A Single-Stage High-Power-Factor Electronic Ballast With ZVS Buck–Boost Conversion

Chin Sien Moo, Kuo Hsing Lee, Hung Liang Cheng, and Wei Ming Chen

Abstract—This paper proposes a novel single-stage high-power-factor electronic ballast via the integration of a derivative buck–boost converter and a half-bridge resonant inverter. The derivative buck–boost converter employs two coupled inductors with an appropriate turn ratio to conduct the current from the input line source into the designated power switches. With the tactful topology and delicately designed circuit parameters, both the active power switches of the resonant inverter can retain the zero voltage switching, resulting in a high circuit efficiency. A prototype circuit designed for a T8-36W rapid-start fluorescent lamp was built and tested to verify the analytical predictions, and satisfactory results were obtained experimentally.

Index Terms—Buck–boost converter, electronic ballast, fluorescent lamp, resonant inverter.

I. INTRODUCTION

FOR SEVERAL decades, fluorescent lamps have been the most important light sources due to their high efficiency and long life cycles. These advantages are particularly highlighted by high-frequency electronic ballasts that replaced electromagnetic ones [1], [2]. Conventionally, electronic ballasts are realized with a diode-bridge rectifier with a bulky capacitor followed by a high-frequency load resonant inverter, providing a high ignition voltage and, then, a stable arc current. Such an electronic ballast circuit, however, introduces a highly distorted input current, resulting in a low power factor and high harmonic contents. To comply with the regulations, such as International Electrotechnical Commission 1000-3-2, an additional conversion stage for power factor correction (PFC) was incorporated between the diode-bridge rectifier and the load resonant inverter. These two-stage conversion circuits use more circuit components, leading to a lower efficiency as well as a higher product cost. Consequently, many single-stage electronic ballasts have sought solutions from integrating the PFC stage and the resonant inverter by sharing one or two active switches [3]–[13].

Among various topologies, the single-stage electronic ballast with a boost-type PFC appears to have the simplest circuit structure and control. It can achieve a high power factor by operating the boost converter at the discontinuous conduction mode (DCM), provided that the dc-link voltage is at least twice the peak of the line voltage. This solution, however, produces a considerably high voltage stress on the circuit components [14], [15]. On the other hand, the PFC with a buck–boost or flyback converter can achieve both high efficiency and high power factor without introducing an excessively high dc-link voltage when operated at DCM [10]–[13].

In these single-stage solutions, power MOSFETs are usually adopted as the active power switches for the operating frequency ranging from several tens to a hundred kilohertz. To eliminate switching-on losses of power MOSFETs, much attention needs to be paid to the zero voltage switching (ZVS) operation. This fact poses a problem to a buck–boost-type PFC. Because of the asymmetrical topology, the ZVS for the commonly shared active power switch is no longer retained [8]–[12]. Moreover, two extra fast recovery diodes are required to accomplish the buck–boost operation, resulting in higher conduction losses [10]–[12]. These all introduce more switching and conduction losses, obstructing the pursuit of high circuit efficiency. These drawbacks can be unraveled by the symmetrical topology, which employs two buck–boost converters sharing both active power switches of the resonant inverter [13]. However, this symmetrical approach needs two bulky capacitors to reduce voltage ripple on the dc link, which is caused by the alternating operation of two buck–boost converters at a low frequency of the input line source.

To solve the problem of the single-stage electronic ballast with buck–boost PFC, this paper proposes a novel configuration with a derivative buck–boost converter. Both active power switches can accomplish ZVS by designing appropriate circuit parameters. Following the description of the proposed circuit is the discussion of the operating principle, the design equations, and an exemplar design for a 36-W rapid-start fluorescent lamp. Finally, the experimental results verify the accuracy of theoretical analysis.

II. CIRCUIT CONFIGURATION AND OPERATION

Fig. 1 shows a two-stage high-power-factor electronic ballast. It consists of a buck–boost converter for PFC and a half-bridge series-resonant parallel-loaded inverter for driving the fluorescent lamp. In the buck–boost PFC converter, two coupled inductors may be employed instead of the inductor $L_b$ to adjust the dc-link voltage. The active power switch $S_2$ and the diode $D_1$ in the resonant inverter can simultaneously play the roles of the active power switch $S_b$ and the diode $D_f$ in
the PFC circuit. By integrating these components, a single-stage electronic ballast with embedded PFC circuit can be obtained.

Fig. 2 shows the circuit configuration of the proposed electronic ballast with an embedded buck-boost PFC converter. Two power MOSFETs $S_1$ and $S_2$ are adopted as the power switches of the inverter. Each MOSFET is composed of an active switch and an intrinsic antiparallel diode. The load resonant circuit of the inverter is formed by the fluorescent lamp and a resonant energy tank consisting of $C_s$ and $L_s$. A capacitor $C_f$ and a starting-aid circuit in parallel with the lamp are used for providing the filament current paths at the steady-state and preheating phases, respectively [16].

The embedded PFC converter shares the active power switch $Q_2$ of the bottom MOSFET and the antiparallel diode $D_1$ of the upper MOSFET of the inverter. Instead of using a single inductor, two coupled inductors $L_1$ and $L_2$ along with a blocking diode $D_3$ are used to accomplish buck-boost conversion. The turn ratio of the two coupled inductors is well designed to induce a voltage higher than the rectified input voltage on $L_2$ to block the current from the line source, so that the buck-boost operation can be accomplished. A small low-pass filter $L_m$ and $C_m$ is used to remove the high-frequency current harmonics at the input line.

With a commonly sharing active power switch, only one control circuit is needed. The two active power switches are gated by two complementary voltages $V_{gs1}$ and $V_{gs2}$ with a short dead time. Neglecting the dead time, the duty ratios of $S_1$ and $S_2$ are regarded as $(1 - d_r)$ and $d_r$, respectively. The resonant current of the inverter can be regulated by controlling the operation frequency and the duty ratio.

The ballast-lamp operation can be divided into three phases: preheating, ignition, and steady state. The starting-aid circuit only operates in the preheating phase. During this phase, the lamp is short circuited by turning on $S_a$, eliminating the glow current and providing a current path for preheating the cathode filaments. When the cathodes reach the emission temperature, $S_a$ turns off. Meanwhile, the operation frequency is increased. It will sweep through the load natural frequency in the case of the lamp being off. Then, a high voltage will appear across the lamp. This voltage will normally ignite the lamp.

At the steady state, $S_a$ remains off, and the operation frequency and the duty ratio are fine tuned to produce the required lamp power with appropriate filament heating.
Fig. 3. Operation modes.

The steady-state operation can be explained by five modes within one high-frequency cycle according to the conducting power switches, as shown in Fig. 3. For simplicity, the input filter is omitted, and the load resonant circuit is replaced by a current source $i_r$. Fig. 4 shows the theoretical waveforms in each mode. To achieve a high power factor, the buck-boost converter is operated at DCM. The operation frequency of the inverter is higher than the resonance frequency of the load resonant circuit to ensure that two active power switches are switched on at zero voltage. The details of each mode are described as follows.

**Mode I:** During this mode, the dc-link capacitor supplies a positive resonant current $i_r$, flowing through $Q_1$ to the load resonant circuit. This mode ends when $Q_1$ is switched off, and then, circuit operation enters to Mode II. Since the load circuit is designed to be inductive, $i_r$ is positive at the instant of switching off.

**Mode II:** Mode II begins as soon as $Q_1$ switches off. It should be noted that $Q_2$ is activated by the gate signal during this interval but will not be turned on since a positive resonant current $i_r$ freewheels through the antiparallel diode $D_2$ and the voltage $\nu_{S2}$ is clamped at a small voltage. At the same time, the rectified line voltage $v_{rec}$ is imposed on the inductor $L_1$. With the DCM operation, the currents in two coupled inductors are zero before entering Mode II, and the inductor current $i_{L1}$ increases linearly from zero with a slope proportional to the line voltage. Mode II ends at the time when $i_{L1}$ increases up to be equal to $i_r$, and $Q_2$ turns on softly at zero voltage.

**Mode III:** Throughout Mode III, $Q_2$ remains on. At first, $Q_2$ carries the current difference between $i_{L1}$ and $i_r$ and then carries both currents when $i_r$ resonates to a negative. $i_{L1}$ keeps increasing until Mode IV begins when $Q_2$ is switched off.

**Mode IV:** At the instant of switching off $Q_2$, $i_r$ is negative and will freewheel through $D_1$ to $C_{dc}$. In the meantime, the diode-bridge rectifier is reverse biased by designing a voltage across the inductor $L_2$ to be higher than the peak of the input voltage. Hence, $i_{L1}$ in the inductor $L_1$ now flows through the inductor $L_2$ into $C_{dc}$. At this moment, the magnetomotive force
Fig. 4. Theoretical waveforms at (a) high and (b) low line voltages.
in the coupled inductors should keep balance in ampere turns; hence,

\[ N_1 i_{L_1}(t_{2-}) = (N_1 + N_2) i_{L_1}(t_{2+}) \]  

(1)

where \( N_1 \) and \( N_2 \) are the turn numbers of the coupled inductors, and \( i_{L_1}(t_{2-}) \) and \( i_{L_1}(t_{2+}) \) are the currents through \( L_1 \) at the instants right before and after \( Q_2 \) is switched off. As shown in Fig. 4, \( i_{L_1} \) drops while \( i_{L_2} \) rises dramatically at the switching-off instant. Henceforth, \( i_{L_1} \) equals \( i_{L_2} \) and flows through \( D_1 \) and \( D_3 \) to charge \( C_{dc} \). The negative dc-link voltage is imposed on the two coupled inductors, and the current declines linearly to zero.

As explained earlier, the rising slope of \( i_{L_1} \) is proportional to the rectified input voltage and so is the peak value of \( i_{L_1} \) in each switching cycle at a constant duty ratio. Therefore, the duration for \( i_{L_1} \) declining to zero varies with the rectified input voltage. As a result, there are two possible modes following Mode IV, depending on whether the current \( i_{L_1} \) or \( i_r \) reaches zero first.

At a high input voltage, \( i_r \) resonates to zero and then becomes positive before \( i_{L_1} \) reaches zero. Thereafter, \( D_1 \) keeps on conducting, carrying the difference between \( i_{L_1} \) and \( i_r \). Mode IV ends when the decreasing \( i_{L_1} \) is equal to the increasing \( i_r \), and then, the operation enters Mode V-a.

**Mode V-a:** As the same switching action of \( Q_2 \), \( Q_1 \) has been activated during Mode IV but is actually turned on until \( i_r \) becomes greater than \( i_{L_1} \). In other words, \( Q_1 \) is turned on softly with ZVS, too. When the inductor current decreases to zero, the operation enters Mode I of the next cycle.

On the other hand, when the input voltage is low, \( i_{L_1} \) decreases to zero earlier than \( i_r \) does. In this case, Mode V-b follows Mode IV, instead of Mode V-a.

**Mode V-b:** In this mode, \( i_{L_1} \) remains zero, and \( i_r \) flows through \( D_1 \). This mode ends at the time when \( i_r \) resonates to zero. Then, \( Q_1 \) is turned on with ZVS to carry the positive \( i_r \). The operation returns to Mode I of the next cycle.

From the mode operation, one can find that both \( Q_1 \) and \( Q_2 \) can retain ZVS, leading to relatively lower switching losses.

### III. Circuit Analysis

For simplifying the theoretical analysis, the following assumptions are made.

1) All circuit components are ideal.
2) The quality factor of the load resonant circuit is high enough to ensure a sinusoidal resonant current.
3) The capacitance of \( C_{dc} \) is large enough, so that the dc-link voltage \( V_{dc} \) can be regarded as a dc voltage source.
4) The lamp arc is regarded as an open circuit before ignition and a resistance at the steady-state operation.

According to the circuit operation described earlier, the features of the buck–boost converter and the load resonant inverter remain the same, even though they share some switches. Therefore, the electronic ballast can be treated as two independent conversions, which are the buck–boost PFC and the load resonant inverter.

**A. Buck–Boost Power Factor Corrector**

The electronic ballast is supplied from the ac line source

\[ \nu_s(t) = V_m \sin(2\pi f_s t) \]  

(2)

where \( f_s \) and \( V_m \) are the frequency and amplitude of the line voltage source, respectively. Based on the fact that \( f_s \) is much lower than the inverter’s operation frequency \( f_r \), the rectified line voltage can be regarded as a constant over a high-frequency cycle of the inverter.

To ensure the buck–boost conversion operation, the voltage across the inductor \( L_2 \) must be always higher than the input line voltage to block the current from the ac line source

\[ \nu_{L2} = \left( \frac{\alpha}{\alpha + 1} \right) V_{dc} \geq V_m \]  

(3)

where \( \alpha \) is the turn ratio of \( N_2 \) to \( N_1 \).

To achieve a high power factor, the buck–boost converter is operated at DCM over an entire line frequency cycle. Hence, \( i_{L_1} \) increases from zero at the beginning of Mode II and reaches its peak at the end of Mode III. It is noted that only the rising part of the \( i_{L_1} \) is supplied from the line source. In other words, the line source supplies a current to the buck–boost converter during Modes II and III. The conceptual waveform of \( i_{rec} \) is shown in Fig. 5. The rectified input current \( i_{rec} \) is equal to \( i_{L_1} \), and the peaks follow a sinusoidal envelope

\[ i_{rec, peak}(t) = \frac{V_m |\sin(2\pi f_s t)|}{L_1 f_s} d_r. \]  

(4)

Then, the average of the input current in every switching period can be expressed as

\[ i_{in, avg}(t) = \frac{V_m d_r^2}{2L_1 f_s} \sin(2\pi f_s t). \]  

(5)

Equation (5) indicates that the input current is proportional to and in phase with the ac line voltage if the operation frequency and the duty ratio of the inverter remain constant over a line cycle, leading to a high power factor.

The input power can be obtained by taking the average of the instantaneous line power over one line frequency cycle

\[ P_{in} = \frac{1}{2\pi} \int_{0}^{2\pi} \nu_s(t) \cdot i_{in}(t) d(2\pi f_s t) = \frac{V_m^2 d_r^2}{4L_1 f_s}. \]  

(6)
Fig. 6. Equivalent circuits of the load resonant inverter. (a) Preheating. (b) Ignition. (c) Steady state.

B. Series-Resonant Parallel-Loaded Inverter

The rectangular input voltage of the load resonant inverter $V_{ab}$ can be expressed as

$$\nu_{ab} = (1 - d_r) V_{dc} + \sum_n \left[ \sqrt{2} V_{dc} \sqrt{1 - \cos(2n\pi d_r)} \sin(2n\pi f_s t + \pi + \theta_n) \right]$$

(7)

where

$$\theta_n = \tan^{-1} \left( \frac{\sin(2n\pi d_r)}{1 - \cos(2n\pi d_r)} \right).$$

With a high load quality factor, the load resonant circuit can filter out almost all the harmonic contents, as well as the dc term. Only the fundamental current flows through the load resonant circuit. The rms value of the fundamental component of $\nu_{ab}$ is given by

$$V_1 = \frac{\sqrt{2} V_{dc} \sin(\pi d_r)}{\pi}.$$  

Using fundamental approximation, the equivalent circuits of the load resonant inverter for the phases of preheating, ignition, and steady state are shown in Fig. 6. At the steady state, the fluorescent lamp is represented by a power-dependent resistance $R_{arc}$ and a resistance $r_f$ for each cathode filament [17], [18].

1) Preheating: During the preheating interval, the starting-aid circuit is equivalently short circuited by turning on the switch $S_a$, and the fluorescent lamp is regarded as an open circuit before being ignited. It ensures that no glow current will occur during the preheating phase [16]. The equivalent circuit of the half-bridge series-resonant parallel-loaded inverter is shown in Fig. 6(a). With a preheating frequency of $f_p$, the reactance of $L_s$ and $C_s$ in series is

$$X_{sp} = 2\pi f_p L_s - \frac{1}{2\pi f_p C_s}.$$  

(9)

and the preheating current $I_p$ can be expressed in terms of $V_1$

$$I_p = \frac{V_1}{\sqrt{4r_f^2 + X_{sp}^2}}.$$  

(10)

The resonant inverter can produce a high preheating current when $f_p$ is close to the natural resonance frequency $f_{rp}$

$$f_{rp} = \frac{1}{2\pi \sqrt{L_s C_s}}.$$  

(11)

The preheating current is preferred to be as high as possible to accomplish a rapid lamp starting, but it should be limited by its rated filament current to prevent the lamp from being damaged. As indicated by (6), the input power is a function of the duty ratio. During preheating, the input power is consumed only by the filaments. Therefore, the duty ratio for preheating should be much smaller than that for the steady-state operation. A proper preheating current can be obtained by adjusting the duty ratio of the active switches

$$d_p = \sqrt{\frac{4L_1 f_p P_f}{V_m^2}} = \frac{I_p}{V_m} \sqrt{8L_1 r_f f_p}.$$  

(12)

2) Ignition: After the cathode filaments have been preheated to an appropriate emission temperature, the starting-aid circuit is open circuited. At this phase, the equivalent circuit of the load resonant inverter is shown in Fig. 6(b), and the natural resonance frequency of the circuit becomes higher

$$f_{ri} = \frac{1}{2\pi \sqrt{L_s \left( \frac{C_s C_f}{C_s + C_f} \right)}}.$$  

(13)

Neglecting the filament resistance, the voltage for ignition can be expressed as

$$V_{ig} = \frac{C_s}{4\pi^2 f_{ri}^2 L_s C_s C_f - (C_s + C_f)} V_1.$$  

(14)

Substituting (13) into (14) yields

$$V_{ig} = \frac{C_s}{\left( \frac{f_{ri}^2}{f_{ri}^2 - 1} \right) (C_s + C_f)} V_1.$$  

(15)

A high voltage can be obtained for ignition by operating the inverter at a frequency close to the resonant frequency $f_{ri}$.

3) Steady State: At the steady state, the inverter is operated at the designated frequency $f_s$. The reactance of the energy tank becomes

$$X_{ss} = 2\pi f_s L_s - \frac{1}{2\pi f_s C_s}.$$  

(16)

The quality factor of the load circuit is given by (17), shown at the bottom of the next page. Then, the resonance frequency can be expressed in term of $Q_L$

$$f_{rs} = \frac{f_{ri}}{\sqrt{1 - \frac{1}{Q_L^2}}}.$$  

(18)

According to (17), $f_{rs}$ is lower than $f_{ri}$. With a high quality factor, the $f_{rs}$ is very close to $f_{ri}$.

Referring to Fig. 6(c), the filament current can be obtained as

$$I_f = \frac{V_{arc}^2}{r_f - f_{ri}^2 f_p C_f}.$$  

(19)
Then, the fundamental voltage of the inverter output can be expressed as

\[ V_1' = V_{arc} + (I_{arc} + I_f)(r_f + jX_{ss}). \]  

(20)

Substituting (19) into (20), the relationship between \( V_1 \) and \( V_{arc} \) can be described by (21), shown at the bottom of the page.

For a rapid-start fluorescent lamp, a filament current is required to maintain the cathodes at the emission temperature for the steady-state operation [19], [20]

\[ I_f = \frac{2\pi f_s C_f V_{arc}}{1 + (2\pi f_s r_f C_f)^2}. \]  

(22)

4) ZVS Operation: As shown in Fig. 6(c), the total impedance of the load resonant circuit can be expressed as in (23), shown at the bottom of the next page. The load angle \( \psi_1 \) of \( Z_{in} \) is given in (24), shown at the bottom of the next page. The phase angle of the fundamental voltage in (7) is

\[ \theta_1 = \tan^{-1} \left( \frac{\sin(2\pi d_r)}{1 - \cos(2\pi d_r)} \right) = \left( \frac{1}{2} - d_r \right) \pi. \]  

(25)

Then, the load resonant current can be expressed as

\[ i_r(t) = \frac{\sqrt{2V_1}}{Z_{in}} \sin(2\pi f_s t + \pi + \theta_1 - \psi_1). \]  

(26)

To fulfill the requirements of ZVS, an inductive load circuit is necessary so that the load current \( i_r \) lags the voltage \( V_1 \) by a phase angle. Therefore, the load angle \( \psi_1 \) should be designed to meet the following:

\[ \psi_1 \geq \theta_1 = \left( d_r - \frac{1}{2} \right) \pi, \quad \text{when} \quad d_r \geq 0.5 \]  

(27)

\[ \psi_1 \geq \theta_1 = \left( \frac{1}{2} - d_r \right) \pi, \quad \text{when} \quad d_r < 0.5. \]  

(28)

5) Switching Frequency: It should be noted that the resonance frequency at the preheating phase \( f_{rp} \) is lower than those for the ignition \( f_{ri} \) and the steady state \( f_{rs} \). To ensure a sufficiently high voltage for ignition, the switching frequency is scheduled to start from a preheating frequency \( f_p \), then goes through \( f_{ri} \), and eventually to the steady state \( f_s \). On the other hand, the ballast circuit is preferred to be with an inductive load to achieve ZVS operation. As stated earlier, \( f_{ri} \) and \( f_{rs} \) are very close to each other. For these reasons, \( f_p \) is designed to be slightly higher than \( f_{rp} \) but lower than \( f_{ri} \), and the steady-state frequency \( f_s \) is higher than \( f_{ri} \) and \( f_{rs} \). The variation of the switching frequency is shown in Fig. 7. Since both \( f_p \) and \( f_s \) are higher than \( f_{rp} \) and \( f_{rs} \), respectively, the ZVS operation can be retained for both preheating and steady-state phases.

IV. Design Example

An electronic ballast designed for a T8-36W rapid-start fluorescent lamp is taken as an illustrative example. Table I lists the circuit specifications, which aim at operating the buck–boost PFC at DCM and operating the active power switches with ZVS to achieve a high power factor and circuit efficiency. The designed procedure is described as follows.

\[
Q_L = \sqrt{\left\{ \frac{L_i \left[ 1 + 4\pi^2 f^2 C_f \left[ C_f R_{arc}^2 + C_f r_f (r_f + R_{arc}) \right] \right]}{C_f \left[ r_f^2 + 1 + 4\pi^2 f^2 C_f^2 \left[ r_f (r_f + R_{arc}) \right] \right]} \right\} \left( f_f + \frac{R_{arc} \left[ 1 + 4\pi^2 f^2 C_f^2 \left( r_f + R_{arc} \right) \right]}{1 + 4\pi^2 f^2 C_f^2 \left( r_f + R_{arc} \right)^2} \right)}
\]  

(17)

\[
V_{arc} = \frac{V_1}{\sqrt{\left( 1 + \frac{r_f}{R_{arc}} + \frac{4\pi^2 f^2 r_f C_f^2}{1 + 4\pi^2 f^2 r_f^2 C_f^2} \right)^2 + \left( \frac{X_{ss}}{R_{arc}} + \frac{2\pi f_s r_f C_f \left[ 1 + 2\pi f_s r_f C_f X_{ss} \right]}{1 + 4\pi^2 f^2 r_f^2 C_f^2} \right)^2}}
\]  

(21)
**Step 1**—Determine $V_{dc}$, $\alpha$, and $d_r$ at Steady State: The dc-link voltage $V_{dc}$ should be high enough so that the current in the coupled inductor always declines to zero in every high-frequency cycle

$$V_{dc} \geq \frac{d_r}{(1 - d_r)} V_m.$$  \hfill (29)

To meet the requirement of buck–boost conversion, the turn ratio of the two coupled inductors is determined from (3)

$$\alpha \geq \frac{V_m}{V_{dc} - V_m}.$$  \hfill (30)

From (29) and (30), it can be found that $V_{dc}$ can be lower with a larger turn ratio and a smaller duty ratio. However, a larger turn ratio and a smaller duty ratio are with a smaller inductance $L_1$, leading to a higher peak value in $i_{L1}$, resulting in higher current stress and losses in the active power switches. In the illustrative case, $V_{dc}$ is set at 285 V, which is slightly larger than the theoretical calculation of 259 V, which is the compromise by a turn ratio $\alpha$ of 1.5 and a duty ratio of 0.4.

**Step 2**—Calculate $L_1$: For a given lamp power $P_{lamp}$ and an estimated circuit efficiency $\eta$, $L_1$ can be calculated as

$$P_{lamp} = P_{in} \cdot \eta = \frac{V_m^2 f \pi r_f^2}{4 L_1 f_s} \cdot \eta.$$  \hfill (31)

In this case, $L_1$ is obtained as 0.756 mH by assuming a circuit efficiency of 90%.

**Step 3**—Calculate $C_f$: The capacitance $C_f$ can be obtained from (22)

$$C_f = \frac{1}{2 \pi f_s \sqrt{\left(\frac{V_m}{f s r_f}\right)^2 - r_f^2}} = 16.4 \text{ (nF)}.$$  \hfill (22)

**Step 4**—Calculate $d_p$, $L_s$, and $C_s$: For a given preheating frequency of 25 kHz and assuming a circuit efficiency of 80%, the duty ratio during preheating is obtained from (12)

$$d_p = \frac{I_p}{V_m} \sqrt{\frac{8 L_1 r_f f_s}{\eta_p}} = 0.09.$$  \hfill (12)

Then, $V_1$ during preheating is calculated from (8)

$$V_1 = 35.8 \text{ (V)}.$$  \hfill (8)

$$Z_{in} = r_f + \frac{R_{arc} \left(1 + 4 \pi^2 r_f f_s^2 C_f^2 (r_f + R_{arc}) \right)}{1 + 4 \pi^2 f_s^2 C_f^2 (r_f + R_{arc})^2} + j \left(2 \pi f_s L_s - \frac{4 \pi^2 f_s^2 R_{arc} C_f C_s + \left[1 + 4 \pi^2 f_s^2 C_f^2 (r_f + R_{arc})^2 \right]}{2 \pi f_s C_s \left[1 + 4 \pi^2 f_s^2 C_f^2 (r_f + R_{arc})^2 \right]} \right)$$  \hfill (23)

$$\psi_1 = \tan^{-1} \left(\frac{\left[4 \pi^2 f_s^2 L_s C_s - 1\right] \left[1 + 4 \pi^2 f_s^2 C_f^2 (r_f + R_{arc})^2 \right] - 4 \pi^2 f_s^2 R_{arc} C_f C_s}{2 \pi f_s C_s \left[r_f + R_{arc} \right] \left[1 + 4 \pi^2 r_f f_s^2 C_f^2 (r_f + 2 R_{arc}) \right]} \right)$$  \hfill (24)

**TABLE II**

<table>
<thead>
<tr>
<th>Circuit Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC-link voltage, $V_{dc}$</td>
<td>285 V</td>
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<tr>
<td>Duty-ratio (steady-state), $d_r$</td>
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</tr>
<tr>
<td>Turn ratio, $\alpha$</td>
<td>1.5</td>
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<tr>
<td>Buck-boost inductance, $L_1$</td>
<td>0.756 mH</td>
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<tr>
<td>Buck-boost inductance, $L_2$</td>
<td>1.701 mH</td>
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<tr>
<td>Parallel capacitance, $C_f$</td>
<td>16.4 nF</td>
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<tr>
<td>Resonant inductance, $L_s$</td>
<td>4.24 mH</td>
</tr>
<tr>
<td>Resonant capacitance, $C_s$</td>
<td>10.29 nF</td>
</tr>
</tbody>
</table>

From (10), $X_{sp}$ can be obtained by

$$X_{sp} = \sqrt{\left(\frac{V_1}{I_p}\right)^2 - 4 r_f^2} = 47.54 \text{ (} \Omega \text{)}.$$  \hfill (10)

By squaring (21) and rearranging their terms yield

$$1 + (2 \pi f_s C_f (R_{arc} + r_f))^2 X_{ss}^2 - 4 \pi f_s C_f R_{arc} X_{ss}$$

$$+ (R_{arc} + r_f)^2 [1 + 4 \pi^2 f_s^2 C_f^2 (r_f)^2] + 4 R_{arc} (3 R_{arc} + 2 r_f)$$

$$\times \pi^2 f_s^2 C_f^2 r_f^2 - R_{arc} [1 + 4 \pi^2 f_s^2 C_f^2 (r_f)^2] \left(\frac{V_1}{V_{arc}}\right)^2 = 0.$$  \hfill (32)

$X_{ss}$ is then given by (33), shown at the bottom of the next page.

There are two solutions in (33). However, only the positive one is preferred to exhibit an inductive load. Thus, $X_{ss} = 369.5 \Omega$ is the only solution.

After solving $X_{ss}$ and $X_{sp}$, $L_s$ and $C_s$ can be obtained from the following:

$$L_s = \frac{f_s X_{ss} - f_p X_{sp}}{2 \pi \left(\frac{f_s^2 - f_p^2}{f_p}\right)}$$  \hfill (34)

$$C_s = \frac{f_s^2 - f_p^2}{2 \pi f_s f_p (f_p X_{ss} - f_s X_{sp})}.$$  \hfill (35)

Solving (34) and (35) yields $L_s = 4.24 \text{ mH}$ and $C_s = 10.29 \text{ nF}$.

Table II lists the circuit parameters. In the design case, the resonance frequency of the resonant circuit is 23 kHz during preheating, while the natural resonance frequency at the ignition transient is 30.6 kHz. A high ignition voltage can...
be promised when the inverter frequency is adjusted from the preheating frequency, passing through the resonance frequency, to the steady-state operation frequency.

V. EXPERIMENTAL RESULTS

A prototype of the proposed electronic ballast was built and tested. Fig. 8 shows the starting transient waveforms. After being switched on, electronic ballast is operated at 25 kHz, which is, to some extent, higher than the resonance frequency of 23 kHz, generating an approximately constant filament current of 0.75 A for preheating. The preheating interval lasts for 1 s. With the starting-aid circuit, no glow current is found in the lamp. Once the cathode filaments reach the appropriate emission temperature, the starting-aid circuit is opened, and the operation frequency is adjusted from 25 to 32 kHz. As the operation frequency passes through a resonance frequency of 30.6 kHz, a high voltage is produced across the lamp for ignition. After being started up, the lamp is operated at 36 W. Meanwhile, the filament current decreases to about 0.35 A to maintain the cathode filaments at the proper emission temperature.

The measured waveforms of $\nu_{ab}$, $i_r$, $i_{L1}$, and $i_{L2}$ at the steady-state operation are shown in Fig. 9. The experimental results are well in agreement with the theoretical prediction seen in Fig. 4. Fig. 10(a) shows the dc-link voltage and the lamp current for six cycles of the ac line source. The lamp voltage and current waveforms in Fig. 10(b) show that the lamp current is nearly sinusoidal. The measured crest factor is below 1.48. Fig. 11 shows the voltage and current waveforms of the active power switches at the steady-state operation, indicating that the ZVS operation for the active switches of the ballast circuit can be retained.

Fig. 12 shows the waveforms of the input voltage and current along with the rectified input current. It can be observed that the buck-boost PFC is operated at DCM over the entire cycle of the line source. The input current is sinusoidal and in phase with the input voltage. The measured power factor is greater than 0.99, and the total current harmonic distortion (THD) is

$$X_{ss} = \frac{2\pi f_s C_f R_{arc}^2 \pm \sqrt{4\pi^2 f_s^2 C_f^2 R_{arc}^4 - \left[1 + (2\pi f_s C_f (R_{arc} + r_f))\right]^2 \left((R_{arc} + r_f)^2 \frac{1 + 4\pi^2 f_s^2 C_f^2 r_f^2}{4 R_{arc} (3R_{arc} + 2r_f) \pi^2 f_s^2 C_f^2 r_f^2} - R_{arc}^2 \left(1 + 4\pi^2 f_s^2 C_f^2 r_f^2 \left(\frac{V_{dc}}{V_{arc}}\right)^2 \right)\right]}}{1 + (2\pi f_s C_f (R_{arc} + r_f))^2}$$

(33)
Fig. 11. Switching voltage and current waveforms at steady state.

Fig. 12. Waveforms of $v_a$, $i_s$, and $i_{sec}$.

less than 9%. The input and output powers were 38.5 and 36.2 W, respectively. With ZVS operation on the two active power switches, the circuit efficiency is as high as 0.94.

VI. CONCLUSION

A novel single-stage high-power-factor electronic ballast has been presented. The circuit topology is based on the integration of a half-bridge resonant inverter for ballasting the fluorescent lamp and a derivative buck–boost converter for PFC. The elaborate design on circuit parameters ensures that both the active power switches can achieve ZVS features, leading to a high efficiency.

The circuit operation is described, and the design equations are derived. A prototype circuit designed for a T8-36W fluorescent lamp is built and measured to verify the theoretical analyses. Experimental results show that the ballast-lamp circuit performs satisfactorily from preheating through ignition to the steady state. Under the nominal operation conditions, a nearly unity power factor and a THD of less than 9% can be achieved. With ZVS operation on both active power switches, the electronic ballast has a high circuit efficiency of 0.94.

REFERENCES

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