Full-range high-frequency high-efficiency class D/sup 2/ resonant power supplyProceedings of IECON'94 - 20th Annual Conference of

Off-Line Full-Range High-Frequency High-Efficiency Class D^2 Resonant Power Supply

M. Bartoli*, A. Reatti*, Member, IEEE, and M. K. Kazimierczuk**, Senior Member, IEEE

*University of Florence, Department of Electronic Engineering, Via di S. Marta, 3, 50139 Florence - Italy
Phone: (39)(55)4796-389 and Fax: (39)(55)494569; E - Mail: CIRCUITI@VM.IDG.CNR.FI

**Wright State University, Department of Electrical Engineering, Dayton, Ohio 45435, USA
Phone: (513)873-5059 and Fax: (513)873-5009; E - Mail: MKAZIM@VALHALLA.CS.WRIGHT. EDU

Abstract **– A full–range off–line power supply based on a class D^2 dc-dc parallel resonant converter is presented in this paper. The class D inverter is operated above the resonant frequency to produce a zero voltage turn-on of the MOSFETs. Turn-off losses are drastically reduced by using only one capacitor connected in parallel to one of the two MOSFETs of the inverter. As a result, switching losses are nearly zero and a full-load, low line high-efficiency h = 86% is achieved at an operating frequency of about 500 kHz including the front-end rectifier. The output voltage V_o = 12 V is regulated over the entire load range, that is, for an output current I_o ranging from 0.25 A to 2.5 A and for both American and European standard input voltages in a narrow frequency range, from f_{min} = 290 kHz to f_{max} = 490 kHz. Other advantages of the power supply are its inherently short-circuit protected operation, the low di/dt at the MOSFET turn-off that allows the parasitic MOSFET diodes to be used, and the low conduction losses in ESR of the rectifier filter capacitor.**

I. Introduction

In last years, designers have been trying to increase the converter switching frequency to achieve volume and weight reduction. This results in small transformers, filter chokes, and filter capacitors. Resonant dc-dc converters represent the most promising solution to the problem of high power-density power supplies and noise level reduction. They are constituted of a resonant inverter and a rectifier coupled with a transformer that provides the galvanic insulation between the high voltage input and the low voltage output and the proper voltage transfer function by means of its turns ratio n. The reduction of switching losses makes them suitable for a high-frequency operation. Main limitations of some resonant converters are high peak currents in the inverter section and high reverse voltage across power transistors; e.g. in the class E inverter the maximum reverse voltage may be four times higher than the dc input voltage V_{DD} [1-3]. Class D resonant inverters have a low voltage stress on power transistors, equal to V_{DD}. For this reason, they can be used for applications in off-line power supplies, where the rectified dc input voltage is high, e.g., $V_{DD} = \sqrt{2} \times 220 \times 1.2 = 370\ V$ in Europe. Moreover, low cost power transistors with a low R_{ON} and low parasitic capacitances can be used and both conduction and switching losses are reduced.

Other papers [4-6] show resonant dc-dc converters operated below the resonant frequency. However, this operation requires fast external antiparallel diodes in the inverter circuit. Moreover, only turn-off losses are eliminated. The power lost at the turn-on is simply transferred to external snubber circuits. Practically, spikes in the switch current waveforms are present at both the turn-on and turn-off.

The purpose of this paper is to present a power supply that operates over a wide-load and line range, I_O = 0.25 – 2.5 A and V_{DD} = 100 – 370 V, which corresponds to line ac voltage from 75 to 275 V_{RMS}. Therefore, the circuit is called a full-range power supply. The power supply is composed of a parallel class D^2 dc-dc resonant converter [7], where switching losses are drastically reduced and, therefore, the converter is suitable for a high switching frequency (f_{max} = 480 kHz) operation.

The significance of the paper is that the power supply operates with a high-efficiency operation with a low number of easily available components and, therefore, represents a satisfactory solution to the need of low-cost and high power-density power supplies.

Other advantages of the presented power supply are as follows:

1) The turn off losses are almost zero over the entire load range because the current in the inverter is almost load independent.
2) It is inerhently protected against load short circuit.
3) a safe no-load topology is achieved if the switching frequency is kept close to the resonant frequency.
4) Parasitic components, as such as MOSFET parasitic diodes and capacitances, are absorbed in the circuit operation.
5) Low number of currently available devices can be used in assembling the power supply so that both the high power-density and low-cost requirements are satisfied.

II. Principle Of Operation

The converter circuit is obtained with a parallel resonant inverter and a class D voltage-driven center-tapped rectifier [7, 8] with one capacitor in parallel with only one power MOSFET, as shown in Fig. 1. This provides reduction of the turn-on losses in both MOSFETs. The parallel resonant inverter circuit is better than the series resonant [11,15,16] because it is able to regulate the dc output voltage V_O over a wide load range, e.g., from 10 % to 110 % of the nominal output current as normally required from design specifications. Moreover, it has an efficiency which increases with output current. The parallel class D inverter has is chosen because the series-parallel class D inverter has a lower full-load efficiency [10,15,16]. The circuit is operated at the continuous current mode

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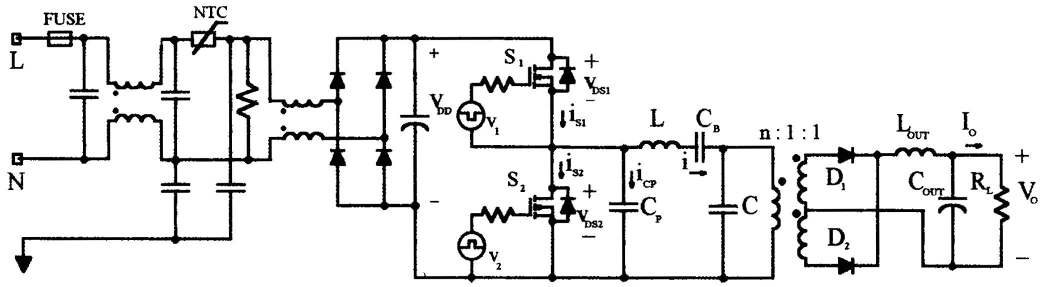

Fig. 1. Class D parallel resonant power supply.

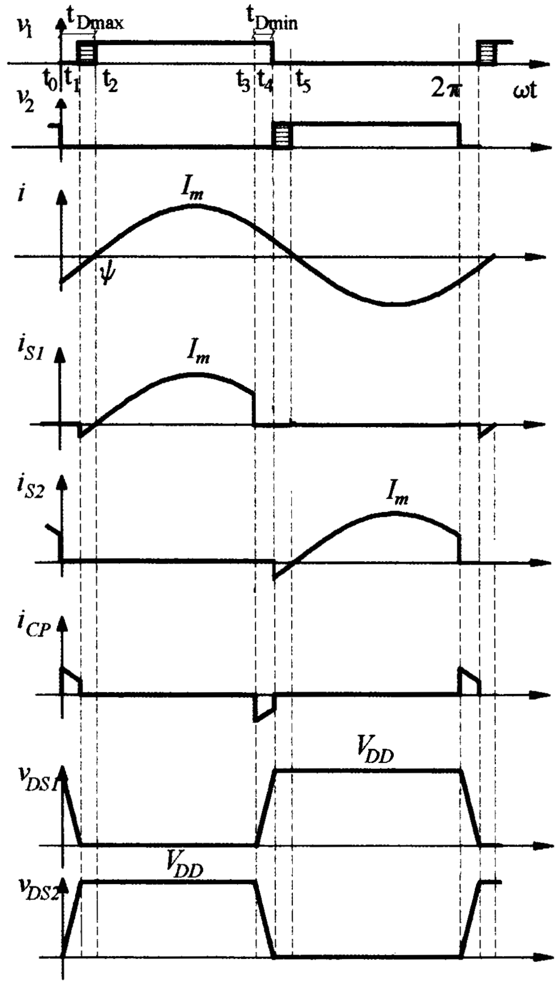

Fig. 2. Waveforms of class D inverter for $f > f_o$ with shunt capacitor C_P.

above the resonant frequency f_o over the entire load and input voltage ranges. The principle of operation of class D inverter is explained by the waveforms shown in Fig. 2: the current in the resonant circuit is lagging the voltage applied across the switches (that is, the fundamental harmonic of the square wave applied). As a result, turn on switching losses are eliminated because the MOSFET body-drain diodes conduct the inverse current and the voltages v_{DS1} and v_{DS2} are zero before the MOSFETs carry the forward currents (time interval t_1 - t_2). Moreover, diodes turn on at a very low *di/dt*, not generating current spikes, and the conducting diodes have a turn off time equal to the forward conduction time of the MOSFETs. So the flyback diodes can be slow and no external diodes are required. The parallel capacitor C_P forces the reverse voltage across both of the MOSFETs to slowly increase during their turn-off [8]. As a result, turn-off losses are drastically reduced.

For operation above resonance ($f > f_o$), MOSFET Q_1 conducts during the time interval t_2 - t_3 , MOSFET Q_2 is OFF and the voltage v_{DS2}, which is also the voltage across C_P , is equal to the dc input V_{DD}. From time t_3 to t_4 , the gate-to-source voltage turns Q_1 off and Q_2 is still off because of the dead time t_D . After time t_4, the current of the resonant circuit is diverted from Q_1 to the shunt capacitor C_P , the voltage v_{DS2} decreases according to $i_{CP} = dv_{DS}/dt$, and v_{DS1} is forced to increase from zero to V_{DD}. As a result, v_{DS1} slowly increases and crosses the decreasing current at almost zero level. After half a period, the voltage and current waveforms of Q_2 have the same waveforms of those of Q_1; therefore, Q_1 turns on with zero power loss and turn-off loss is also reduced for Q_2.

III. Design Example

An off-line power supply was designed to meet the following specifications:

1) output power P_o = 24 W at an output voltage V_o = 12 V;
2) input dc voltage V_{DD} ranging from 100 V to 370 V, that is, an ac rms voltage in the 80–240 V range;
3) f_o = 300 kHz.

Using the results of previous analysis [7] and using a loaded quality factor $Q = 3$, a class D inverter efficiency $h_I = 0.9$, a diode forward drop V_F = 0.5 V, a diode series resistance R_F = 0.03 W, and a transformer winding resistance r_F = 0.1 W, the design procedure is as follows. The dc load resistance is

$$R_L = \frac{V_o^2}{P_o} = 6\ \Omega\ . \tag{1}$$

The dc-to-dc voltage transfer function for the center-tapped rectifier is

$$M_R = \frac{M_R'}{n} = \frac{1}{n}\frac{2\sqrt{2}}{\pi\left(1+\frac{V_F}{V_O}+\frac{R_F+r_F}{R_L}\right)} = \frac{0.84}{n} \tag{2}$$

where n is the transformer turns ratio. The equivalent input resistance seen at the terminals of the primary winding is

$$R_R = R_L\frac{n^2\pi^2}{8}\left(1+\frac{V_F}{V_o}+\frac{R_F+r_F}{R_L}\right) = 7.87\,n^2\ \Omega\,. \tag{3}$$

If a minimum operating frequency $f = 1.05f_o$ is assumed at full power, the dc-to-ac voltage transfer function results

$$M_I = \frac{\sqrt{2}\,M_{I2}}{\pi} = \frac{\sqrt{2}}{\pi}\frac{1}{\sqrt{\left[1-\left(\frac{f}{f_o}\right)^2\right]^2+\left(\frac{1}{Q}\frac{f}{f_o}\right)^2}} = 1.23. \tag{4}$$

The rectifier efficiency at full power is

$$\eta_R = \left(1+\frac{V_F}{V_o}+\frac{R_F}{R_L}+\frac{r_f}{R_L}\right)^{-1} = 0.94\,. \tag{5}$$

The dc-to-dc voltage transfer function of the converter is

$$M = \frac{V_o}{V_{I\,min}} = \sqrt{\eta_I}\,M_I\,\frac{M_R'}{n} \tag{6}$$

which leads to n = 8.17. By choosing n = 9, we obtain M_{I2max} = 3.01, M_{I2min} = 0.79, and therefore, the maximum frequency for load regulation is

$$f_{max} = \sqrt{1-\frac{1}{2Q^2}-\sqrt{\left(1-\frac{1}{2Q^2}\right)^2+\frac{1}{\left|M_{I2min}\right|^2}-1}} = 1.46\,f_o\,. \tag{7}$$

The component values of the resonant circuit are calculated as follows

$$L = \frac{n^2 R_R}{2\pi f Q} = 112\ \mu\text{H}. \tag{8}$$

$$C = \frac{Q}{n^2 R_R 2\pi f} = 2.5\ \text{nF}. \tag{9}$$

Hence the characteristic impedance is $Z_o = \sqrt{L/C} = 211\,\Omega$. The maximum current through inductor L at full load operation and at maximum input voltage V_{DD} = 370 V is calculated as

$$I_m = \frac{2V_{DDmax}\,M_{I2min}\sqrt{1+\left(Q\frac{f_{max}}{f_o}\right)^2}}{\pi Z_o\,Q} = 1.32\ A\,. \tag{10}$$

The efficiency of the Class D parallel resonant inverter operated at full load with an input voltage V_{DD} = 100 V is

$$\eta_{PI} = \frac{P_O}{P_O+P_C+P_{TOFF}} = \left\{1+\frac{r_P}{R}\left[1+(Qf/f_o)^2\right] + \frac{2f\left[1+(Qf/f_o)^2\right]}{3I_M^2R}\left[V_I I_{1P}(t_{1P}+t_{2P})+\frac{V_{DD}I_M\sin\psi_P\,t_{1P}^2}{t_{2P}}\right]\right\}^{-1} = 0.96 \tag{11}$$

where

$$r_P = \frac{r_{DS1}+r_{DS2}}{2}+r_L+r_{RESR}\frac{(Qf/f_o)^2}{(Qf/f_o)^2+1}+r_{BESR} = r_M+r_L+r_{BSR} \tag{12}$$

is the total equivalent parasitic resistance of a class D parallel resonant inverter, r_{DS1} and r_{DS2} are the drain-to-source resistances of the two MOSFETs when ON, r_L is the equivalent series resistance (ESR) of the resonant inductor L, r_{RESR} is the ESR of the resonant capacitor C, and r_{BESR} is the ESR of the blocking capacitor C_B.

IV. Experimental Results

A breadboard of the converter was built and tested, using two International Rectifier IRF740 MOSFETs, Motorola MBR4035CT Schottky diodes, C_P = 1.5 nF, L = 110 mH, C = 3.9 nF, C_B = 1 mF, L_{OUT} = 35 mH, and C_{OUT} = 300 nF. The measured resonant frequency was f_o = 251 kHz. The driver voltages with adjustable dead time of 250 ns were implemented using a Unitrode U1825IC. The experimental results were measured as functions of the load resistance and input voltage at a fixed output voltage V_o = 12 V. The measured output power was varied from P_{Omin} = 2.5 W to P_{Omax} = 33 W and the switching frequency varied from 280 kHz at V_{DD} = 100 V and I_o = 2.8 A to 480 kHz at V_{DD} = 370 V and I_o = 0.2 A. Fig. 3 shows the efficiency of the converter as a function of the output current for two values of the ac input voltage V_{Irms} = 100 V and V_{Irms} = 235 V at V_o = 12 V.

Figs. 4 and 5 show the experimental voltage and current waveforms of the bottom MOSFET at V_{DD} = 141 V and V_{DD} = 333 V at a constant output resistance R_L = 4.7 W. The output voltage was regulated over the entire input voltage range with a narrow frequency variation from 294 kHz to 381 kHz. Fig. 6 zooms the turn-off transition and shows that the voltage v_{DS2} and the drain current i_{DS2} cross each other at almost zero values, resulting in nearly zero turn-off loss of the MOSFET.

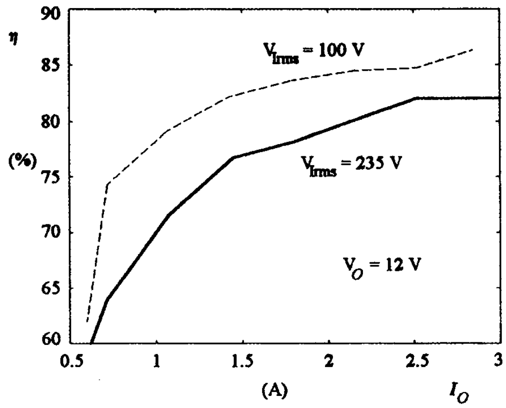

Fig. 3. Converter efficiency versus output current.

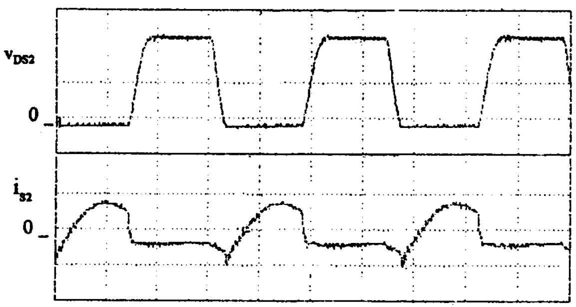

Fig. 4. Waveforms of voltage v_{DS2} and current i_{S2} at an operating frequency $f = 294$ kHz, ac input voltage $V_{Irms} = 100$ V, output voltage $V_O = 12$ V, and $R_L = 4.7\ \Omega$ ($I_O = 2.5$ A). Vertical: 40 V/div. for v_{DS2} and 1 A/div. for i_{S2}; horizontal: 1 μs/div.

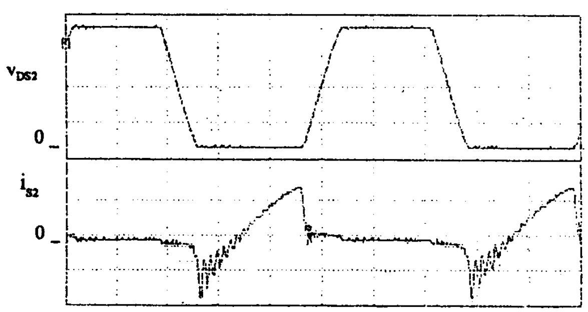

Fig. 5. Waveforms of voltage v_{DS2} and current i_{S2} at an operating frequency $f = 382$ kHz, ac input voltage $V_{Irms} = 235$ V, output voltage $V_O = 12$ V, and $R_L = 4.7\ \Omega$ ($I_O = 2.5$ A). Vertical: 185 V/div. for v_{DS2} and 1 A/div for i_{S2}; horizontal: 500 μs/div.

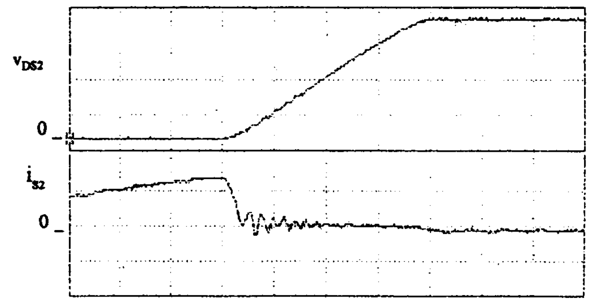

Fig. 6. Detail of voltage v_{DS2} and current at the turn-off of the MOSFET Q_2. Operating conditions are the same as in Fig. 4. Vertical: 185 V/div. for v_{DS2} and 1 A/div. for i_{S2}; horizontal: 500 μs/div.

Fig. 7 shows the current waveform through the primary winding of the transformer and the output voltage ripple v_r. The current was a square wave as expected, and the peak-to-peak voltage ripple was as low as 100 mV, which is less than 1% of the output voltage.

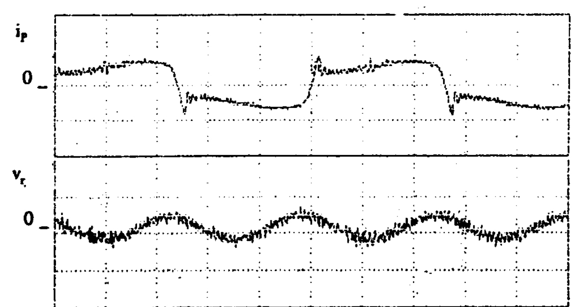

Fig. 7. Waveforms of the current i_p through the primary winding of the transformer and the output voltage ripple v_r at an operating frequency $f = 386$ kHz, ac input voltage $V_{Irms} = 235$ V ($V_{DD} = 235$ V), output voltage $V_O = 12$ V, and $R_L = 4.7\ \Omega$ ($I_O = 2.5$ A). Vertical: 500 mA/div. for i_p and 100mV/div. for v_r ; horizontal: 500 ns/div.

References

[1] Kazimierczuk, M. K. and X.T. Bui, "A family of class E resonant dc-dc power converters", *Proc. 16th Intern. PCI '88 Conf. (SATECH '88)*, Dearborn, MI, October 3-6, 1988, pp.383-394.

[2] Kazimierczuk, M. K. and K. Puczko, " Power-output capability of class E amplifier at any loaded Q and switch duty cycle", *IEEE Trans. Circuits Syst.*, vol CAS-36, pp. 1142-1144, August 1989.

[2] Kazimierczuk, M. K. and K. Puczko, " Power-output capability of class E amplifier at any loaded Q and switch duty cycle", *IEEE Trans. Circuits Syst.*, vol CAS-36, pp. 1142-1144, August 1989.

[3] J. Jozwik and M.K. Kazimierczuk, "Analysis and design of Class E^2 dc-dc converter", *IEEE Trans. Ind. Electron.*, vol IE-37, pp.173-183, April 1990.

[4] K. Schmidtner, "A new high frequency resonant converter topology", *Proc. High Freq. Power Conversion Conf.*, pp.390-403, May 1988, San Diego, CA.

[5] S.L. Smith and S. Robinson, "An off-line, one MHz 350-Watt parallel resonant converter (PRC) utilizing an RF transformer", *High Frequency Power Conversion Conference*, pp. 446-466, May 1988, San Diego, CA.

[6] R. Myers and R.D. Peck, "200-KHz Power FET Technology in New Modular Power Supplies", *Hewlett-Packard Journal*, pp. 3-10, August 1981.

[7] Kazimierczuk, M. K., Szaraniec, W., and Wang, S., "Analysis and design of parallel resonant converter at high Q", *IEEE Trans. Aerospace Electron. Syst.*, Vol. 28, No. 1, January 1992, pp. 35-50.

[8] Kazimierczuk, M. K. and W. Szaraniec, "Class zero-voltage-switching inverter with only one shunt capacitor", *Proc. Inst. Elect. Eng., Pt. B, Elec. Power Appl.*, vol. 139, Nov. 1992.

[9] M. K. Kazimierczuk, "Class D current-driven rectifiers for dc/dc converter applications", *IEEE Trans. Ind. Electron.*, vol. IE-38, pp. 344-354, Nov. 1991.

[10] Kazimierczuk, M. K., Thirunarayan, N., and Wang, S., "Analysis of series parallel resonant converters", *IEEE Trans. Aerospace Electron. Syst.*, Vol. 29, No. 1, January 1993, pp. 88-99.

[11] R. J. King and T. A. Stuart, "A normalized model for the half-bridge series resonant converter", *IEEE Trans. Aerospace Electron. Syst.*, vol. AES-17, pp. 190-198, Mar. 1981.

[12] V. Vorpérian and S. Cuk, "A complete dc analysis of the series resonant converter", in *IEEE Power Eletron. Specialists Conf. Rec.*, Cambridge, MA, June 14-17, 1982, pp. 85-100.

[13] Kazimierczuk, M. K. and Wang, S., "Frequency domain analysis of series converter for continuous conduction mode", *IEEE Trans. Power Electron.*, Vol. 7, No. 2, April 1992, pp. 270-279.

[14] R. L. Steigerwald, "High-frequency resonant transistor dc-dc converter", *IEEE Trans. Power Electron.*, pp. 181-191, May 1984.

[15] R. L. Steigerwald, "A comparison of half-bridge resonant converter topologies", *IEEE Trans. Power Electron.*, vol. PE-3, pp. 174-182, Apr. 1988.

[16] Kazimierczuk, M. K. and W. Szaraniec, "Elctronic ballast for fluorescent lamps", *IEEE Trans. Power Electron.*, Vol. PE-8, pp. 386-395, Oct. 1993.