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## Hybridge Zero-Current-Switching Rectifier for High-Frequency DC-DC Converter Applications

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**Abstract** – An analysis and experimental verification of a hybrid current-driven zero-current-switching rectifier are given. In the hybrid rectifier, the diodes turn on at zero  $di/dt$ , low  $dv/dt$ , and turn off at low  $di/dt$ . Therefore, the hybrid rectifier is suitable for high-frequency high-efficiency applications. The current source driving the rectifier can be constituted by a class D or a class E inverter. Since these inverters operate with nearly zero switching losses, high power-density dc-dc converters can be assembled by using resonant inverters and hybrid rectifiers. A dc-dc converter based on a class D series resonant inverter and a hybrid rectifier was designed and tested for a constant output voltage  $V_O = 12$  V and an output current  $I_O$  from no-load to 12 A. The converter was operated at a dc input voltage ranging from 200 to 380 V and a minimum frequency of  $f = 600$  kHz. The theoretical results and measured performances were in good agreement. A maximum converter efficiency as high as 90% has been achieved for the dc-dc converter operated at full load.

### I. INTRODUCTION

High-frequency rectifiers are needed to build high power-density dc-to-dc converters. Bridge rectifiers provide a full-wave rectification and require a simple transformer with only one secondary winding. However, two diodes are simultaneously ON during each half-period. This results in high power losses at a high output current operation. Moreover, voltage drops across the diodes make the full-bridge rectifier unsuitable for low output voltage applications. Rectifier losses can be reduced using the well-known center-tapped rectifier. However, the transformer in this circuit has two secondary windings which are not efficiently used because the current flows in each secondary winding only for one-half of the switching period. In the hybrid rectifier, a transformer with only one secondary winding is used [1]–[7]. Moreover, the rms value of the secondary side ac-current is half of that circulating in the center-tapped rectifier. The limitation of this rectifier circuit is that high-switching losses and high noise level are produced at frequencies above 500 kHz if it is operated with a pulse-width-modulated (PWM) voltage or current source at the primary side. Thus, soft-switching operation is highly desirable to achieve an high-frequency high-efficiency full-wave rectification. This is achieved if the hybrid rectifier is driven by a sinusoidal current source. Actually the current source can be realized by using a class E or a class D series resonant inverter [7]–[15]. Switching losses are nearly

zero in these inverters, and, therefore, they can be operated at high frequency with an high efficiency. As a result, high power-density dc-dc converters can be assembled by using resonant inverters and hybrid rectifiers.

The purpose of this paper is to present design equations and experimental results for a hybrid current-driven zero-current-switching (ZCS) low  $di/dt$  rectifier. This rectifier preserves major advantages of conventional bridge and center-tapped rectifiers and overcomes many limitations of these circuits. As in full-wave full-bridge and center-tapped rectifiers, the filter capacitors are small because they are operated at a frequency which is twice that of the input source. Conduction losses in the diodes are reduced because the average current through each diode is one-half of the output dc current as in center-tapped rectifiers. However, each inductor of the hybrid rectifier carries one-half of the dc output current while the entire output current flows through the output inductor in the center-tapped rectifier. As a result, conduction losses are lower in the hybrid rectifiers. The hybrid current-driven zero-current-switching (ZCS) low  $di/dt$  rectifier overcomes also the limits of its PWM counterpart because the diode current and voltage waveforms do not overlap at diode transitions. Moreover, the diodes turn on at zero voltage and zero current with zero  $di/dt$  and limited  $dv/dt$  and turn off with a limited  $di/dt$ . As a result, the switching losses and noise level are drastically reduced in the hybrid current-driven zero-current-switching (ZCS) low  $di/dt$  rectifier. Therefore, this rectifier can be operated at higher frequencies than a PWM hybrid rectifier, resulting in a lower volume and weight. This allows for operations at low output voltage ripple also when small capacitances and inductances are used.

The hybrid current-driven zero-current-switching (ZCS) low  $di/dt$  rectifier circuit was used in combination with a class D series resonant inverter used as a current source. A 144 W dc-dc converter was assembled and operated as a voltage regulator at a dc voltage ranging from 200 to 380 V dc, an output voltage  $V_O = 12$  V regulated over the entire line voltage range and from no-load to a full load  $I_O = 12$  A. The dc-to-dc converter was frequency-controlled in the 650-to-950 kHz range. A maximum dc-to-dc converter efficiency  $\eta$  above 90% was achieved at full load.

II. RECTIFIER CIRCUIT DESCRIPTION AND OPERATION

A circuit for a hybrid current-driven ZCS low  $di/dt$  is shown in Fig 1. Figs 1(a) and (b) show the non-inverting and the inverting circuit of the hybrid rectifier. Each of the two circuits are driven by a sinusoidal current source with

$$i = I_m \sin \omega t \tag{1}$$

where  $I_m$  is the current amplitude,  $\omega = 2\pi f$  is the angular frequency of the current, and  $f$  is the operating frequency of the current source and rectifier transformer. The rectifier circuit consists of two diodes  $D_1$  and  $D_2$ , two inductors  $L_1$  and  $L_2$ , and a filter capacitor  $C_f$ . Resistance  $R_L$  represents the dc load. The two inductors and single-pole low-pass rectifier composed by the parallel combination of  $C_f$  and  $R_L$  are operated at a frequency of  $2f$  while diodes are operated at a frequency of  $f$ . This results in small volume inductors and capacitors. Moreover, a low ac ripple of the output voltage can be easily achieved with reduced values of capacitance  $C_f$  because the filter capacitor is operated at twice the input current frequency.

In the inverting rectifier circuit of Fig. 1(b), the two diodes have a common cathode connection. Therefore, the center tapped rectifiers normally available on the market with two diodes built in the same case can be used. This results in a reduction of the number of components and volume of the rectifier circuit.

Figs. 2(a) to (d) and (e) to (h) show the four topological modes the rectifier of Fig. 1(a) goes through during one switching period for a diode on-duty cycle  $0 < D \leq 0.5$  and  $0.5 < D \leq 1$ , respectively.

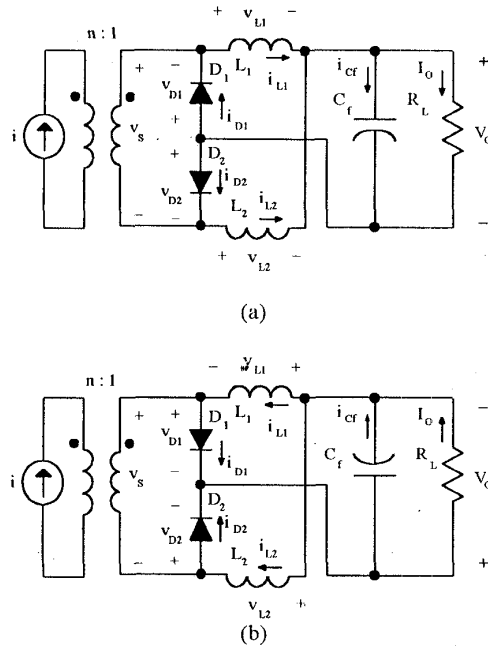


Fig. 1. Schematic circuit of the hybrid current-driven rectifier.

- (a) Non-inverting circuit.
- (b) Inverting circuit

The inverting rectifier circuit of Fig. 1(b) goes through four topological modes as the non-inverting rectifier. The difference between the two rectifier topological modes are the current directions in the circuit components.

The voltage and current waveforms for the rectifier of Fig. 1(a) are shown in Fig. 3. These waveforms are the same also for the inverting rectifier of Fig. 1(b).

A. Operation for  $0 < D \leq 0.5$

During the first topological mode of Fig. 2(a),  $D_1$  is ON, the load is supplied by the current source through the transformer secondary side represented by a current source  $ni$ . When  $D_1$  is ON, the voltage across  $L_1$  is equal to the reverse output voltage  $V_O$ , and the current through diode  $D_1$  is  $i_{D1} = i_{L1} - ni$  and increases with limited  $di/dt$ . Moreover,  $di/dt$  is zero at diode  $D_1$  turn-on. When the  $i_{D1}$  reaches zero, diode  $D_1$  turns off and mode II begins. During mode II inductors  $L_1$  and  $L_2$  and the transformer are in the same current path and the output capacitor  $C_f$  supplies the current to the dc load. The voltages across the inductors are sinusoidal and the reverse voltage across the diodes are portion of a sinusoid shifted downward by a DC voltage  $V_O$ . When the voltage across  $D_2$  reaches its threshold,  $D_2$  turns on at low  $dv/dt$  and zero  $di/dt$  with theoretically zero switching losses. During mode III,  $I_O$  partially flows through diode  $D_2$  in series with inductor  $L_2$  and partially through inductor  $L_1$ . When diode  $D_2$  turns off, mode III ends and mode IV begins. The current path of mode IV is identical to that of mode II.

B. Operation for  $0.5 < D \leq 1$

The circuit operation for  $0.5 < D \leq 1$  is similar to that described for  $0 < D \leq 0.5$ . The main difference is that both  $D_1$  and  $D_2$  are ON during modes I and III. As a consequence, both inductors  $L_1$  and  $L_2$  store part of the energy absorbed by the load resistance, reducing the current through the transformer windings and the ac voltage ripple across  $C_f$ .

III. RECTIFIER DESIGN EQUATIONS

The equations for the rectifier design are derived under the following assumptions:

- 1) Diodes are modelled as ideal switches.
- 2) The transformer is ideal, the turns ratio is  $n$ , and the transformer leakage inductances are neglected.
- 3) The output voltage is equal to  $V_O$  and is assumed to have a zero ripple.
- 4) The rectifier circuit is symmetric. Therefore, we have  $L_1 = L_2 = L$ .

Results of the analysis are valid for both the operation at  $0 < D \leq 0.5$  and  $0.5 < D \leq 1$ .

The initial turn-on delay angle of diode  $D_1$  is

$$\phi = \text{tg}^{-1} \left[ \frac{\sin 2\pi D - 2\pi D}{1 - \cos 2\pi D} \right] \tag{2}$$

Since a symmetrical operation of the rectifier is assumed, angle  $\phi$  is also the delay angle of diode  $D_2$ . A plot of  $\phi$  as a function of  $D$  is depicted in Fig. 4. The amplitude of the sinusoidal input current is given as function of  $\phi$  as follows

$$I_m = - \frac{V_O}{n\omega L \cos \phi} \tag{3}$$

The load resistance  $R_L$  normalized with respect to the reactance  $\omega L = \omega L_1 = \omega L_2$  as a function of the on-diode duty cycle is

$$\frac{R_L}{\omega L} = \frac{\pi(1 - \cos 2\pi D)}{(1 - \cos 2\pi D)^2 + (\sin 2\pi D - 2\pi D)^2 + 2\pi^2 D^2(1 - \cos 2\pi D)} \tag{4}$$

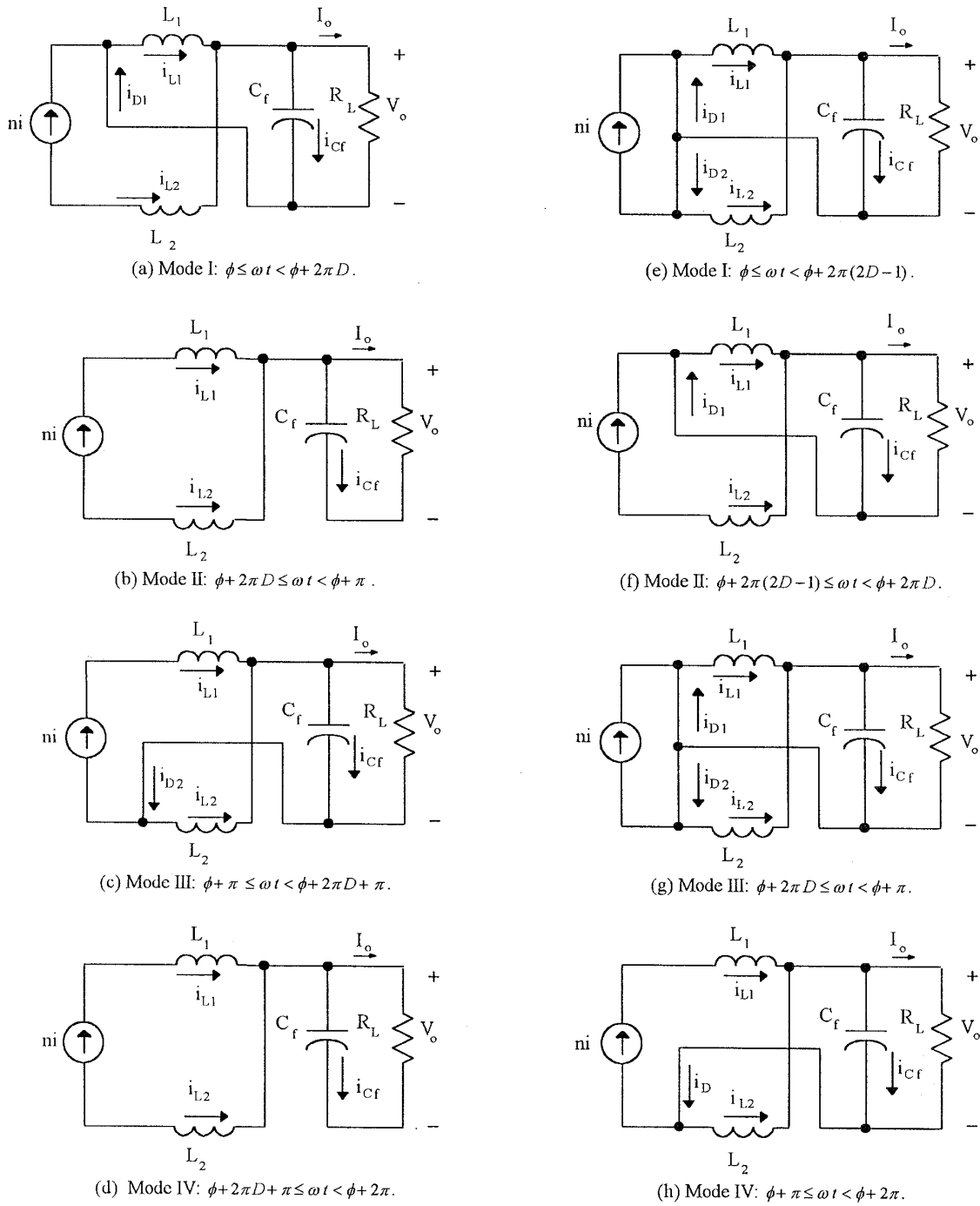


Fig. 2. Equivalent circuits of the hybrid current rectifier for various intervals.

- (I) Operation at  $0 \leq D < 0.5$ : (a)  $D_1$  ON and  $D_2$  OFF, (b)  $D_1$  OFF and  $D_2$  OFF, (c)  $D_1$  OFF and  $D_2$  ON, (d)  $D_1$  OFF and  $D_2$  OFF.
- (II) Operation at  $0.5 \leq D < 1$ : (e)  $D_1$  ON and  $D_2$  ON, (f)  $D_1$  ON and  $D_2$  OFF, (g)  $D_1$  ON and  $D_2$  ON, (h)  $D_1$  OFF and  $D_2$  ON.

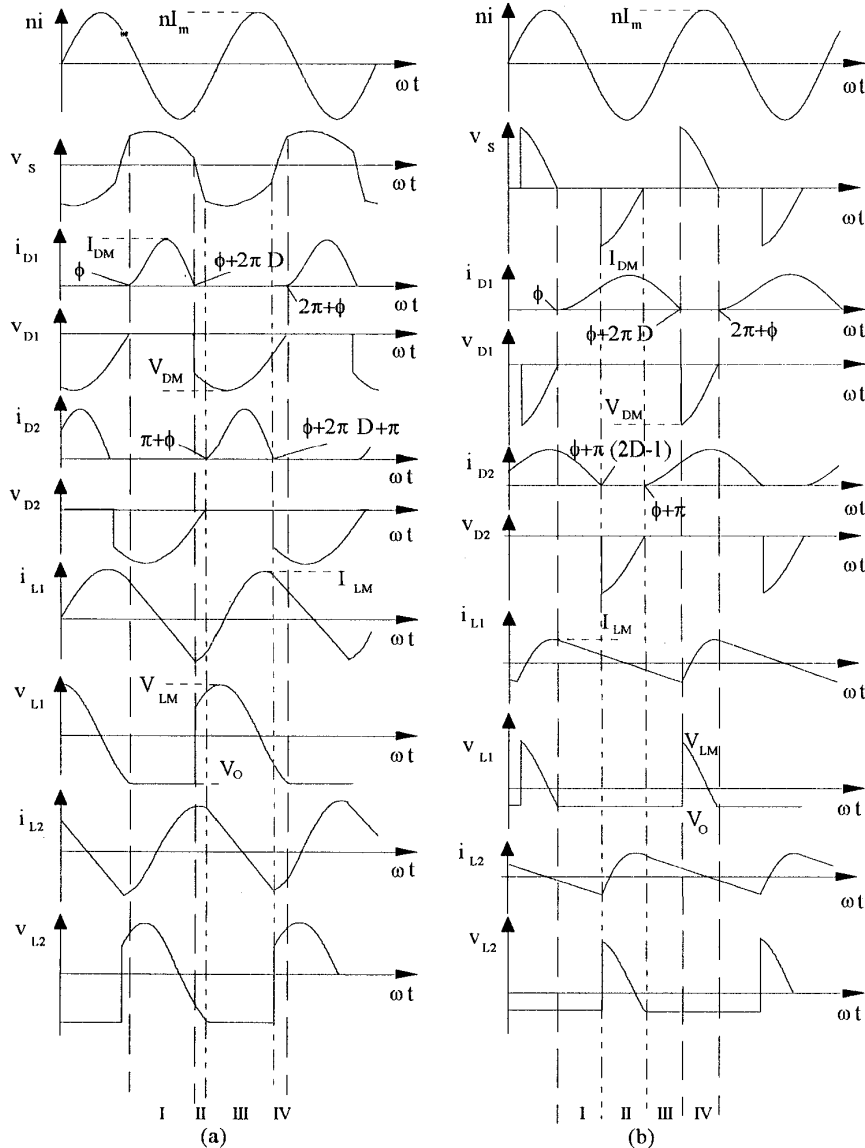


Fig. 3. Voltage and current waveforms in the hybrid ZCS rectifier. (a) For  $0 < D \leq 0.5$ . (b) For  $0.5 < D \leq 1$ .

Fig. 5 shows the on-duty cycle  $D$  as a function of the load resistance  $R_L$  normalized with respect to the reactance  $\omega L$ . The maximum current through the inductors normalized with respect to  $I_O$  is given by

$$\frac{I_{LM}}{I_O} = -\frac{R_L}{\omega L \cos \phi} \frac{\pi(\cos 2\pi D - 1)}{\cos \phi \left[ (1 - \cos 2\pi D)^2 + (\sin 2\pi D - 2\pi D)^2 + 2\pi^2 D^2 (1 - \cos 2\pi D) \right]} \quad (5)$$

and is plotted in Fig. 6 as a function of  $R_L / \omega L$ . The maximum current through each diode is expressed as

$$\frac{I_{DM}}{I_O} = -2 \frac{R_L}{\omega L} (\pi - \phi + \tan \phi) \frac{2\pi[(\phi - \pi)(1 - \cos 2\pi D) - \sin 2\pi D + 2\pi D]}{(1 - \cos 2\pi D)^2 + (\sin 2\pi D - 2\pi D)^2 - 2\pi^2 D^2 (1 - \cos 2\pi D)} \quad (6)$$

Fig. 7 shows the plot of  $I_{DM} / I_O$  versus  $R_L / \omega L$ . The maximum reverse diode voltage across each diode normalized with respect to  $V_O$  is

$$\frac{V_{DM}}{V_O} = 1 - \frac{1}{\cos \phi} \quad (7)$$

and is plotted as a function of  $R_L / \omega L$  in Fig. 8.

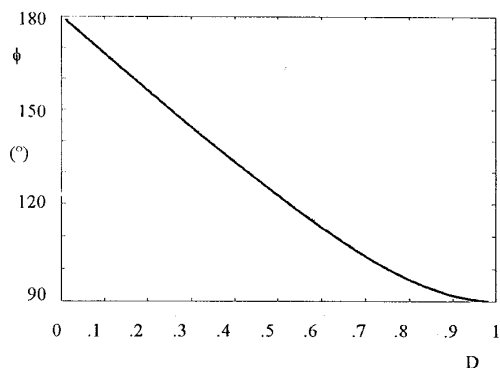


Fig. 4. Delay angle  $\phi$  as a function of the diode on-duty cycle  $D$

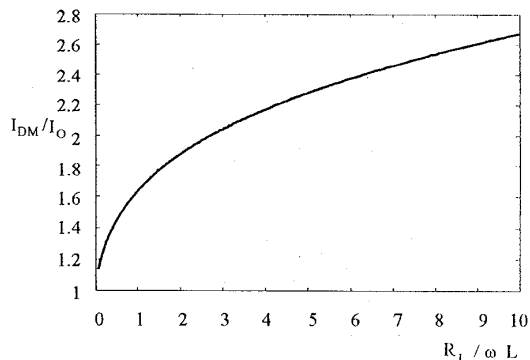


Fig. 7. Maximum normalized diode current  $I_{DM}/I_O$  as a function of the normalized load resistance  $R_L/\omega L$ .

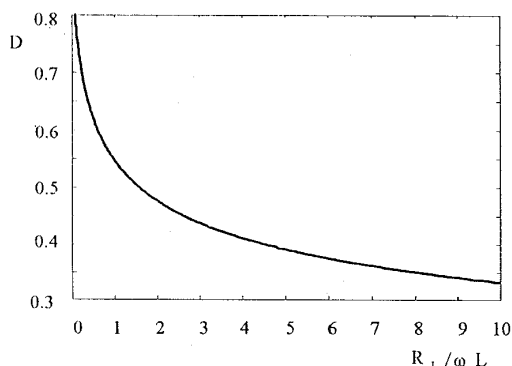


Fig. 5. Duty-cycle  $D$  as a function of the normalized load resistance  $R_L/\omega L$ .

The voltage across the equivalent input resistance of the rectifier as seen at the terminals of the transformer primary side normalized with respect to the output voltage  $V_O$  is given by

$$\frac{V_{Rim}}{nV_O} = 2\cos(\phi + 2\pi D) - 2\cos\phi + \frac{\sin^2(\phi + 2\pi D) - \sin^2\phi}{\cos\phi} \quad (8)$$

and is plotted in Fig. 9 as a function of the normalized load resistance.

The input resistance of the rectifier normalized with respect to the load resistance  $R_L$  is given by

$$\frac{R_I}{n^2 R_L} = \frac{\omega L}{\pi R_L} x \left[ 2 - \sin^2\phi + \cos^2\phi - \cos\phi \cos(\phi + 2\pi D) - \sin^2(\phi + 2\pi D) \right] \quad (9)$$

and is plotted in Fig. 10 as a function of the normalized load resistance  $R_L/\omega L$ .

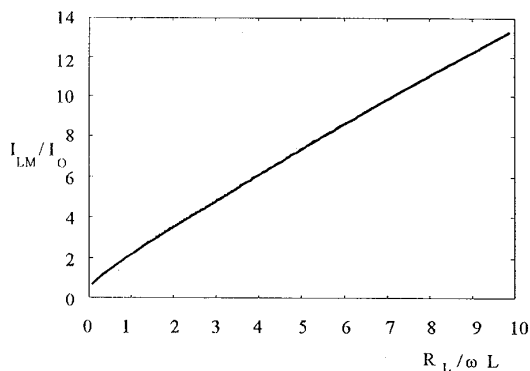


Fig. 6. Maximum normalized inductor current  $I_{LM}/I_O$  as a function of the normalized load resistance  $R_L/\omega L$ .

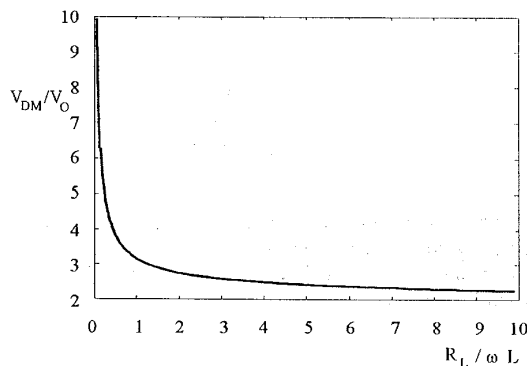


Fig. 8. Maximum normalized diode voltage  $V_{DM}/V_O$  as a function of the normalized load resistance  $R_L/\omega L$ .

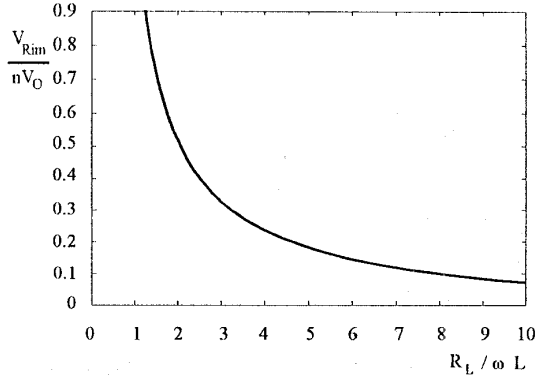


Fig. 9. Normalized primary side resistive voltage  $V_{Rim} / n V_O$  as a function of the normalized load resistance  $R_L / \omega L$ .

The design parameters of the hybrid ZCS rectifier are summed up in Table I.

TABLE I. RECTIFIER DESIGN PARAMETERS

$D$	$\phi$ (°)	$R_L / \omega L$	$I_{DM} / I_O$	$V_{DM} / V_O$
0.1	168	143	8.89	2.02
0.2	156	86	4.44	2.09
0.3	144	15.8	2.96	2.23
0.4	133	4.49	2.22	2.46
0.5	122	1.57	1.78	2.86
0.6	112	.6	1.49	3.61
0.7	104	.233	1.28	5.21
0.8	96.6	.079	1.137	9.7
0.9	91	.0167	1.04	33.7

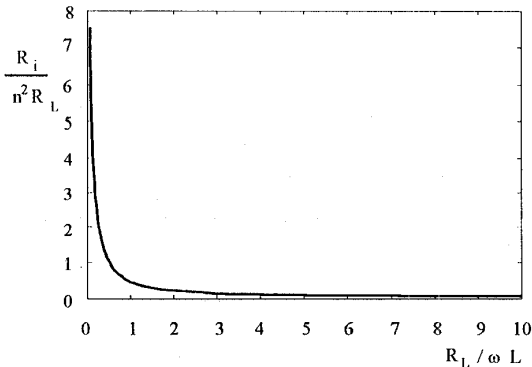


Fig. 10. Normalized input resistance  $R_i / n^2 R_L$  as a function of the normalized load resistance  $R_L / \omega L$ .

#### IV DESIGN EXAMPLE

A dc-dc converter was built and tested to achieve the following design specifications:

- Output power  $P_O = 144$  W.
- Output voltage  $V_O = 12$  V.
- Output current  $I_O$  from 3 to 100 % of the nominal output current  $I_{ON} = 12$  A.
- Input voltage  $V_{Imin} = 200$  Vdc and  $V_{Imax} = 380$  Vdc.
- Output power  $P_O = 144$  W.
- Minimum transformer operating frequency  $f = 600$  kHz.

The schematic circuit of the dc-dc converter constituted by a class D resonant inverter and a hybrid rectifier is shown in Fig. 11. The inverter is operated as a nearly sinusoidal current source.

The minimum load resistance is  $R_{Lmin} = V_O / I_{ON} = 1 \Omega$ . Assuming a diode duty cycle  $D = 0.5$ , the normalized load resistance resulting from the plot of Fig. 5 is  $R_L / \omega L = 1.57$ . Therefore, the two inductors  $L_1$  and  $L_2$  have an inductance of  $L = R_L / 1.57 \omega = 1 / 1.57 \times 2 \pi 500 \times 10^3 = 200$  nH.

From Figs. 5 and 6 the maximum diode currents and voltages are  $I_{DM} = 1.78 I_O = 21.3$  A, and  $V_{DM} = 2.86 V_O = 35$  V, respectively. The converter dc voltage transfer function must be  $M_{Vdc} = V_O / V_{Imin} = 12 / 200 = 0.06$ . The class D series resonant inverter voltage transfer functions is given in [14]. Assuming an inverter efficiency  $\eta_I = 90\%$ , a loaded quality factor  $Q_L = 5$ , and an inverter operating frequency  $f = 1.05 f_O$ , where  $f_O$  is the resonance frequency of the inverter LC circuit, the dc-to-ac inverter voltage transfer function is calculated as

$$M_I = \frac{\sqrt{2} \eta_I}{\pi \sqrt{1 + Q_L^2 \left( \frac{f}{f_O} - \frac{f_O}{f} \right)^2}} = \frac{\sqrt{2} \times 0.9}{\pi \sqrt{1 + 25(1.05 - 1.05^{-1})^2}} = 0.37. \quad (10)$$

The rectifier ac-to-dc transfer function is  $M_R = V_O / V_{Rin} = M_{Vdc} / M_I = 0.06 / 0.37 = 0.16$ . As shown in Fig. 9, the voltage across the equivalent resistance of the rectifier seen at the terminals of the transformer primary side is  $V_{Rin} / n V_O = 0.68$  for  $R_L / \omega L = 1.57$ . The transformer turns ratio is calculated as  $n = (0.16 \times 0.68)^{-1} = 10$ . As shown in Fig. 10 the normalized input resistance of the rectifier is  $R_i / n^2 R_L = 0.234$ , and, therefore, the input resistance is  $R_i = 10^2 \times 1 \times 0.234 = 23.4 \Omega$ . Assuming a resonance frequency of  $f_O = 600$  kHz for the inverter, the inductance and capacitance of the resonant circuit are calculated as  $L_R = Q_L R_i / 2 \pi f_O = 5 \times 23.4 / 2 \pi \times 600 \times 10^3 = 31 \mu\text{H}$  and  $C_R = 1 / (2 \pi f_O)^2 L_R = 1 / (2 \pi \times 600 \times 10^3)^2 \times 31 \times 10^{-6} = 2.2$  nF.

#### V. EXPERIMENTAL RESULTS

The dc-to-dc converter assembled with a class-D series resonant dc-ac inverter and an inverting hybrid rectifier was frequency-controlled in the 600-to-950 kHz range. Two International Rectifier IRF740 MOSFETs were used in the inverter. A Philips 2.2 nF/250 V polyester capacitor was used as a resonant capacitor. The resonant inductor was assembled by winding 60 turns of a 90x0.1 Litz wire on an Amidon iron-powder Mix #2 T-94 toroidal core. A quality factor  $Q = 200$  was measured for the resonant inductor at an operating frequency varying from 500 to 700 kHz. The transformer was built on a ETD29 Siemens N87 ferrite core. The primary was wound with 24 turns of 60 x 0.1 mm LITZ wire and the secondary winding with 2 turns of copper strips 15 x 0.15 mm. A Motorola 30C145 Schottky diode was used for the rectifier. The filter inductances  $L_1 = L_2 = 0.2 \mu\text{H}$  were assembled by using two Amidon T50 iron-powder Mix #2 toroidal cores. One electrolytic Oscon 100  $\mu\text{F}$  capacitor was used as filter capacitors. The operation at the minimum frequency is achieved for the minimum input voltage  $V_I = 200$  V and the maximum load  $I_O = 12$  A and the converter is operated at the maximum frequency when the input voltage is  $V_{Imax} = 380$  V and output current  $I_O = 1.2$  A. Fig. 12 shows the experimental waveforms of the voltage and current of the inverter lower MOSFET  $M_2$ . Since the inverter is operated above the resonant frequency, the current lags the voltage and is negative at the MOSFET turn-on.

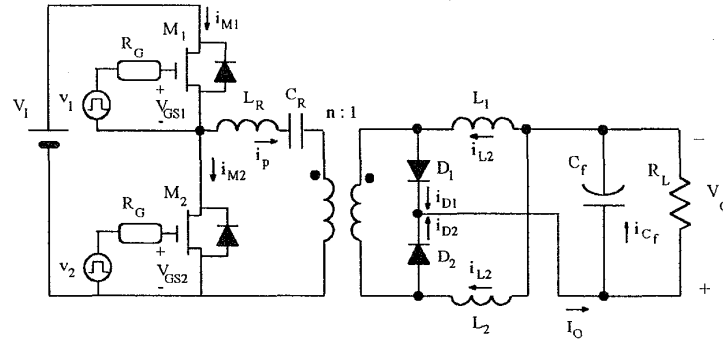


Fig. 11. Schematic circuit of the dc-dc converter constituted by a class D series resonant inverter and an inverting hybrid rectifier.

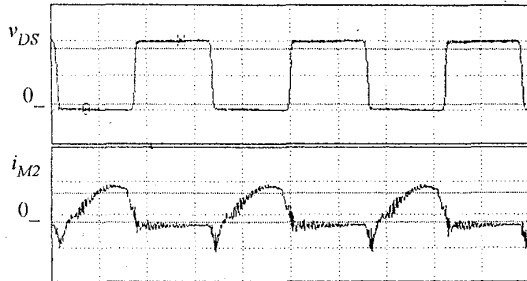


Fig. 12. Waveforms of MOSFET  $M_2$  drain to source voltage  $v_{DS}$  and source current  $i_{M2}$  at an operating frequency of  $f = 615$  kHz, dc input voltage  $V_1 = 200$  V, output voltage  $V_O = 12$  V, and  $I_O = 12$  A. Vertical: upper trace 100V/div.; lower trace 2 A/div., horizontal: 500 ns/div..

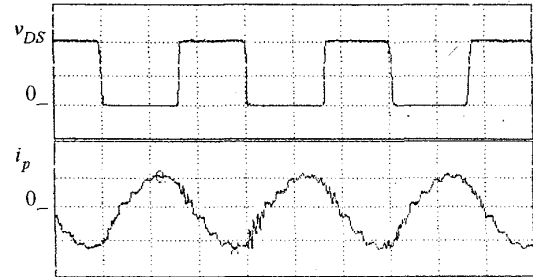


Fig. 13. Waveforms of MOSFET  $M_1$  drain to source voltage  $v_{DS}$  and transformer primary side current  $i_p$  at an operating frequency of  $f = 660$  kHz, dc input voltage  $V_1 = 250$  V, output voltage  $V_O = 12$  V, and  $I_O = 12$  A. Vertical: upper trace 125 V/div.; lower trace 2.5 A/div., horizontal: 500 ns/div..

Therefore, the current flows through the MOSFET parasitic diode which clamps the drain-to-source voltage at a low value, e.g., 0.5 V. This allows for a nearly zero-loss turn-on of the inverter MOSFETs.

The transformer primary side current waveform shown in Fig. 13 is nearly sinusoidal. This demonstrates that the class D series resonant inverter is operated as a sinusoidal current source also at low loaded quality factors. The waveforms of the voltage  $v_p$  across the terminals of the transformer primary winding and the current through MOSFET  $M_2$  are shown in Fig. 14. The parasitic oscillation of the experimental waveform  $v_p$  was caused by the transformer leakage inductance which resonates with the diode parasitic capacitances.

The waveforms of the output voltage ripple  $v_{oac}$  and the ac component of current  $i_{L1ac}$  through inductor  $L_1$  are shown in Fig. 15. Current  $i_{L1ac}$  was observed with a passive current probe was nearly the same as the theoretically predicted current shifted downwards by the inductor dc component  $I_{L1dc} = I_O/2 = 6$  A. Since the measured maximum ac current was  $i_{L1ac} = 8$  A, the inductor maximum currents through the inductors were  $I_{LM} = 14$  A.

A plot of the dc-to-dc converter efficiency versus the ac input voltage measured at an output current  $I_O = 12.5$  A and an output voltage  $V_O = 12$  V is depicted in Fig. 15(a). The efficiency was higher than 84% over the entire voltage range and reached a maximum of 88.5%. A plot of the measured efficiency as a function of  $I_O$  at a dc input voltage  $V_1 = 310$  V and an output voltage  $V_O = 12$  V is shown in Fig. 15 (b). An efficiency  $\eta = 90\%$  was at full load, and the efficiency was higher than 80% for a dc output current varying from 2.3 A to 12 A, which corresponds to a load in the 30% to 100% range.

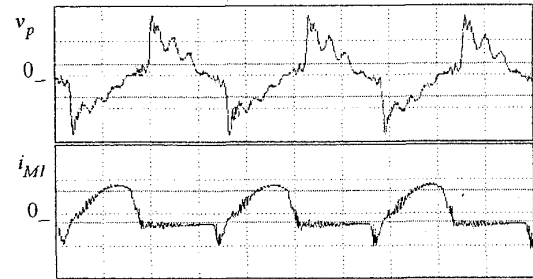


Fig. 14. Waveforms of the transformer primary side voltage  $v_p$  and MOSFET  $M_1$  source current  $i_{M1}$  at an operating frequency of  $f = 615$  kHz, dc input voltage  $V_1 = 200$  V, output voltage  $V_O = 12$  V, and  $I_O = 12$  A. Vertical: upper trace 100 V/div.; lower trace 2 A/div., horizontal: 500 ns/div..

#### IV. CONCLUSIONS

A circuit of a hybrid zero  $di/dt$  low  $dv/dt$  rectifier has been introduced. The equations describing the circuit operation been presented and verified experimentally. This rectifier preserves major advantages of conventional bridge and center-tapped rectifiers and overcomes many limitations of these circuits. Conduction losses are reduced because the average current through each diode and inductor is one-half of the output dc current as in center-tapped rectifiers.



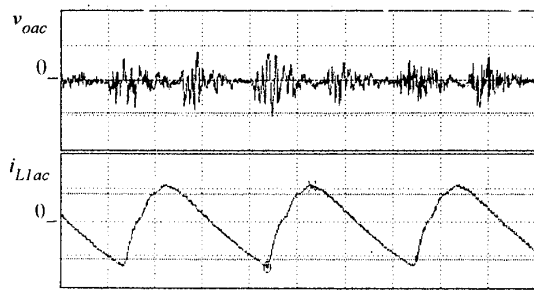


Fig. 15. Waveforms of output voltage ripple  $v_{oac}$  and ac component  $i_{L1ac}$  of the current through inductor  $L_1$  at an operating frequency of  $f = 660$  kHz, dc input voltage  $V_I = 250$  V, output voltage  $V_O = 12$  V, and  $I_O = 12$  A. Vertical: upper trace 50 mV/div.; lower trace 8 A/div., horizontal: 500 ns/div..

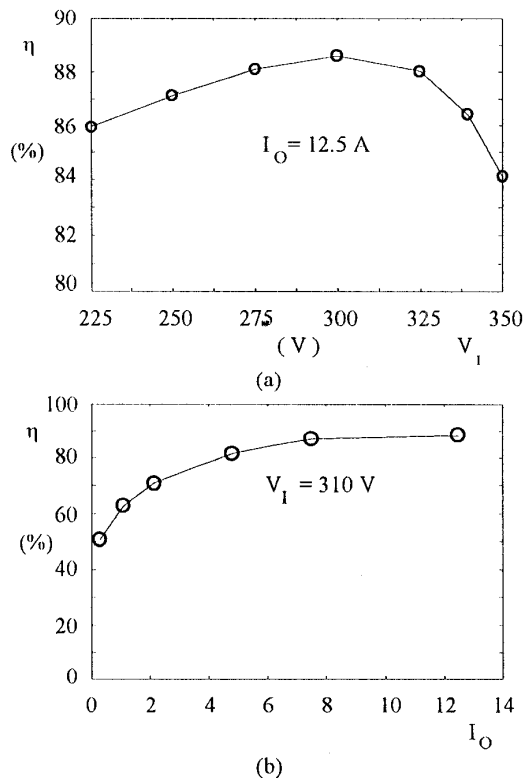


Fig. 16 Efficiency of the dc-dc converter at an output voltage  $V_O = 12$  V. (a) Efficiency versus input voltage  $V_I$ . (b) Efficiency versus output current  $I_O$ .

A full-wave rectification is achieved with two diodes and a transformer with only one secondary winding. The diode current and voltage waveforms do not overlap at diode transitions and the diodes turn on at zero voltage and zero current with zero  $di/dt$  and limited  $dv/dt$  and turn off with a limited  $di/dt$ . As a result, switching losses and noise level are drastically reduced in the current-driven circuit. The feasible high-frequency operation of the hybrid rectifier is possible allows for small capacitances and inductances to be used. Finally, the current-driven hybrid ZCS low  $di/dt$  full-wave rectifier is compatible with high-frequency inverters as class D and class E series resonant inverters in assembling dc-dc

converters.

A possible limitation of the current-driven hybrid rectifier is the high number of wound components. This does not significantly affect the overall volume and weight of the rectifier but contributes to the cost increase of the rectifier. However, if the two inductors are integrated onto one core this problem is overcome and the overall size of the integrated inductors is lower than in a center-tapped current-driven rectifier. Therefore, as a future work, it is recommended to evaluate the feasibility of a hybrid rectifier with the two inductor and the transformer integrated onto only one core. Moreover, the use of power MOSFETs with low on-resistance could improve the rectifier efficiency.

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