycle control. However, the shared switch usually suffers from high current stress because it needs to conduct the reflected load current and line current. A high-current rating device has to be used. Recently, a family of single-stage CP-PFC electronic ballasts have been proposed and become attractive topologies [4-8]. They employ a capacitor instead of an inductor to achieve PFC, and a capacitor is usually cheaper and more reliable than an inductor. Among them, the current-source charge-pump PFC (CS CPPFC) electronic ballast presented in [5] has the attractive merits of good performance, simple structure, and small current and voltage stresses. The switches only deal with the resonant current. However, due to the modulation effect of the charging capacitor, the lamp current has twice of the line frequency ripple, which deteriorates the crest factor. To improve the crest factor, an average lamp-current control with switching frequency modulation was developed in [5]. However, it needs a lamp-current sensor, PI compensator, and oscillator for negative feedback control. Usually, the series-resonant parallel-load inverter with the self-oscillating drive is one of the most simple, efficient, and cost-effective circuits for ballasting lamps [10,12]. The switches are controlled by the power taken from the oscillating power stage. The circuit has a simple structure with a very small number of components. However, the switching frequency is determined by the circuit itself, and hence, the circuit analysis and design are not simple.

In this paper, current-source (CS) CPPFC electronic ballast with self-oscillating drive is presented and analyzed. The circuit employs MOSFETs as power switches and a linear core for a current transformer. The principle of the self-oscillating drive for MOSFETs is analyzed. A general structure for the self-oscillating electronic ballast is then proposed. Based on the general structure, current drive source selection and current injection concept for the self-oscillation drive are originally proposed to improve ballast performance in terms of lamp current crest factor and lamp power variation. Based on the steady-state analysis, a design guideline is proposed. Finally a self-oscillating drive is implemented to verify the theoretical analysis.
II. Circuit Description

Fig. 1 shows the CS CPPFC with self-oscillating drive electronic ballast. Diodes $D_6$–$D_9$ compose of a regular full-wave rectifier, $L_q$ and $C_f$ consists of input filter. The fast diode $D_1$ ensures unidirectional input current and $D_4$ enables the boost-up function of the circuit. The series-resonant parallel-load tank is composed of $L_r$, $C_r$, and the lamp. The capacitor $C_d$ functions as a DC blocking device to the resonant inductor. The capacitor $C_m$ functions as the charge-pump capacitor, together with the resonant inductor current, to obtain a high power factor. The self-oscillating drive circuit is composed of a current-transformer (CT) $T_r$ with three windings ($N_p$, $N_s1$, and $N_s2$), gate drive resistors $R_{g1}$ and $R_{g2}$, and two sets of back-to-back zener diodes ($D_3$ to $D_4$). The resonant inductor current (drive source current) is fed back through the CT and converted into a complementary voltage to drive the two MOSFETs $M_1$ and $M_2$. Because this configuration cannot start-up naturally, a starter circuit, which is composed of the resistor $R_{st}$, capacitor $C_{st}$, diode $D_{st}$, and diac $D_{ac}$, is needed.

III. Principle of Self-Oscillating Operation

To analyze the principle of the self-oscillating drive for MOSFETs, a half-bridge inverter without charge-pump stage, shown in Fig. 2a, is simulated. In approximately one second after power is applied, the voltage across $C_u$ will reach about 30V break-over voltage of the diac $D_{ac}$. With the $D_{ac}$ conducting, a positive turn-on voltage pulse is applied to the gate of $M_2$. When $M_2$ is turned on by this pulse, the drain voltage of $M_2$ is rapidly switched to ground, and thus initiating the resonant circuit oscillation. As long as $M_2$ is turned on each half cycle, electrical charge developed across $C_u$ is discharged through $D_{ac}$ preventing build-up voltage of further start-up pulses. The CT, $T_r$, is in series with the resonant inductor. The polarities of the CT are chosen in such a way that resonant inductor current, flowing through the primary side of the CT, will generate the complementary gate drive voltages for the power MOSFETs $M_1$ and $M_2$, causing the circuit to oscillate. An inductive appearance of the resonant tank is naturally obtained by this configuration so that zero-voltage switching (ZVS) is achieved for both MOSFETs. During this start-up mode, a high striking voltage across the lamp tube is generated due to the lightly damped resonant tank, which is composed of $L_r$, $C_r$, and the lamp filaments. With the lamp turned on, the lamp impedance drops quickly, and after several switching cycles, the circuit reaches a steady state. Fig. 2b shows the equivalent circuit of Fig. 2a, assuming the gate resistor $R_g$ is zero. Basically, the gate of the MOSFET is a capacitance network. $C_{gd}$ and $C_{gs}$ represent the gate capacitance, and miller feedback drain-gate capacitance, which is usually voltage dependent. The current transformer is modeled as a current-control current-source shunted by a magnetizing inductance, $L_m$, with two secondary windings coupled by an ideal transformer, as shown in Fig. 2b, where $N=N_s2/N_p$. Fig. 3 shows the simulated key switching waveforms. Eight topological stages exist over a complete self-oscillating switching cycle. Because of symmetrical gate drive circuits for $M_1$ and $M_2$, four topological stages over half switching cycle are discussed as follows.

Stage1 [$t_0$, $t_1$]: Assume before $t_0$, the secondary winding current $I_{ns2}$ charges the $M_2$ gate capacitor and cause the gate voltage to increase. At $t_0$, the gate voltage reaches the zener break-over voltage and is clamped. $M_2$ is fully on and $M_1$ is cut-off. The equivalent circuit is shown in Fig. 4a. The drive current $i_s$ is bigger than magnetizing current $i_{m0}$. The magnetizing current linearly increases due to the positive volt-second while drive source current increases in a resonant way until the magnetizing current reaches the drive current at time $t_1$ where secondary winding current $I_{ns2}$ equals zero.

$$i_{Lm}(t) = \frac{1}{L_m} \int_{t_s}^{t} (V_s + V_d) dt + I_{m0}$$

$$= \frac{V_s + V_d}{L_m} (t - t_s) + I_{m0}$$

where $V_s$ and $V_d$ are zener clamping voltage and diode forward voltage respectively, and $I_{m0}$ is the initial magnetizing current at $t_0$.

Stage2 [$t_1$, $t_2$]: At $t_1$, the magnetizing current reaches the drive current $i_s$ and the secondary winding current $I_{ns2}$
becomes negative. The zener diode is turned off and the gate capacitors show up, causing resonance with the magnetizing inductance. M₂ is still fully on and M₁ is still turned off. However, the energy stored in the gate capacitors begins to transfer to the magnetizing inductor, and the gate voltage of M₂ begins to drop while the gate voltage of M₁ increases. The switch M₂ is carrying the resonant current until the gate voltage falls to \( (V_{th} + i_{Lr}/g_{fs}) \) at \( t_2 \), which corresponds to the gate voltage needed to sustain the drain current \( i_{Lr} \). 

\( V_{th} \) and \( g_{fs} \) 

Fig. 3. Simulated switching waveforms.

Fig. 4. Topological stages over half-switching cycle, (a) stage 1 \([t_0, t_1]\), (b) stage 2 \([t_1, t_2]\), (c) stage 3 \([t_2, t_3]\), and (d) stage 4 \([t_3, t_4]\).
are the gate threshold voltage and the transconductance of the MOSFET, respectively.

Stage 3 \([t_2, t_3]\): The drain of \(M_2\) still carries most of the resonant current, and part of the resonant current goes through the Miller capacitance, \(C_{gs1}\), causing the drain voltage of \(M_1\) to decrease. Since the drain current and gate voltage are tied inextricably to one another by the MOSFET’s transfer characteristic, the gate voltage of \(M_2\) keeps constant. Fig. 4c shows the equivalent circuit. This stage won’t end until the drain voltage of \(M_2\) reaches the bus voltage \(V_B\), where the body diode of \(M_1\) begins to turn on at \(t_3\), realizing ZVS.

Stage 4 \([t_3, t_4]\): At \(t_3\), the gate voltage of \(M_2\) begins to drop and the drain-current \(i_{D2}\) decreases according to the MOSFET’s transfer characteristic curve. The difference between the resonant-current \(i_{Lr}\) and \(i_{D2}\) is carried by the body diode of \(M_1\). This commutation mode does not end until gate voltage of \(M_2\) drops below the threshold voltage \(V_{th}\) and \(M_2\) is cut off. The \(i_{Lm}\) reaches its maximum \((i_{Lm}/N)\) when the gate voltage of \(M_2\) drops to zero. This mode ends when the gate voltage of \(M_2\) is reverse clamped by the zener diode at time \(t_4\), then another half-switching cycle begins.

The switching period is load-dependent and determined by four time intervals. The first time interval corresponds to the stage 1 where magnetizing inductance is linearly charged by the clamped gate voltage \(V_{g1}+V_d\). This period is determined by magnetizing inductance seen from the secondary side, zener clamped voltage, and source current \(i_s\). The second time interval corresponds to the stage 2 where the gate capacitance begins resonance with the magnetizing inductance. This time period is determined by the gate stored charge and magnetizing inductance. Different MOSFET has different gate charge so that this time interval varies. The third time interval corresponds to the stage 3 where the MOSFET operates in saturation region. The gate-to-drain capacitance discharging current plays important role to determine this period. From the waveforms, it is clearly to see that the MOSFET drain-to-source voltage increases to bus voltage and at the same time the MOSFET carries full load current (resonant current). The main turn-off loss of MOSFET occurs in this period. To reduce turn-off loss, a bigger discharging current is preferred. The fourth time interval corresponds to the stage 4 where the gate capacitance becomes resonance with the magnetizing inductance.

IV. DRIVE SOURCE SELECTION AND CURRENT INJECTION TECHNIQUE

From the analysis above, the self-oscillating switching frequency is basically self-determined. However, it is not so difficult to obtain a desired switching frequency if we see the circuit from a system point of view. The whole system can be divided into three sub-systems as shown in Fig. 5a. The first sub-system is a power amplifier that provides a high-frequency square-wave voltage source \(V_{NM}\). The second sub-system is a LC series-resonant parallel-load tank fed by \(V_{NM}\). The third sub-system is a converter, which converts the feedback resonant current (drive source), \(i_{Lr}\), into suitable voltage signals for the power amplifier inputs. Fig. 5b shows the system block diagram. Fig. 6 shows a more general structure of the self-oscillating electronic ballast. The function of the converter, \(C\), is to remove the phase shift (lag or leading) between the square-wave voltage source and the drive source. The loop gain of the system is thus one, and the phase angle \(2\pi k\) to maintain the circuit oscillation.

So two key design considerations should be highlighted: one is drive source selection, the other is converter \(C\) design. The drive source provides drive power to MOSFET gate and converter \(C\) compensates the phase shift between the drive source and \(V_{NM}\). To improve circuit performance, the drive source should have some inherent negative feedback mechanism. Here we only consider about current drive source selection. Fig. 7 shows the simulated switching waveforms: resonant inductor current, \(i_{Lr}\), resonant capacitor current, \(i_{Cr}\), and lamp current, \(i_{Lamp}\), with respect to the voltage \(V_{NM}\). All of them have a phase lag so that a simple zener clamping circuit together with magnetizing inductance can provide the required phase compensation. Among these three drive
sources, only resonant capacitor current, $i_{Cr}$, has negative feedback mechanism due to the discharging lamp negative impedance characteristic.

In some applications, the switching frequency needs to modulate to achieve good performance. Switching frequency is determined by the turn-off time instant in ZVS operation. From the analysis above, the turn-off time instant in self-oscillating drive is determined by the polarity of the secondary currents $i_{Ns1}$ and $i_{Ns2}$, and the secondary currents equal the difference between the drive current and magnetizing current over turns ratio. If we inject some current into the current transformer, we can change the polarity of the secondary current, and thus modulate the switching frequency.

V. EXPERIMENTAL RESULTS

Based on the above analysis, the circuit shown in Fig. 1 was implemented to verify its operation. A transformer is used to match the lamp impedance as shown in Fig. 8. The ballast is designed under the following condition.

- AC input voltage $V_g$: 200 ± 10% Vrms
- Output lamp current: 0.23A
- Output lamp power: 64 W (Two 32W lamps in series)
- Power transformer turn ratio: 1:3.873
- Switching frequency: 47.2kHz
- Resonant capacitor Cr: 3nF

Based on the design considerations in [5], the charge capacitance and the resonant inductance are obtained as $C_{in} = 50.2 \text{ nF}$ and $L_r = 0.51 \text{ mH}$.

Like in conventional self-oscillating circuit, we first select resonant inductor current as drive source. Four 15-voltage zener diodes are composed of two sets back-to-back clamping circuit. TDK PC40 EI16 core was used for CT. The turns-ratio is 15.

Fig. 9a shows the measured line current waveform with a 0.976 power factor and 16.1% THD. Fig. 9b shows the lamp current envelope during line cycle. It is seen that the lamp current is highly modulated by twice the line frequency. The measured crest factor is as high as 1.88, much higher than 1.7 requirement. As discussed in [5], the crest factor is poor due to the modulation of the charge-pump capacitor. Another problem is the high sensitivity of lamp power to input line variation. The measured lamp power variation range is around 15% when input line has ±10% variation. The high crest factor and power variation may deteriorate lamp life. In order to improve the lamp crest factor, the switching frequency modulation method is employed. As discussed in previous section, a current injection scheme is developed to modulate the self-oscillating frequency. From the measured lamp-current envelope shown in Fig. 9b, the lamp has high current near the line zero crossing and small lamp current near line peak voltage. It is required that the switching frequency is high near line zero crossing and low near line peak voltage. The injected current should have such property.
that the magnitude is low near line zero crossing and high near line peak voltage, and also the phase should not conflict with the drive source. Fig. 10 shows the simulation waveform of the circuit in Fig. 1 using an external gate drive with a fixed switching frequency. It is seen that the current through the charge-pump capacitor $C_{in}$ meets the requirements. In order to minimize lamp power variation, resonant capacitor current is used as drive source. Fig. 11 shows the self-oscillating drive with switching frequency modulation for the single-stage current-source charge-pump power-factor-correction electronic ballast. Fig. 12 shows the switching-current waveforms near line zero crossing, half-line, and line peak voltage. It can be seen that the corresponding switching frequencies are 50, 44.48, and 37.3 kHz. The measured lamp-current waveforms are shown in Fig. 13 with ±10% line variation. The measured crest factors are 1.48, 1.49, and 1.46, respectively. Lamp power variation range is within 10%. Fig. 14 shows the line input current waveforms at different line input voltages. The measured each harmonic component meets the IEC 61000-3-2 Class C requirement. And overall efficiency is 84% including filament loss.

**VI. CONCLUSION**

In this paper, the principle of the self-oscillating drive for MOSFETs is analyzed from both of switching cycle and system point of view. A general structure for the self-oscillating electronic ballast is then proposed. Based on the analysis, a current injection scheme is proposed to achieve switching modulation. The theoretical analysis was verified by a 64-watt current-source CP-PFC electronic ballast.
REFERENCES


Fig.13. Measured lamp current waveforms at different line voltage, $i_{\text{Lamp}}$: 0.2A/div, time: 2ms/div.
(a) $v_g=200$ V, (b) $v_g=180$ V, and (c) $v_g=220$ V.

Fig. 14. Measured line input current waveforms at different line voltage, $v_g$: 100V/div, $i_g$: 0.5A/div, time: 2ms/div.
(a) $v_g=200$ V, (b) $v_g=180$ V, and (c) $v_g=220$ V.