Self-Oscillating Constant-Current Fluorescent Lamp Driver: Theory and Application

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***Abstract*- Here, a high frequency AC current-sourcing driver that can feed electrical loads with a constant current independent of the load resistance is presented, analyzed and tested experimentally. The power source is a self-oscillating inverter that includes a resonant network and a Current Comparing and Toggling Circuit (CCTC) that is used to generate the control signals of the inverter's power switches. The power driver can be used to ignite and operate one or several serially connected fluorescent lamps, without the need of additional resonant networks and redesign of the power stage, while still maintaining high efficiency and low output crest factor. The CCTC is realized, in the present work, by a multi-winding transformer that is designed to change the polarity of the gate drive of the power switches whenever the input signal drops to the level of the desired output current and hence, will cause the driver to behave as a current source.**

***Index Terms* — Constant AC Current sourcing, Current Comparing and Toggling Circuit (CCTC), Fluorescent lamps, Self-Oscillating electronic ballast.**

I. INTRODUCTION

Several types of loads need to be driven by a constant ac current in order to optimize their performance and maximize the system's efficiency [1] – [3]. For example, fluorescent lamps need to be driven by a source that features high output impedance to stabilize their operating point [4], and by high frequency signals to increase their light output. That is, a fluorescent lamp driver must feed the lamp with a constant current that will be independent of the lamp resistance.

One way to obtain a high impedance source is by applying an electromagnetic ballast that is based on a large bulk inductor in series with the power line voltage. A better approach would be to generate a high frequency signal by a switch mode resonant inverter (Fig. 1) since the lamp's efficacy is increased when fed by a high frequency signal. The latter solves some design problems of the first solution such as weight and a higher efficacy, although it still suffers from a number drawbacks. Firstly, there is a need for a dedicated controller and gate drivers for the high side and low side of the power switches. Secondly, in a multi-lamp fixture, there is a need to add interface (resonant) networks for each added lamp.

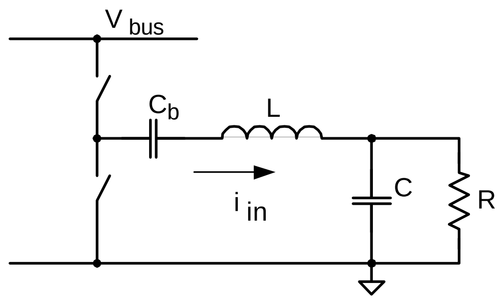

Figure 1. A half-bridge inverter feeding a resonant network

The first drawback can be solved by converting the conventional inverter topology into a self-oscillating topology [5] – [7]. This will eliminate the need for complicated and rather costly gate drives and control. However, this solution still requires a ballast per lamp, or redesign of the circuit for a specific number of lamps. It would be highly desirable if one self-oscillating ballast could drive, as is, one or more serially connected lamps. This would reduce the number of models that need to be stocked by manufacturers and may reduce production costs due to the higher production volume of the same circuit. This possibility was explored in this study.

In this work we propose a mathematical model and method for the selection of the components of the self-oscillating inverter, such that the circuit will oscillate at a desired, pre-selected, switching frequency and will behave as a current source, with the ability to feed one or more serially connected lamps (up to some practical limit) while maintaining constant current and high efficiency of the power driver.

II. SELF-OSCILLATING INVERTER

The topology (Fig. 2) includes a half-bridge inverter stage and a resonant network (L_r, C_r). The inverter drives the resonant inductor (L_r) that is placed in series with the resonant capacitor (C_r) to which the load is connected in parallel. The driver also includes a blocking capacitor (C_s) that is intended

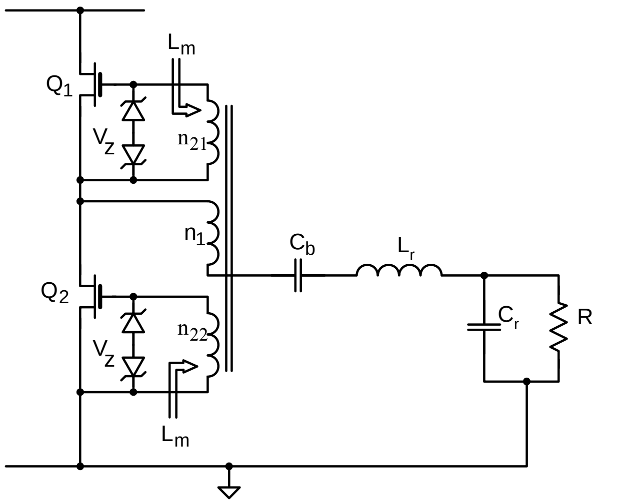

Figure. 2. Self-Oscillating resonant inverter

to pass the AC current to the load while filtering out the DC component. The capacitance of C_s is normally much larger than C_r, thus its influence on the resonant network is negligible. The load R represents the serially connected lamps.

The purpose of the capacitor C_r is to generate the high voltage needed to ignite the lamp. The peak voltage required for ignition is obtained by driving the inverter, during the start-up period, by a frequency that is close to the resonant frequency formed by the resonant network (L_r, C_r). Consequently, the voltage across C_r will accumulate rapidly, due to the high impedance of the lamps when not ignited, until the voltage level is sufficient for ignition. Once the lamps ignite, their impedance quickly drops with the system's Q-factor, which will bring the system to steady state after several cycles.

The self-oscillation operation of the driver is accomplished by using a Current Comparing and Toggling Circuit (CCTC) that, in present work, is realized by a multi-winding current transformer (n_1:n_{21}:n_{22}, Fig. 2). The primary side of the transformer (n_1, Fig. 2) is placed in series with the resonant network, while each of the secondary windings (n_{21}, n_{22}, Fig. 2) is connected to the gates of the power switches. These windings are shunted by Zener diodes in order to fix the gates' voltages [5].

The self-oscillatory commutation process is triggered by the interaction of the current accumulating in the magnetizing inductance of (n_{21}, n_{22}) and the sinusoidal resonant current that passes through n_l [5]. In particular, commutation occurs when the two are equal. It should be noted that the shape of the resonant current is sinusoidal whereas the magnetizing current increases linearly due to the fixed voltage imposed by the Zener diodes.

III. Theoretical Analysis and Mathematical Model

Considering the circuit of Fig. 2 when working around the resonant frequency, the half-bridge inverter can be simplified by first harmonic approximation to a parallel loaded resonant network that is fed by a sinusoidal signal (Fig. 3) given by

$$V_{inmax} = \frac{V_{bus}}{2}\frac{4}{\pi} = V_{bus}\frac{2}{\pi} \qquad (1)$$

where V_{bus} is the rectified input voltage (Fig. 2) and the resonant frequency ω_0 is

$$\omega_0 = \frac{1}{\sqrt{L_r C_r}} \qquad (2)$$

By applying Notron's theorem the circuit of Fig. 3 can be transform into the parallel equivalent depicted in Fig. 4. When the circuit is driven at the resonant frequency, the impedance of the parallel resonant network (Lr, Cr) is infinitely high and consequently all the current of the equivalent current source is flowing into the load Fig. 5. The output current can be derived from Figs. 4 and 5

$$I_o = \frac{V_{in}}{Z_r} \qquad (3)$$

where Z_r is the characteristic impedance (Z_r=L_r/C_r).

Hence, the output current I_o is a function of the input voltage and resonant elements only and independent of the load resistance R.

It can also be observed from (3) that when R is very large (e. g. the fluorescent lamp's resistance prior to ignition), the voltage across C_r will become very large. Hence, the driver is capable of automatically self-generating the high voltage needed for igniting fluorescent lamps.

The input impedance Z_{in} of the loaded resonant tank is

$$Z = j\omega L + \frac{R}{j\omega RC + 1} = \frac{-\omega^2 LCR + j\omega L + R}{j\omega RC + 1} \qquad (4)$$

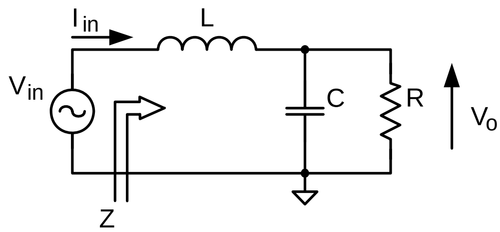

Figure. 3. Equivalent circuit, simplified by first harmonic theorem

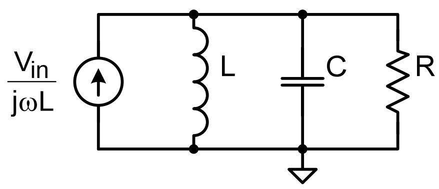

Figure. 4. Equivalent parallel network by Norton's theorem

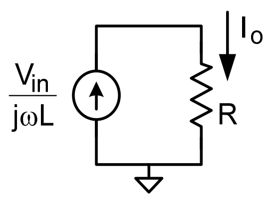

Figure. 5. Equivalent parallel network by Norton's theorem at ω=ω_0

and when $\omega=\omega_0$

$$Z = \left.\frac{j\omega L}{j\omega RC + 1}\right|_{\omega=\omega_0} = \frac{jZ_r}{j\frac{R}{Z_r} + 1} \qquad (5)$$

Consequently, the input current to the resonant section (I_{in}) will be

$$I_{in} = \frac{V_{in}}{Z} = V_{in}\frac{j\frac{R}{Z_r} + 1}{jZ_r} = \frac{jR + Z_r}{jZ_r^2}V_{in} \qquad (6)$$

and the magnitude of I_{in} will be

$$\left|I_{in}\right| = V_{in}\frac{\sqrt{R^2 + Z_r^2}}{Z_r^2} \qquad (7)$$

Based on (7), and after combining with (3) and further manipulations, the output to input voltage ratio (V_o/V_{in}) is found to be

$$V_o = I_o R \qquad (8)$$

$$V_o = \frac{V_{in}}{Z_r}R \;\Rightarrow\; \frac{V_o}{V_{in}} = \frac{R}{Z_r} = Q \qquad (9)$$

It should be noted that when the ratio of (9) is greater than unity ($R>Z_r$.), the voltage drop across the serially connected lamps will be higher than the bus voltage and the current will be sinusoidal. This is the preferred operation mode.

The phase of the input current φI_{in}, referred to the first harmonic approximated input voltage (V_{in}, Fig. 3) is found by dividing the input voltage by the characteristic impedance Z_r

$$I_{in} = \frac{V_{in}}{Z_r}\left(\frac{\frac{R}{Z_r} - j}{1}\right) = \frac{V_{in}}{Z_r^2}R\left(1 - j\frac{Z_r}{R}\right) \qquad (10)$$

from which

$$\varphi I_{in} = tg^{-1}\left(-\frac{Z_r}{R}\right) \qquad (11)$$

For the condition that the output voltage is greater than the input voltage (needed for normal operation), φI_{in} is small and since

$$\varphi I_{in} \approx tg\,\varphi I_{in}\,, \qquad (12)$$

one finds

$$\varphi I_{in} \approx -\frac{Z_r}{R} \qquad (13)$$

Due to the first harmonic approximation used in the analysis, the angle of the approximated input voltage (V_{in}, Fig. 3) is essentially the angle of the actual square-wave generated by the power switches (Fig. 2). Hence, the angle φ obtained in (13) is the phase shift between the inverter's square-wave and the input current. It should b noticed that the fact that φI_{in} is negative (lagging current), ZVS for the half bridge switched will be achieved. The magnitude of the input current (I_x) at the commutation time (Fig. 6) is found as follows:

The peak input current is

$$I_{in\,max} = V_{in\,max}\frac{\sqrt{R^2 + Z_r^2}}{Z_r^2} \qquad (14)$$

for

$$V_o > V_{in} \;\Rightarrow\; Z_r^2 < R^2 \qquad (15)$$

Hence, Eq. 13 can be approximated to

$$I_{in\,max} = V_{in\,max}\frac{R}{Z_r^2} \qquad (16)$$

and I_x is found to be

$$I_x = I_{in\,max}\sin\varphi \approx V_{in\,max}\frac{R}{Z_r^2}\left(-\frac{Z_r}{R}\right) \approx V_{in\,max}\frac{1}{Z_r} \qquad (17)$$

The last equation and (3) imply that the I_x is equal to the desired maximum output current I_{omax}

$$I_x = \frac{V_{in\,max}}{Z_r} = I_{o\,max} \qquad (18)$$

Hence, a load independent operation can be achieved by forcing the commutation of the input square wave to occur when the instantaneous input current drops to the level of the desired maximum output current (Fig. 6).

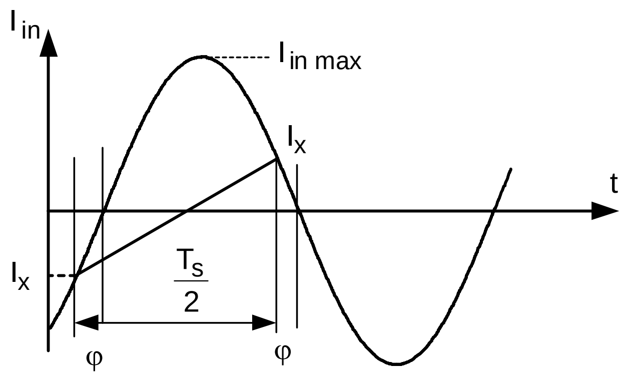

Figure. 6. The desired commutation intersection point, referred to the approximated input current.

Thus, the criterion for constant current sourcing with respect to the circuit parameters of Fig. 2 can be found by equating I_o to the value of the magnetization current of L_m at the commutation instance (I_x)

$$I_x = \frac{V_Z T_S}{L_m 4} n = I_{o\,max} \quad (19)$$

where V_z is the clamping Zener diode voltage (Fig. 2), L_m is the magnetizing inductance of the current transformer as measured from the secondary windings (n_{21}=n_{22}), T_s is the period of the oscillations frequency (f_s) and (n=n_{21}/n_1=n_{22}/n_1).

Hence, to facilitate a constant load current I_o, one needs to choose Lm according to

$$L_m = \frac{V_Z T_S n}{4 I_{o\,max}} \quad (20)$$

IV. Components Selection Procedure

Based on the mathematical description of the system given in the previous section, the following design procedure is suggested for proper operation of the self-oscillating, constant-current inverter. Considering first the given set of specifications: rectified input voltage ($V_{bus} = V_{in}\sqrt{2}$), output current (I_{o_rms}), switching frequency (f_s), clamping voltage of the gate drives (V_z) and output power, or more specifically, load resistance (R).

A. *Resonant components selection:*

Calculate the rms input voltage from (1)

$$V_{in\_rms} = V_{bus}\frac{\sqrt{2}}{\pi} \quad (21)$$

Extract the value of the desired characteristic impedance Z_r by manipulating (3)

$$Z_r = \sqrt{\frac{L}{C}} = \frac{V_{in\_rms}}{I_{out}} \quad (22)$$

Given the frequency of oscillations, calculate the value of the resonant capacitor (C_r) by

$$Z_r f_S = \frac{1}{2\pi C_r} \quad \Rightarrow \quad C_r = \frac{1}{2\pi Z_r f_S} \quad (23)$$

Then the resonant inductance is given by

$$f_S{}^2 = \frac{1}{4\pi^2 L_r C_r} \quad \Rightarrow \quad L_r = \frac{1}{f_S{}^2\, 4\pi^2 C_r} \quad (24)$$

Verify your selection, for proper operation the resonant network must be underdamped (Q>1) as stated in (9). Calculate the output voltage by

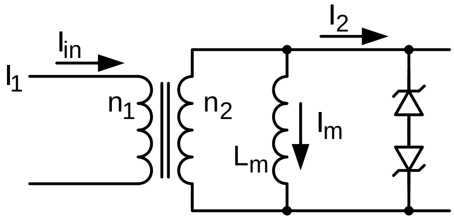

Figure. 7. Current transformer, simplified circuit.

$$V_{out} = I_{out} R \quad (25)$$

$$Q = \frac{R}{Z_r} = \frac{V_{out}}{V_{in}} > 1 \quad (26)$$

If (26) does not hold, the initial specifications need to be changed.

B. *Current transformer design for constant output current*

1) *Inductance value*

Determine the number of turns of the transformer by the following steps. First, calculate the input current by

$$\frac{V_{out}}{V_{in}} > 1 \rightarrow I_{in} = V_{in}\frac{R}{Z_r{}^2} \quad (27)$$

Calculate the current needed at the secondary output by

$$I_2 = I_z + I_{mag} \approx 1.2 I_z \quad (28)$$

where I_z is the Zener diodes current at clamping mode (can be drawn from the datasheet) and I_{mag} is the magnetizing current of the transformer (Fig. 7), a factor of 0.2 of the Zener diode is a good starting point.

The number of turns (n) can now be derived by

$$n = \frac{I_{in}}{I_2} \quad (29)$$

The magnetizing inductance (secondary side) needed for constant load current, as found in (20), is

$$L_m = \frac{V_Z T_S n}{4 I_{o\,max}} \quad (30)$$

2) *Core selection*

Calculate the core area (A_e) and the window area (A_w) by

$$A_e = \frac{L_m \cdot I_{out\_max}}{B_{max} \cdot n_2} \quad (31)$$

$$A_W = \frac{I_{in} \cdot n_1 + I_2 \cdot n_2}{J \cdot K} \quad (32)$$

where B_{max} [Tesla] is the maximal flux density allowed, J is the current density [A/m^2] and K is the fill factor (constant smaller than 1).

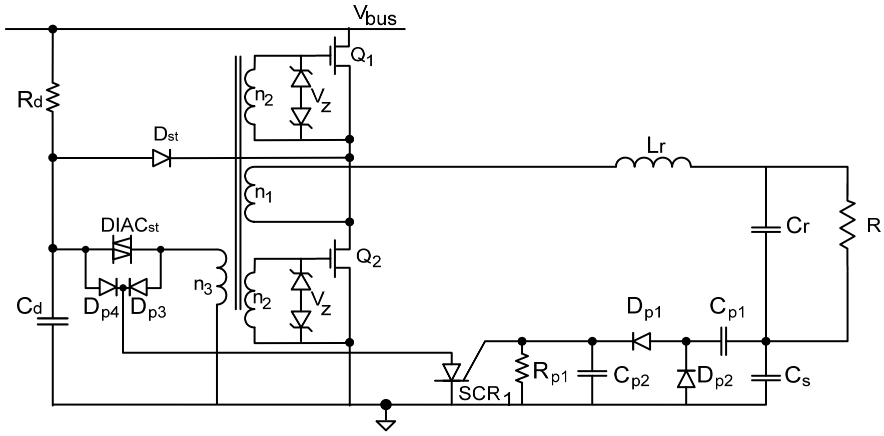

Figure. 8. Experimental setup

The area product needed for magnetic element to comply with design specifications (inductance value, current, etc.) is given by

$$A_{p\_des} \geq A_e \cdot A_w \tag{33}$$

C. Resonant inductor core selection:

Given the design specifications, the area product needed for the magnetic element to be used as the resonant inductor is

$$A_p = A_e \cdot A_w \geq \frac{L \cdot \sqrt{2} \cdot {I_{out}}^2}{J \cdot K \cdot B_{max}} \tag{34}$$

V. Experimental Verification

An experimental unit of the self-oscillating inverter was designed and built, following the guidelines described in previous chapters. The experimental circuit included some additional circuitry for start-up lamp ignition and over-voltage protection (Fig. 8).

At start-up, a resistor (R_d) charges the capacitor (C_d) until the trigger voltage of the Diac (Fig. 8) is reached. At this point, the Diac breaks down and a pulse is fed to the gates of (Q_1, Q_2) via an auxiliary winding (n_3, Fig. 8) to start oscillations. If the triggering fails, the Diac will keep generating pulses until the inverter will self-oscillate. Once the lamp ignites, the diode (D_{st}, Fig. 8) discharges C_d to prevent the accumulation voltage of further ignition pulses.

Prior to ignition, or if the lamp is broken or not inserted in the circuit, the inverter will develop high current through the resonant elements and potentially high output voltage that may damage the circuit. For this purpose, the voltage of the blocking capacitor C_s, which is proportional to the resonant current is sensed, rectified, filtered and applied to thyristor's gate (SCR_1, Fig. 8), so that when it is turned on, it will block oscillation.

By the addition of pull down diodes (D_{p3}, D_{p4}, Fig. 8), once fired, the SCR will hold it's on state by the charging resistor (R_d). The circuit will reset only when disconnected from power.

The experimental unit was design for US standard grid voltage of 115VAC (V_{bus}=150VDC) and operation frequency of 100KHz. The measurements were obtained with loads of 300Ω, 600Ω, 1000Ω and 1600Ω to emulate 1 to 4 serially connected lamps.

Figs. 9 to 12 show the half bridge current and the load current for four power levels. It can be observed that the output current remain constant for all power levels, which indicates on the current source nature of the unit and supports the design procedure.

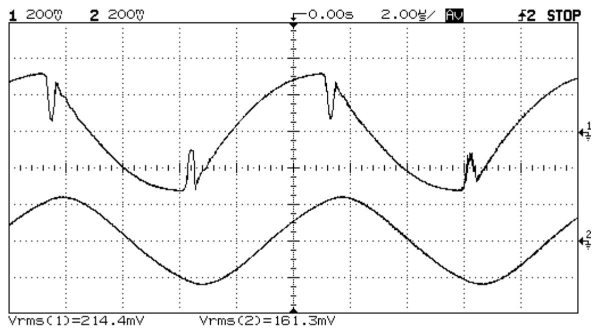

Figure. 9. Experimental results, constant current when loaded by 300Ω : top: I_{out} (0.2A/div); bottom I_{lamp} (0.2A/div); Horizontal scale: 2µS/div

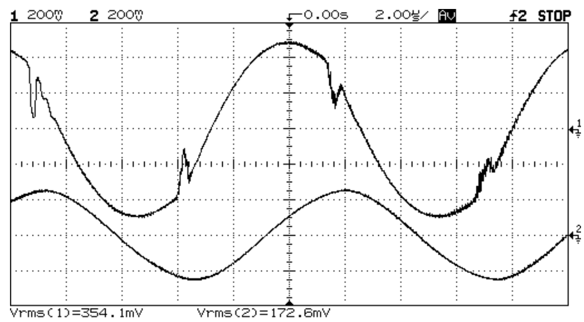

Figure. 10. Experimental results, constant current when loaded by 600Ω : top: I_{out} (0.2A/div); bottom I_{lamp} (0.2A/div); Horizontal scale: 2µS/div

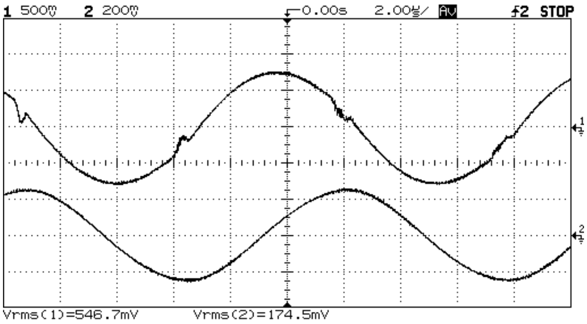

Figure. 11. Experimental results, constant current when loaded by 1000Ω : top: I_{out} (0.2A/div); bottom I_{lamp} (0.2A/div); Horizontal scale: 2μS/div

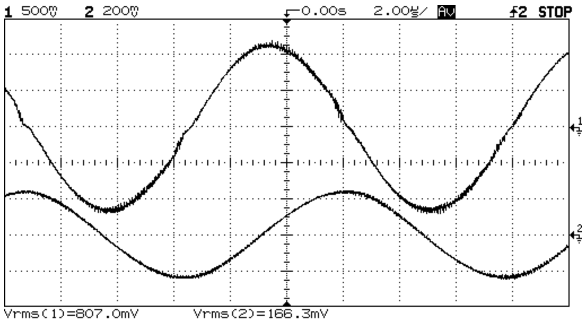

Figure. 12. Experimental results, constant current when loaded by 1600Ω : top: I_{out} (0.2A/div); bottom I_{lamp} (0.2A/div); Horizontal scale: 2μS/div

A summary of this set of measurements is given in Table I and depicted in Fig. 13 for better visualization. The inherent ZVS quality of the power driver, when properly designed, is depicted in Figs. 14 and 15, which validated the design of the current transformer in charge of the commutation process. The over-voltage protection operation is demonstrated in Fig. 16 whereas the system was ignited under 'no load' conditions, oscillations ceased after 10mS.

VI. Discussion and Conclusions

The proposed design of a self-oscillating inverter shows the capability of driving fluorescent lamps by high frequency constant AC current that is independent of the load (from 1 to at least 4 lamps). The power source drives the lamps safely and when properly designed, following the guidelines detailed in this work, the power source operates under soft-switching conditions (ZVS in turn on and turn off).

The load current that was measured in the experiment was found to be almost sinusoidal. This implies that the output current will have a very good crest factor, an attribute that contributes to long lamp life.

The theoretical study and the mathematical model derived in this work, which was supported by experimental measurements, cover the aspects of self-oscillating and current sourcing. The contribution of this research is the straightforward approach for practical design and use of the proposed modification of the self-oscillating ballast that converts it to a constant current source. The proposed ballast is covered by a US patent [8].

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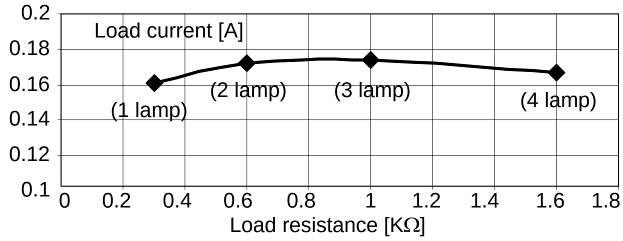

Figure. 13. Experimental results, Load current as a function of the load resistance. Constant-current driver operation.

Table I. Summary or measurements. Measured lamp and resonant current as a function of the load

Load [Ω]	Equ. no. of lamps	I_{Lamp} [Arms]	IL_r [A]
1600Ω	4	0.167A	0.8A
1000Ω	3	0.174A	0.546A
600Ω	2	0.172A	0.354A
300Ω	1	0.161A	.21A

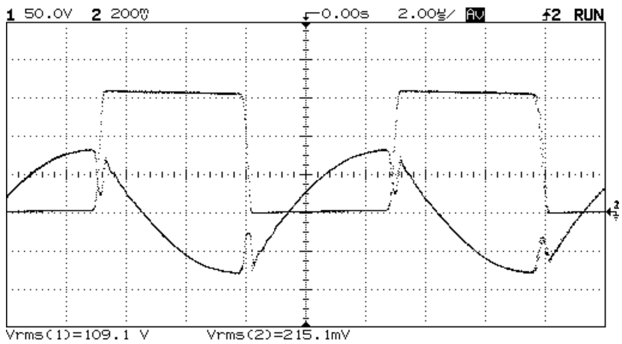

Figure. 14. Experimental results, ZVS operation when loaded by 300Ω : top: drain voltage of Q_2 (50V/div); bottom I_{out} (0.2A/div); Horizontal scale: 2μS/div

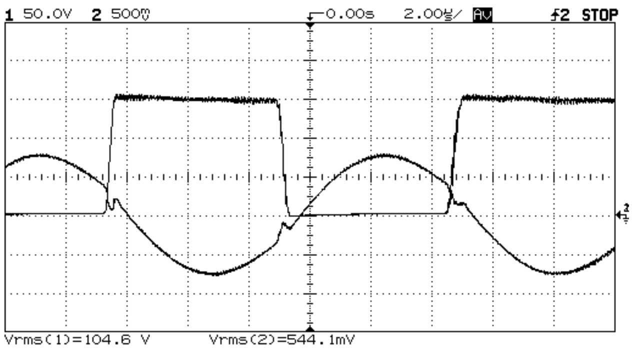

Fig. 15. Experimental results, ZVS operation when loaded by 1000Ω : top: drain voltage of Q_2 (50V/div); bottom I_{out} (0.5A/div); Horizontal scale: 2μS/div

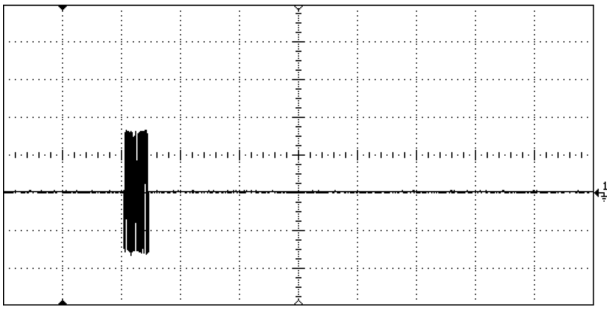

Figure 16: Experimental results. Over voltage protection operation under no load conditions. Gate voltage of Q_2. Horizontal scale: 20ms/div.

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