A Self-Excited Half-Bridge Series-Resonant Ballast with Automatic Input Current Shaping

Yueh-Ru Yang

Chern-Lin Chen

Power Electronics Laboratory
Department of Electrical Engineering
National Taiwan University
Taipei, Taiwan 10764

Abstract - A single-stage electronic ballast with power factor correction and soft switching is presented and analyzed. This circuit utilizes half-bridge series-resonant power stage to obtain the zero-voltage-switching clamping-voltage merits. The input current waveform is automatically shaped by an additional high-frequency inductor. Near unity power factor can be achieved.

I. Introduction

Fluorescent lamps which exhibit negative resistance characteristics require ballast of positive impedance characteristics to limit the lamp current. From the viewpoint of impedance, resistors, inductors, and capacitors are all candidates for ballast in ac circumstances. However, resistors are seldom utilized because it deteriorates the efficiency of fixture. Owing to the negative resistance features the voltage waveshape of lamp is similar to a square-wave when it works at line frequency. Since capacitor voltage can not change abruptly, which means capacitors can not absorb the abrupt changes of lamp voltage at each discharge instant, it is unsuitable for ballasting. On the other hand, using inductors as ballast are suitable because its voltage can change instantly and accompany a smooth current. For all this, the size, weight, and audible noise of line-frequency inductors are still annoying. The distorted lamp voltage and the line-frequency flickering also deserves to be improved for higher lamp efficacy. Such factors make high frequency electronic ballast receive lots of attention recently[1-5].

There are optimal diameters and current densities for fluorescent lamps. Hence, current control topologies are often utilized for ballast. The self-excited half-bridge series-resonant inverter as shown in Fig.1 which possesses current control features and zero-voltage-switching clamping-voltage (ZVS-CV) merits is widely employed[6]. A current transformer, wound on a saturable core[7], drives the two power switches. Filament preheating and overcurrent protection are inherent due to series resonance and current driving, respectively. One serious drawback of this circuit is

the low input power factor, around 0.6, owing to bulk capacitor in series with diode rectification.

An electronic ballast usually consists of power factor corrector, high-frequency dc/ac inverter, and control circuitry. In this paper, a single-stage electronic ballast with power factor correction and soft switching is presented. This circuit, as shown in Fig.2, utilizes the half-bridge series-resonant power stage to obtain the ZVS-CV merits. The input current waveform is automatically shaped by an additional high-frequency inductor. Near unity power factor can be achieved.

Fig.1: A self-excited half-bridge series-resonant ballast (starting circuit is not shown)

Fig.2: A self-excited half-bridge series-resonant ballast with automatic input current shaping

II. Circuit Operations

In the presented single-stage ballast, as shown in Fig.2, the switching mechanism of the automatic input current shaper is attached on the dc/ac inverter. Hence, the dc/ac inverter is analyzed first and then the current shaper adopts the inverter switching pattern.

2.1 Dc/ac Inverter

In the positive(negative) resonant half cycle, operations can be divided into three stages: transistor-on, resonant-transition, and diode-on, as shown in Fig.3a.~c.(d.~f.). In the transistor-on stage, the current transformer(s.c.) drives the transistor $Q_2(Q_1)$ into deep-saturation. The inverter behaves as a RLC circuit supplied by $C_2(C_1)$. In the resonant-transition stage, transistor $Q_2(Q_1)$ comes into active region and becomes a high-impedance component. The resonant current flows through the transition-capacitors C_z . As the C_z is charged to the half dc bus voltage, The inverter comes into the diode-on stage. In this stage, the RLC circuit returns energy to $C_1(C_2)$ through D_1 . As the resonant current decays to zero, the diode-on stage ends and next transistor-on stage starts.

Fig.3: Dc/ac inverter operations in positive(negative) resonant half cycle, a.~c.(d.~f.)

2.2 Automatic current shaper

The additional high-frequency inductor(L_s) is charged by the ac input source and discharged to the dc bus. Its

operations are controlled by the dc/ac inverter operations. During the positive(negative) half cycle of the input source, L_s combined with $Q_2(Q_1)$ and fast recovery diode $D_1(D_2)$ forms a boost-type input power factor corrector, as shown in Fig.4a. & b.(c. & d.). Because the input current shaper adopts the switching pattern of dc/ac inverter, the inductor current in L_s is discontinuous.

Fig.4: Automatic current shaper operations in positive(negative) ac source half cycle, a. & b.(c. & d.)

III. Mathematical Analysis

Some phenomena about fluorescent lamps must be recognized in advance.

(a) As long as the discharge current is restricted, the lamp voltage depends on the length of arc. For a ready-made lamp these two quantities which exhibit nonlinear relationships determine its consumed power and operating resistance.

(b) The current waveshape and frequency are all concerned with lamp life expectancy. Hence the current crest factor(CCF) and switching frequency must be considered at first.

(c) The negative tendency of lamp resistance is not obvious above audio frequency. Hence the lamp resistance can be regarded as a constant value at a fixed high frequency.

The lamp power of Fig.3 can be expressed by rms. resonant current and lamp resistance. It also can be approximated by the energy losses per second in equivalent capacitor(C_{eq}) or resonant inductor(L_r). The equivalent capacitor C_{eq} is the series combination of resonant capacitor and bulk capacitor C_1 .

$$P_{lamp} = [\text{rms}.i(t)]^2 R \quad (1a)$$

$$= \frac{1}{2} L_r \{ [\max.i(t)]^2 + [\min.i(t)]^2 \} f \quad (1b)$$

$$= \frac{1}{2} C_{eq} \{ [(V_{C1} + \max.v_{C_r}(t))^2 - (V_{C1} + \min.v_{C_r}(t))^2] \} f \quad (1c)$$

Ideally, the switching frequency(f) is a constant and V_{C1} equals V_{C2} . The resonant current of a series-connected RLC circuit, which supplied by a dc voltage source, except for critical-damped cases can be expressed as

$$i(t) = A_1 e^{s_1 t} - A_2 e^{s_2 t} \quad (2)$$

where $S_1 = -\alpha_0 + \sqrt{\alpha_0^2 - \omega_0^2}$, $S_2 = -\alpha_0 - \sqrt{\alpha_0^2 - \omega_0^2}$,

$$\alpha_0 = \frac{R}{2L_r}, \quad \omega_0 = \frac{1}{\sqrt{L_r C_r}},$$

$$A_1 = \frac{V_{C2} - v_{C_r}(0) + i(0)S_1 L_r}{(S_1 - S_2)L_r}, \quad A_2 = \frac{V_{C2} - v_{C_r}(0) + i(0)S_2 L_r}{(S_1 - S_2)L_r}.$$

Generally, the quality factor Q and characteristic impedance Z_0 are used to describe the circuit characteristics,

$$Q = \frac{\omega_0}{2\alpha_0}, \quad Z_0 = \sqrt{\frac{L_r}{C_r}} \quad (3)$$

Fig.2 is a low Q circuit due to the high equivalent resistance of fluorescent lamps. Its complex frequency can be rewritten as

$$S_1 = -\frac{\omega_0}{2Q} + j\omega_0 \sqrt{1 - \frac{1}{4Q^2}}, \quad S_2 = -\frac{\omega_0}{2Q} - j\omega_0 \sqrt{1 - \frac{1}{4Q^2}}.$$

The current and voltage of resonant capacitor with zero initial conditions are

$$i(t) = \frac{V_{C2}}{kZ_0} \cdot e^{-\alpha_0 t} \sin \omega_d t \quad (4)$$

$$v_{C_r}(t) = \frac{V_{C2}}{k} \cdot e^{-\alpha_0 t} \left(-\frac{R}{2Z_0} \sin \omega_d t - k \cdot \cos \omega_d t \right) + V_{C2} \quad (5)$$

where $k = \sqrt{1 - \frac{1}{4Q^2}}$ and $\omega_d = k\omega_0$.

From Eq.(2) or (4), the peak current occurs at

$$t_m = \frac{\ln A_2 S_2 - \ln A_1 S_1}{\omega_d} \tan^{-1} \frac{\omega_d}{\alpha_0}. \quad (6)$$

The current slope is dominated by the decaying time constant

$$T_d = \frac{1}{\alpha_0}. \quad (7)$$

The relationship between t_m and T_d can be derived from Eq.(6)&(7) and expressed as Eq.(8). It can be viewed as a

reference index to choose the proper value of Q . To obtain low CCF and high switching frequency, the index generally approaches unity.

$$\frac{t_m}{T_d} = \frac{\alpha_0}{\omega_d} \tan^{-1} \frac{\omega_d}{\alpha_0} = \frac{\tan^{-1} \sqrt{4Q^2 - 1}}{\sqrt{4Q^2 - 1}} \quad (8)$$

Eq. (8) is depicted by Fig.5. The relationship among k , kZ_0 , and Cr/Lr at different lamp resistance is shown in Fig.6. It shows that the needed ratio of Cr/Lr for different resistance at a fixed value of k .

Fig.5: Relationship between t_m/T_d and Q

Fig.6: Relationship among k , kZ_0 , and Cr/Lr

In Eq.(4), if the resonant current starts at $t=0$, current transformer(sc) saturates at time $t_{sat,sc}$, transistor is turned off at t_{on} , and the peak current of nature response occurs at time t_m , then the following constraints must be satisfied.

$$N_{sc} \cdot i(t_{sat,sc}) = H_{sat,sc} \cdot \ell_{sc} \quad (9)$$

$$N_{Lr} \cdot i(t_{sat,sc}) < H_{sat,Lr} \cdot \ell_{Lr} \quad (10)$$

$$i_b(t) = \frac{N_P}{N_S} i(t) \gg \frac{1}{h_{FE}} i(t), \quad t \leq t_{sat,sc} \quad (11)$$

where $t_{sat,sc} < t_{on} < t_m$ and $i(t_{sat,sc}) < i(t_{on}) < i(t_m)$.

The saturable core of current transformer must satisfy the constraint of Eq.(9), and the core of resonant inductor(L_r) must satisfy the constraint of Eq.(10). Eq.(11) is the constraint of base-driving. Since the high current injection of Eq.(11) saturates the transistor, the storage time(Δt_{stg}) must be analyzed by transistor model[7]. Hence, the duration of transistor-on stage(Δt_{on}) equals the summation of $\Delta t_{sat,sc}$ and Δt_{stg} . The effective current of Eq.(4) can be

estimated by a postulated current crest factor(CCF) expectancy

$$\text{rms.}i(t) = \frac{\max.i(t)}{CCF} \quad (12)$$

where $\max.i(t)=i(\text{ton})$. The crest factor is concerned with the lifetime of lamp and must be specified with a proper value[8].

In the resonant-transition stage, the resonant current increases slightly and the collector-emitter voltage changes linearly due to the transition-capacitors(C_z).

In the diode-on stage, the current expression is the same as Eq.(2) except that initial conditions are different.

$$i(\tau) = A_3 e^{s_1 \tau} - A_4 e^{s_2 \tau}, \quad \tau > 0 \quad (13)$$

where

$$A_3 = \frac{i(t_{on})S_1 L_r - v_{C_z}(t_{on}) - V_{C_1}}{(S_1 - S_2)L_r}, \quad A_4 = \frac{i(t_{on})S_2 L_r - v_{C_z}(t_{on}) - V_{C_1}}{(S_1 - S_2)L_r}.$$

If $i(\tau)$ falls to zero at τ_{on} , then from Eq.(13) we can obtain the duration of diode-on stage($\Delta\tau_{on}$)

$$\Delta\tau_{on} = \frac{\ln A_4 - \ln A_3}{S_1 - S_2} \quad (14)$$

and the switching period $T \approx 2(\Delta\tau_{on} + \Delta\tau_{off})$.

Shown in Fig.7 is the simulation of resonant current and voltage. It uses the following parameters: $P_{lamp}=18W$, $R_{lamp}=180\Omega$, $V_{C_1}=V_{C_2}=140V$, $L_r=2mH$, $C_r=0.1\mu F$, core TDK H5B2 T5-10-2.5 $H_{sat,sc}=2$ Oersted, $I_{sc}=2.18cm$, $N_p/N_s=8/2$, $C_z=4nF$.

Fig.7: Simulation of resonant current and voltage

To ensure discontinuous boost operations and volt-seconds balance, as shown in Fig.4, the discharge duration of L_s should be :

$$t_{dis}^*(t) = \sum_{k=0}^{\infty} \frac{V_m |\sin \omega_s t \cdot \delta(t - kT)| t_{on}}{V_d - V_m |\sin \omega_s t \cdot \delta(t - kT)|} \quad (15)$$

where the average voltage of dc bus $V_d = V_{C_1} + V_{C_2}$.

Since the boost operation is attached to dc/ac inverter, Eq.(15) must satisfy

$$\max.t_{dis}^*(t) = \frac{V_m t_{on}}{V_d - V_m} \leq \tau_{on} \quad (16)$$

The discrete peak inductor current can be expressed as

$$i_{pk,L_s}^*(t) = \sum_{k=0}^{\infty} I_{pk,L_s} \sin \omega_s t \cdot \delta(t - kT) \quad (17)$$

where $I_{pk,L_s} = (V_m/L_s)t_{on}$.

Shown in Fig.8 is the simulation of input inductor current under $V_d=2V_m$, $t_{on}/T=0.25$, $\omega_s=377$ radian/sec. and switching frequency 2.52kHz. This input inductor current adds to the resonant current in transistor.

Fig.8: Simulation of input inductor current

$$i_{pk,Q}^* = i_{pk,L_s}^*(t) + i(t_{on}) \quad (18)$$

Hence, Eq.(11) must be modified as

$$i_b(t) \gg \frac{1}{h_{FE}} \left[\frac{V_m}{L_s} \sin \omega_s t + i(t) \right], \quad t \leq t_{sat,sc} \quad (19)$$

The discrete average current of L_s can be expressed as

$$i_{avg,L_s}^*(t) = \frac{t_{on} + t_{dis}^*(t)}{2T} i_{pk,L_s}^*(t) \quad (20)$$

With high-frequency filtering, its low-frequency component is continuous.

$$i_{avg,L_s}^*(t) \approx i_{L_s}(t) = \left(\frac{t_{on}^2 V_m}{2 L_s T} \right) \left(\frac{\sin \omega_s t}{1 - \alpha |\sin \omega_s t|} \right) \quad (21a)$$

where voltage ratio $\alpha = V_m/V_d$. Moreover, in positive ac source half cycle, the following relationships exist.

$$i_{avg,L_s}^*(t) = i_{avg,Q_2}^*(t) + i_{avg,D_1}^*(t), \quad 0 < \omega_s t < \pi. \quad (21b)$$

Similarly, the switch and diode currents can also be regarded as continuous after high frequency filtering.

$$i_{Q_2}(t) = i_{L_s}(t) \cdot (1 - \alpha \sin \omega_s t), \quad 0 < \omega_s t < \pi \quad (21c)$$

$$i_{D_1}(t) = i_{L_s}(t) \cdot \alpha \sin \omega_s t, \quad 0 < \omega_s t < \pi \quad (21d)$$

$$i_{Q_1}(t) = -i_{L_s}(t) \cdot (1 - \alpha |\sin \omega_s t|), \quad \pi < \omega_s t < 2\pi \quad (21e)$$

$$i_{D_2}(t) = i_{L_s}(t) \cdot \alpha \sin \omega_s t, \quad \pi < \omega_s t < 2\pi \quad (21f)$$

Shown in Fig.9 is the simulation of normalized currents in inductor(L_s), switch(Q_2), and diode(D_1), but the resonant current in switch Q_2 is not included.

Fig.9: Simulation of the averaging currents in positive ac source half cycle and $V_m/V_d=0.5$

Because the diode current equals the charging current, the ripple of steady state dc bus voltage in one ac source cycle can be expressed as

$$v_{d,ac}(t) = \frac{1}{C} \int_0^t [i_{D_1}(t) + i_{D_2}(t)] dt - V_d \quad (22a)$$

where C is the series combination of C_1 and C_2 . If power factor equals unity,

$$V_m |\sin \omega_s t| \cdot I_m |\sin \omega_s t| = V_d \cdot [i_{D_1}(t) + i_{D_2}(t)] \quad (22b)$$

, then the dc bus voltage can be written as

$$v_d(t) = \frac{\alpha I_m t}{C} - \frac{\alpha I_m}{4\omega_s C} \sin 2\omega_s t = V_d + v_{d,ac}(t). \quad (22c)$$

From Eq.(22c), it is obvious that at

$$t_{ss} = \frac{V_d C}{\alpha I_m} \quad (22d)$$

the dc bus voltage comes into steady state. The ripple voltage, Eq.(22a), results in switching frequency drift. This drift can be estimated by Eq.(4). The relationship between input power and inductor L_s can be derived[Appendix] as

$$P_i = \frac{1}{\pi} \int_0^\pi V_m \sin \omega_s t \cdot i_{L_s}(t) d\omega_s t = \frac{V_m^2 t_{on}^2}{2\pi T L_s} \cdot \beta \quad (23)$$

$$\text{where } \beta = \frac{1}{\alpha^2} \left[-\pi - 2\alpha + \frac{2}{\sqrt{1-\alpha^2}} \left(\frac{\pi}{2} + \tan^{-1} \frac{\alpha}{\sqrt{1-\alpha^2}} \right) \right].$$

The input rms. current can be obtained by definition

$$\text{rms. } i_{L_s}(t) = \left[\frac{1}{\pi} \int_0^\pi i_{L_s}^2(t) d\omega_s t \right]^{\frac{1}{2}} = \frac{V_m t_{on}}{2\pi T L_s} \sqrt{\frac{\gamma}{\pi}} \quad (24)$$

$$\text{where } \gamma = \frac{1}{\alpha^2} \left[\pi + \frac{2\alpha}{(1-\alpha^2)} + \frac{4\alpha^2-2}{\sqrt{(1-\alpha^2)^3}} \left(\frac{\pi}{2} + \tan^{-1} \frac{\alpha}{\sqrt{1-\alpha^2}} \right) \right].$$

Substituting Eq.(24) into the definition of input power factor,

$$PF = \sqrt{2} \frac{P_i}{V_m \cdot \text{rms. } i_{L_s}(t)} = \sqrt{2} \frac{\beta}{\sqrt{\pi \cdot \gamma}}. \quad (25)$$

Another expression of input rms. current can be derived from Eq.(26)

$$\text{rms. } i_{L_s}(t) = \frac{P_i \sqrt{\pi \cdot \gamma}}{V_m \cdot \beta}. \quad (26)$$

IV. Experiments And Results

A 18W prototype of the presented circuit is implemented and tested. Fig.10 shows that when the transistor-on stage ends, Q_2 is off, C_z accepts i_{Q2} and V_{Cz} rises linearly. Since V_{Cz} is clamped at dc bus voltage as i_{Cz} falls to zero, the ZVS-CV switching is achieved. The negative half cycle illustrates that i_{D2} accepts the resonant current from C_z and returns energy to C_2 .

The current and voltage of resonant capacitor is shown in Fig.11. The current crest factor is 1.6 and the consumed power is 18W. Fig.12 gives the switch current and the dc bus voltage. It shows that i_{Q2} in positive ac half cycle is larger than the fourfold of i_{Q2} in negative ac half cycle. Fig.13 illustrates that the input current is sinusoidal and in phase with the input voltage. Input power factor of 0.97 is observed.

Fig.10: ZVS-CV switching

Fig.11: Current and voltage waveshapes of resonant capacitor

Fig. 12: Switch current and dc bus voltage

Fig. 13: Input current and voltage

V. Conclusions

We have completed the circuit analysis and verified the derived design rules by a 18W SRB prototype. Experimental recordings are in accordance with the analysis. This presented topology combines the power factor corrector and dc/ac inverter into single stage and hence uses only two power transistors. There are merits of low switching loss and high reliability because of inherent soft-switching and current-limiting characteristics. Nevertheless, this circuit features large peak switch currents. The switch utilization is deteriorated. Therefore, the presented circuit is appropriate for low power ballast applications.

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Appendix

$$\int_0^{\pi} \frac{\alpha^2 \sin^2 \theta}{1 - \alpha \sin \theta} d\theta,$$

$$= \int_0^{\pi} \left[-(1 + \alpha \sin \theta) + \frac{1}{1 - \alpha \sin \theta} \right] d\theta = -\pi - 2\alpha + \int_0^{\pi} \frac{1}{1 - \alpha \sin \theta} d\theta,$$

$$= -\pi - 2\alpha + \frac{2}{\sqrt{1 - \alpha^2}} \left(\frac{\pi}{2} + \tan^{-1} \frac{\alpha}{\sqrt{1 - \alpha^2}} \right).$$

where

$$\int_0^{\pi} \frac{1}{1 - \alpha \sin \theta} d\theta,$$

$$\left(\text{let } u = \tan\left(\frac{\theta}{2}\right), d\theta = \frac{2}{1 + u^2} du, \sin\left(\frac{\theta}{2}\right) = \frac{u}{\sqrt{1 + u^2}}, \cos\left(\frac{\theta}{2}\right) = \frac{1}{\sqrt{1 + u^2}} \right)$$

$$= \int_0^{\infty} \frac{2 / (1 + u^2)}{1 - \alpha [2u / (1 + u^2)]} du = \int_0^{\infty} \frac{2}{1 + u^2 - 2\alpha u} du,$$

$$= \int_0^{\infty} \frac{2}{(u - \alpha)^2 + (1 - \alpha^2)} du, \left(\text{let } u - \alpha = y, \sqrt{1 - \alpha^2} = x \right)$$

$$= \int_{-\alpha}^{\infty - \alpha} \frac{2}{y^2 + x^2} dy, \left(\text{let } y = x \tan \theta, \theta = \tan^{-1} \frac{y}{x} = \tan^{-1} \frac{y}{\sqrt{1 - \alpha^2}} \right)$$

$$= \int_{\theta_0}^{\pi/2} \frac{2x \cdot \sec^2 \theta}{x^2 (1 + \tan^2 \theta)} d\theta, \left(\theta_0 = \tan^{-1} \frac{-\alpha}{\sqrt{1 - \alpha^2}} \right)$$

$$= \frac{2}{\sqrt{1 - \alpha^2}} \left(\frac{\pi}{2} - \tan^{-1} \frac{-\alpha}{\sqrt{1 - \alpha^2}} \right).$$