PWM Control of Electronic Ballast for High-pressure Na Lamps in Comparison to Fluorescent Lamps. Introduction to Quasi-optimum Control

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Abstract: **The design of electronic ballast to drive high and low-pressure lamps should fulfill both the high-voltage necessities for the start-up and the current limiting requirements in steady-state operation. The full-bridge parallel resonant converter reduces the current stress compared to the half-bridge converter, so it is presented as a solution to drive lamps over 100W. The analysis of the full-bridge control based on the pulsewidth modulation is presented for HPS and fluorescent lamps. As the inverter input-voltage is usually quite high compared to the lamp voltage requirements in steady-state operation, the quality factor of the resonant tank is low, resulting in an appreciable level of reactive current. The PWM dimming control at fixed frequency is proposed when the total current level is acceptable. In this case, while maintaining the transistors' turn-on at zero voltage, the PWM control improves the converter efficiency by reducing the reactive current. When the reactive current causes an important efficiency reduction, a quasi-optimum control method is proposed, that consists of fixing, at each control point, the frequency vs. duty-cycle relation that minimizes the reactive current while maintaining soft turn-on.**

I. INTRODUCTION

The efficiency of high-pressure sodium (HPS) lamps, over 100lm/W, makes them suitable for use in illumination systems for outdoor spaces. The low-power version, up to 100W, and the widely used fluorescent tubes are locates in spaces such as industrial plants and warehouses. Because of their light spectra, these lamps achieve good color identification.

It is known that fluorescent tubes and high-pressure discharge lamps, when powered by a high frequency source behave resistive [1][2], considering any particular steady-state operation point. The equivalent resistance of the HPS lamps is lower than in fluorescent tubes, and has no dependence on the power delivered to the lamp [3]. The major drawback, when using high-frequency converters to drive HPS lamps, is the acoustic resonance phenomenon [2][4] that may produce an unstable operation, i.e. the arc length, voltage, and power change erratically, leading to light flickering and even light extinction. The resonant frequencies depend on the arc tube geometry, temperature, pressure, and gas composition. To avoid the acoustic resonance phenomenon, the switching frequency of the ballast is fixed in a range of frequency free of resonance.

The full-bridge parallel resonant converter topology (PRC) is a good option when driving high-power HPS lamps or a set of various fluorescent tubes because it can achieve high start-up voltage without excessively increasing the current. However, in steady-state operation, the reactive component of the current through the converter is usually large due to the small quality factor of the resonant tank required to fix the nominal voltage drop. The control based on the pulsewidth modulation (PWM) has the advantage of reducing the power delivered to the lamp while the reactive component is also reduced [9]. The efficiency of the converter is then clearly improved.

The paper is structured as follows: In Section II, the PWM controlled PRC inverter is analyzed to give the most significant design expressions and the control parameters. Section III presents the experimental results obtained with a 70W OSRAM NAV70E HPS lamp under PWM control compared to a 40W Sylvania F40 fluorescent tube. The design and experimental results of an electronic ballast for a 150W OSRAM HPS lamp, using a quasi-optimum control that minimizes the reactive current in the resonant circuit, are also given in this section, finalizing with conclusions.

II. RESONANT INVERTER DESIGN AND PWM CONTROL

The full-bridge PRC circuit of fig. 1 is analyzed considering the fundamental component of the voltage and current of the resonant tank (see fig. 2). R_{lamp} represents the equivalent HPS lamp resistance. At this point, we consider neither the transistors on-resistance, r_{dson}, nor other circuit losses.

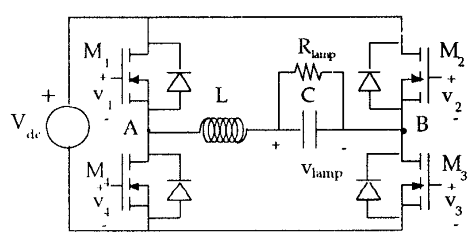

Fig.1 Full-bridge PRC loaded with the equivalent HPS lamp resistance.

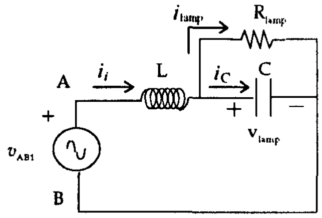

Fig.2 Fundamental approach of the lamp-loaded PRC.

The resonant tank loaded with R_{lamp} is described by the natural frequency ω_o, the characteristic impedance Z_o and the quality factor Q_P [6][7].

As R_{lamp} is nearly an open circuit before the start-up, a large voltage drop rises on the lamp terminals when the switching frequency ω is close to ω_o, achieving the gas ionization [8]. During the start-up the value of R_{lamp} decays, and so does the output voltage until reaching its nominal V_{lamp} value. Contrary to what happens when driving fluorescent tubes, the voltage drop in HPS lamps depends on the converter design.

If the switching frequency is constant, $\omega=\omega_o$, and the converter is controlled by the PWM method, then equation (1) gives the amplitude of the first harmonic of the voltage V_{AB} (see fig.1) as a function of the duty cycle, D.

$$\hat{V}_{AB1} = \frac{4V_{dc}}{\pi}\cos(1-D)\frac{\pi}{2} \tag{1}$$

As is reported in [3], R_{lamp} in steady-state conditions is nearly constant and presents no dependence on the power delivered to the lamp. Assuming this behavior, the maximum voltage applied to the HPS lamp when the switching frequency is ω_o is given by:

$$\hat{V}_{lamp} = \frac{4\cdot V_{dc}}{\pi}\cdot Q_p\cos(1-D)\frac{\pi}{2} \tag{2}$$

Since Q_P remains constant, the lamp voltage is controlled by parameter D.

The peak value of the resonant current, $\hat{I}_i$, also depends on the duty cycle

$$\hat{I}_i = \frac{4V_{dc}\sqrt{1+Q_p}}{Z_o\pi}\cos(1-D)\frac{\pi}{2} \tag{3}$$

On the other hand, fluorescent tubes maintain the voltage drop approximately constant after the start-up, even under variations of the power delivered to the lamp [5]. Thus, the power control produces changes in Q_P that in (3) should be replaced by:

$$Q_p(D) = \frac{Q_o}{\cos(1-D)\frac{\pi}{2}} \tag{4}$$

where Q_o is the quality factor for D=1 (nominal power). Note that Q_P(D) in equation (4) makes V_{lamp} of equation (2) constant. An advantage of the PWM control method is that it makes it possible to maintain the switching frequency constant, simplifying the design of the resonant tank and the line filter. Expression (3) denotes the current dependence on Q_P and the duty cycle, so that the circuit is protected under short circuit (low Q_P) but needs special attention under open circuit operation (Q_P).

The lamp current behaves as a D-controlled-current-source when the switching frequency is ω_o, so that the load is permanently protected from overcurrent.

$$\hat{I}_{lamp} = \frac{4V_{dc}}{Z_o\pi}\cos(1-D)\frac{\pi}{2} \tag{5}$$

As expression (5) has no dependence on the load, it is valid for both HPS and fluorescent lamps.

By knowing the lamp voltage and current, the power delivered to the lamp can be directly calculated. The power expression for HPS lamp is:

$$P_{lamp} = \frac{8\cdot V_{dc}^2 Q_p}{Z_o\pi^2}\cos^2(1-D)\frac{\pi}{2} \tag{6}$$

and for fluorescent lamps:

$$P_{lamp} = \frac{8\cdot V_{dc}^2 Q_o}{Z_o\pi^2}\cos(1-D)\frac{\pi}{2} \tag{7}$$

Both control functions are depicted in figs. 3 and 4, the output power being normalized with respect the maximum value.

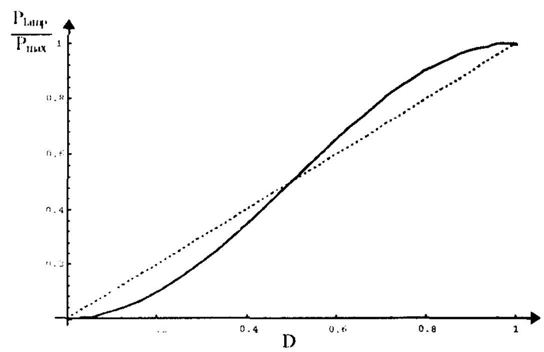

Fig.3 Power delivered to HPS lamps as a function of the duty cycle D.

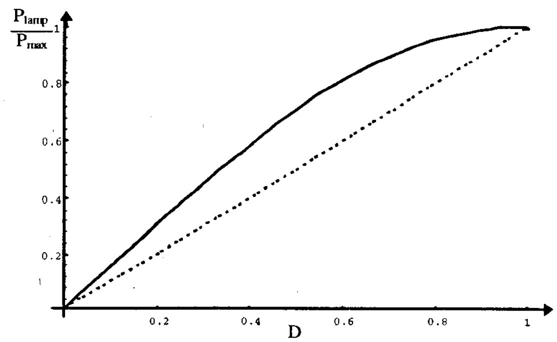

Fig.4 Power delivered to fluorescent lamps as a function of the duty cycle D.

The dotted line corresponds to the normalized function P_{lamp} equal to D, which is closer to the PWM regulation for HPS lamps. The variation range of the control parameter D is limited by the condition of maintaining the MOSFETs turn-on at zero voltage (ZVS).

This condition is guaranteed for large values of D since $Q_p<1$ after the lamp starting. Fig. 5 shows the theoretical waveform of the control signals and the inverter section.

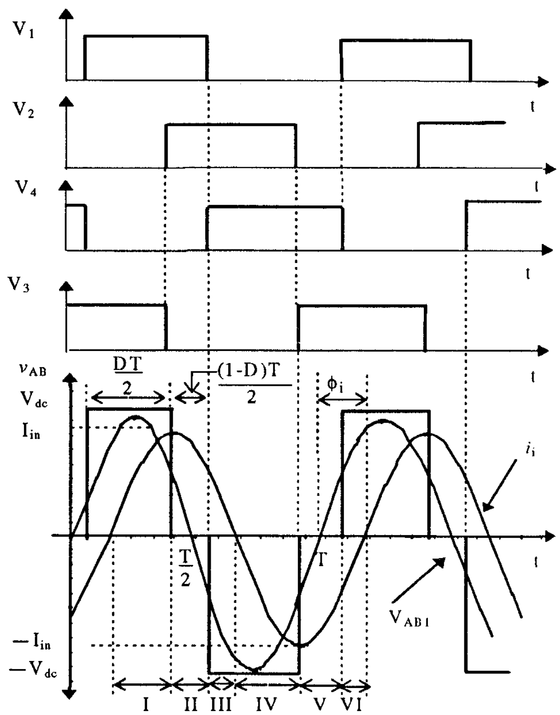

Fig.5 Inverter control signals. Input voltage and current in the resonant circuit.

In fig. 5, it is observed that as the pulse-width D is reduced, the intervals t_{III} and t_{VI}, where the power flow is negative, are also reduced and the efficiency of the power transferred to the load improves. The equal values t_{III} and will be denoted as t_r.

Each switching situation of fig.5 is presented in fig. 6 where the current path through the converter is depicted in each stage.

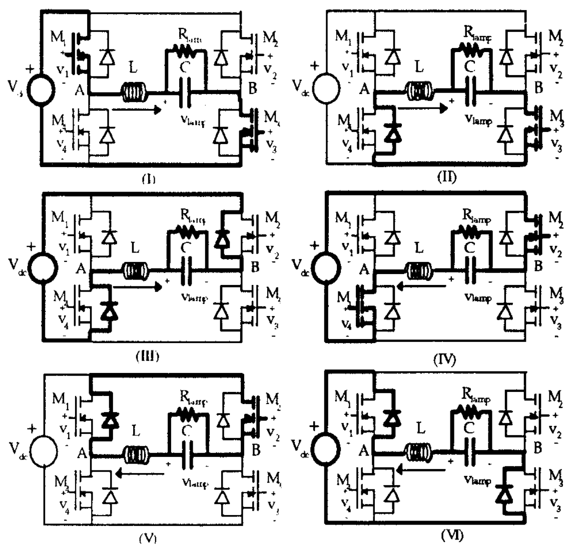

Fig.6 Sequence of theinverter devices in a switching period.

Since Q_p is constant under the dimming control when driving HPS lamps, the current i_i vs. v_{AB1} phase lag depends only on the switching frequency.

$$\phi_i = -tan^{-1}\left\{Q_p\left(\frac{\omega}{\omega_o}\right)\left[\left(\frac{\omega}{\omega_o}\right)^2+\frac{1}{Q_p^2}-1\right]\right\} \tag{8}$$

The angle, ϕ_i, determines the minimum value of D in which we still find ZVS (D_{min}). To obtain the value of D_{min}, the value t_r is evaluated by the following expression:

$$t_r = \frac{\phi_i}{2\pi}\cdot T - \frac{(1-D)T}{4} \tag{9}$$

The value D_{min} is then calculated by equating expression (9) to zero:

$$D_{min} = 1-\frac{2\cdot\phi_i}{\pi} \tag{10}$$

Expression (11) gives the amount of energy E_r returning in each cycle to the supply V_{dc} as a function of D and ϕ_i, valid only for values of $D>D_{min}$.

$$E_r = \frac{V_{dc}\hat{I}}{2\cdot\pi}\cdot T\left[\cos\left((1-D)\frac{\pi}{2}-\phi_i\right)-1\right] \tag{11}$$

The graphic of fig. 7 shows how the reactive energy is reduced when the duty cycle is reduced for different given values of the phase lag ϕ_i.

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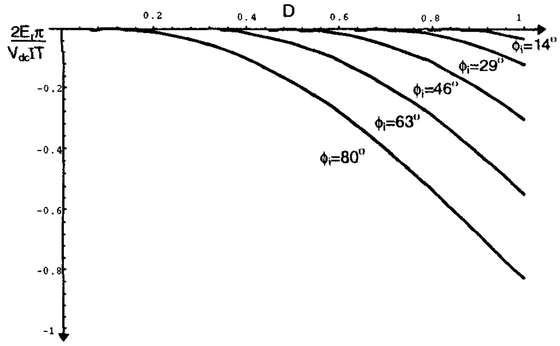

Fig. 7 Reactive energy vs. duty cycle for a given value of phase.

The power of commercial HPS lamps varies from 35W to 1000W, the nominal current being over $1A_{rms}$ for lamps that receive more than 100 W. Such current levels make it necessary to take into account the transistor-on-losses. The reactive energy increments the input current to the resonant circuit that increases the transistor on losses and reduce the efficiency of the resonant inverter.
The graph of fig. 7 leads to a generalization of the dimming control in order to minimize E_r. By substituting expression (8) into (10), we obtain D_{min} as a function of the frequency, for a given value of Q_p.

$$D_{min} = 1 - \frac{2 \cdot tan^{-1}\left\{ Q_p\left(\frac{\omega}{\omega_o}\right)\left[\left(\frac{\omega}{\omega_o}\right)^2 + \frac{1}{Q_p^2} - 1\right]\right\}}{\pi} \tag{12}$$

The relation between the switching frequency, ω, and D_{min} is a control characteristic of the electronic ballast that eliminates the reactive power and maintains ZVS. Fig. 8 shows the optimum ω vs. D_{min} relation given in (12) using Q_p as a parameter.

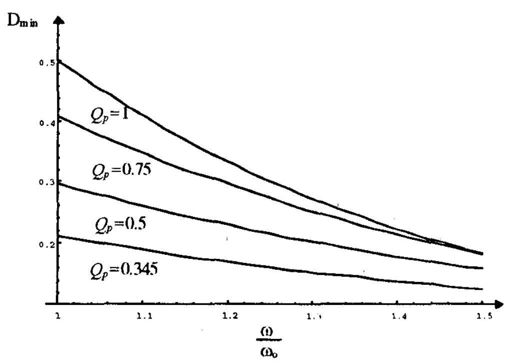

Fig.8 Optimum control function of the parallel resonant converter to eliminate the reactive power and maintain ZVS

If the switching frequency is used as a control parameter then the power delivered to a HPS lamp as a function of ω, D and Q_p varies according to the expression (13).

$$P_{lamp} = \frac{8V_{dc}^2 \cdot \cos^2\left[(1-D)\frac{\pi}{2}\right] \cdot Q_P}{\pi^2 \cdot Z_o \cdot \left\{ Q_P^2\left(1 - \left(\frac{\omega}{\omega_o}\right)^2\right)^2 + \left(\frac{\omega}{\omega_o}\right)^2 \right\}} \tag{13}$$

By introducing the optimum relation ω vs. D given in fig. 8 into expression (13), the power control function that minimizes the reactive power is found.
Following this control strategy, the first design step is to calculate the value of the duty cycle in nominal conditions, D_o. An expression for D_o is obtained from (8) and (10) considering the switching frequency $\omega=\omega_o$.

$$D_o = 1 - \frac{2 \cdot tan^{-1}\left\{ \frac{4V_{dc}}{V_{lamp}\pi} \cos\left(1 - D_o\right)\frac{\pi}{2} \right\}}{\pi} \tag{14}$$

Knowing the values of D_o and V_{lamp} the quality factor, Q_p, is calculated from (2). Using Q_p, R_{lamp} and ω_o, the reactive components L and C of the resonant tank are calculated.
Once Q_p is known, the control circuit must be able to adjust D_{min} for each switching frequency following expression (12).

III. DESIGN AND EXPERIMENTAL RESULTS

A. PWM Control at Fixed Switching Frequency.

A full-bridge PRC, PWM controlled at fixed frequency, has been designed to drive a 70 W HPS lamp. The design input data are the nominal lamp voltage and current at full output power: $V_{lamp}=80V_{rms}$, $I_{lamp}=0.875A_{rms}$, so $R_{lamp}=91\Omega$. The inverter input voltage is $V_{dc}=300V$. Natural and switching frequency have been fixed at $f_o=50kHz$. According to the previous data and considering that D=1 leads to the nominal output power, expression (2) gives $Q_p=0.2961$. By knowing Q_p, R_{lamp} and f_o, the resonant tank is calculated; the characteristic impedance is $Z_o=307\Omega$, L=1mH and C=10.36nF. The MOSFETs chosen for the inverter section are the IRF840 controlled by the driver IR2111. A digital circuit has been designed to generate eight different values of D so that ΔD=0,125.

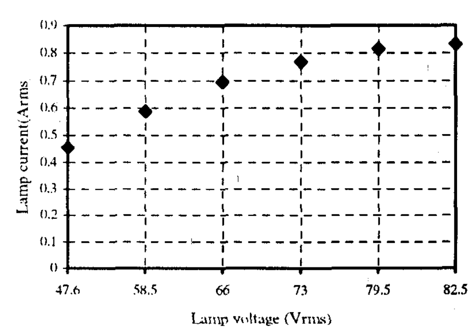

Fig.9. RMS voltage and current for the HPS lamp under PWM control

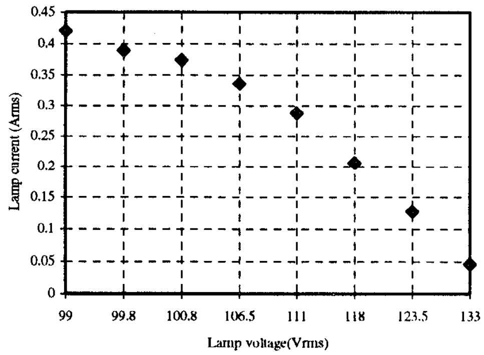

Fig.10. RMS voltage and current for the fluorescent lamp under PWM control.

Fig. 9 shows six steady-state operation-points, I_{lamp} vs. V_{lamp}, obtained by the PWM control at constant switching frequency, while maintaining ZVS for the 70W HPS lamp. The experiment confirms that the value of the equivalent resistance, R_{lamp}, is constant at any value of D. The result of the same experiment is shown in fig. 10 for a 40W fluorescent lamp Sylvania F40, where the increase of the equivalent resistance with D is observed.
The fit by a linear function of the points depicted in fig. 9, $I_{lamp}=0.011018V_{lamp}+0.05525$, gives a correlation coefficient ρ=0.99227. According to this function the value of the equivalent resistance is $R_{lamp}=90.75\Omega$.
In fig. 11, the measurements on the power delivered to the HPS lamp are compared to the graph of equation (6). The resulting output power has been normalized with respect to its nominal value, P_{max}=70W. The linear function that fit the measured power ($P_N=1.069D-0.0564$) has a correlation coefficient ρ=0.9889, close to 1, thus showing very good linearity control. The same results are presented in fig. 12 for the fluorescent tube, where equation (7) is used to compare with the experimental measurements and the power is normalized with respect to P_{max}=40W.

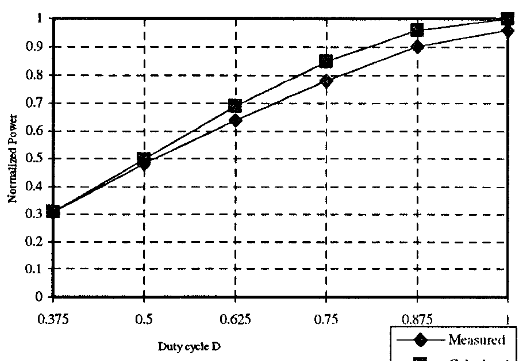

Fig.11. Measured and calculated values of the normalized power delivered to the HPS lamp.

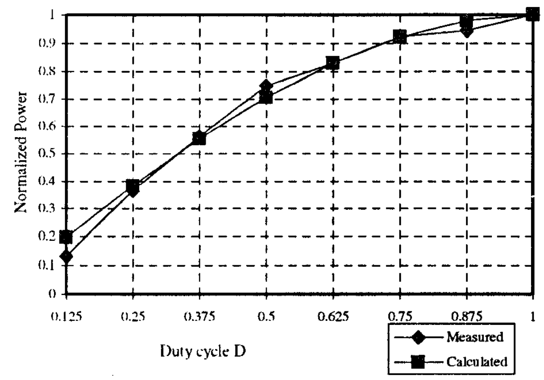

Fig.12. Measured and calculated values of the normalized power delivered to the fluorescent lamp.

B. Design of the Electronic Ballast Minimizing the Reactive Energy

For the practical implementation of this control strategy, an increase of D over D_{min} has been taken into account to guarantee a safe margin that maintains ZVS mode. For this reason, this control method is called quasi-optimum.

$$D_{pract} = D_{min} + D_s \tag{15}$$

D_s is estimated by considering the lamp aging [2] and the tolerance of the ballast components.
The HPS lamp selected was the OSRAM NAV150E. The electric characteristics are $V_{lamp}=83V_{rms}$, $I_{lamp}=1.8A_{rms}$ and $R_{lamp}=45\Omega$. The input inverter voltage is V_{dc}=300V.
Substituting V_{lamp} into (14), an iterative process obtains D_o. In this case D_o=0.343. Knowing D_o, from (2) the quality factor is calculated, Q_P=0.598. The nominal switching frequency f_o is fixed at 50kHz and the reactive components are : L=239μH and C=42.36nF. The control method is implemented by a specific digital circuit that allows the setting of four power levels in the lamp. Using expression (12), the D_{min} corresponding to each switching frequency is calculated and given in TABLE I

TABLE I

i	f_i	D_{min}	D_{pract}
0	50kHz	0.343	0.37
1	55.5kHz	0.294	0.33
2	62.5kHz	0.241	0.31
3	71.4kHz	0.18	0.25

The experimental waveforms obtained at full power and at 30% of full power are shown in figs. 13 and 14. Note that, in both cases, the level of reactive energy in the resonant circuit is very low.

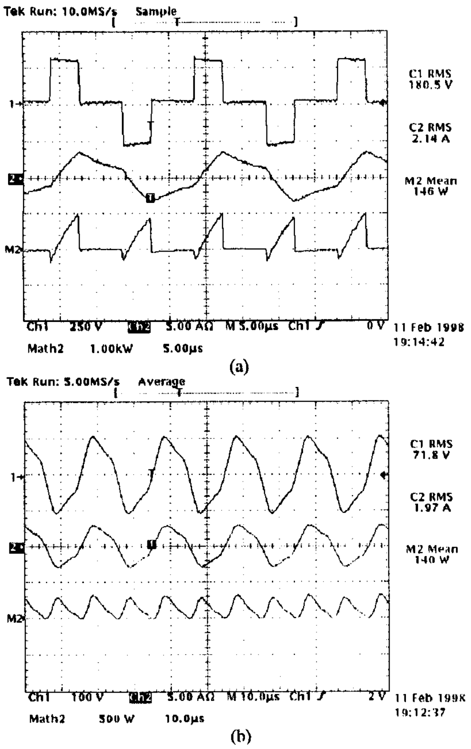

Fig.13 a) Input voltage, current and instant power at resonant circuit for nominal conditions (full power). b) HPS lamp voltage, current and power under nominal conditions.

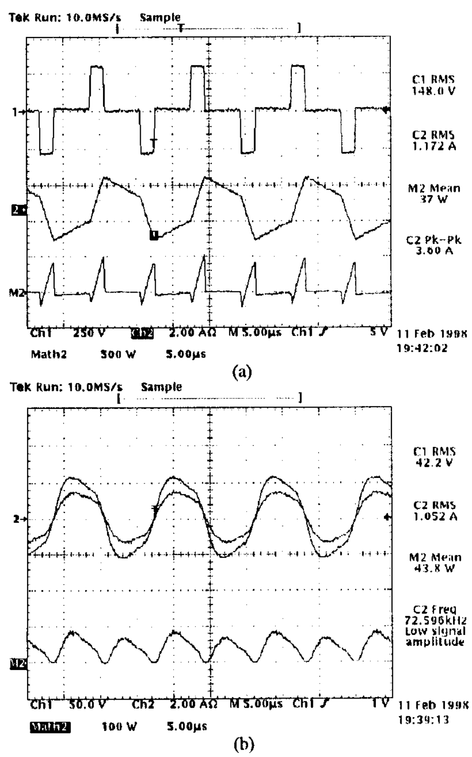

Fig.14 a) Inverter section output voltage, current and instant power at 30% of full power. b) HPS lamp voltage, current and power at 30% of full power.

IV. CONCLUSIONS

The criteria for the electronic ballast design based on the full-bridge PRC topology have been presented. This topology is a good option to drive discharge lamps over 100W and to implement control strategies based on pulsewidth modulation techniques. The characteristics of the PWM control method at constant switching frequency have been given for HPS lamps and fluorescent tubes. The main advantages of this method are the reduction of the reactive current in the converter as long as D is reduced, the simple implementation and the fact that the switching frequency is constant. Since HPS lamps make the reactive current in the converter increase excessively, the proposed quasi-optimum control technique has been verified to be an appropriate control method to minimize the converter losses.

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