High-precision Current Source Using Low-Loss Single-Switch Three-Phase AC-DC Converter

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Abstract

A three-phase ac/dc converter based on isolated Cuk topology feeding an inductive load is presented. The main goal is to get a compact highly stable current source to feed an electromagnet.

A high power factor is achieved, at constant duty-cycle and switching frequency, by discontinuous input current mode operation. The converter presents a linear relationship between the duty-cycle and the output current, making easier to design the control system. Additionally the voltage stress on the power transistor is constant and does not depend on the duty-cycle.

An auxiliary circuit allows zero voltage turn-off while limiting the over-voltage on the switch produced by the transformer leakage inductance.

PWM control is used to reduce sensitivity to line disturbances and to eliminate the 300 Hz ripple on the output current.

Experimental measurements taken on a 400W prototype confirm theoretical forecasts.

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I. Introduction

DC current sources used to feed magnets in particle accelerators, electronic microscopes, etc., must feature high stability of the average current and very low ripple. Typically the maximum current deviation allowed is between 0.1% and 0.01%. Moreover, as far as line currents quality and EMI aspects are concerned, it is desirable to have high power factor and soft-commutations.

A three-phase ac/dc converter based on isolated Cuk topology is proposed, in which the output filter is substituted by the load (Fig. 1). Designing the input stage so as to operate in Discontinuous Input Current Mode (DICM), high input power factor is achieved at constant duty-cycle and switching frequency [1]. Moreover, a suitable auxiliary circuit shown in figure 3, provides both zero voltage turn-off and limitation of the over-voltage on the switch caused by the transformer leakage inductance.

To obtain smooth output current, PWM control is adopted, which allows excellent rejection to line disturbances.

This particular application presents some interesting characteristics as compared to voltage sources [2,3]: constant voltage stress on the switch (independent of the duty-cycle), and a linear relationship between duty-cycle and output current.

II. Main Converter Equations

A detailed analysis of the voltage-fed converter similar to that of figure 1 is reported in [2-3].

A similar analysis is carried out here taking into account that the output variable is now the load current. The converter is supposed to operate in DICM, while the output stage works in the continuous mode. This means that, during the switch off-interval, rectified current i_r decreases to zero while freewheeling diode D is always conducting. In the following analysis, the output current ripple is neglected, being very small compared to the DC value I_L.

Power stage design equations are derived first and then the behavior of the auxiliary circuit for soft commutation is analyzed.

In the converter equations δ is the duty-cycle, T is the switching period and V is the RMS line-to-line voltage.

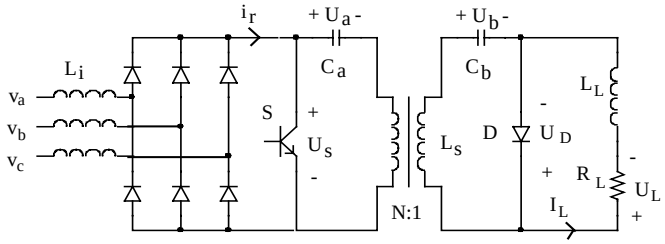

Fig. 1. Isolated Cuk converter with inductive load.

a) Switch voltage

The voltage applied to the main switch during the off stage, without the effect of the transformer leakage inductance is:

$$U_s = U_a + N U_b \tag{1}$$

b) Rectified input current

Let us consider the situation when 2 of the 3 input voltages have the same value. Two of the three input inductances are in parallel. In this situation the maximum line-to-line voltage is:

$$\hat{V}_l = \frac{V\sqrt{6}}{2} \tag{2}$$

The rectified current i_r coincides with the phase current corresponding to the maximum (in absolute value) phase voltage, and has the waveform shown in figure 2. In the particular instant considered the three phase currents goes to zero at the same time t_2 [2].

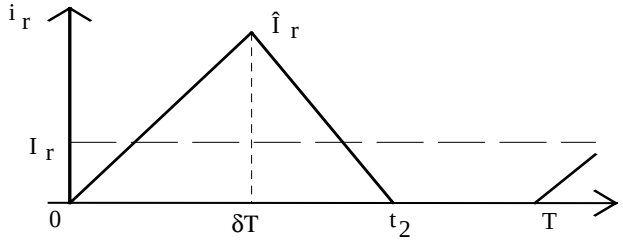

Fig. 2. Input current waveform

The peak and average current values are:

$$\hat{I}_r = \frac{\sqrt{6}\,V\delta T}{3L_i} \tag{3}$$

$$I_r = \frac{t_2\,\hat{I}_r}{2\,T} \tag{4}$$

$$t_2 = \delta T \cdot \left(\frac{2\,U_s}{2\,U_s - V\sqrt{6}} \right) \tag{5}$$

c) Output voltage

The voltage applied to the load during the switch on-interval δT is U_s/N. During the remaining part of the switching period the voltage is zero. Thus the output voltage is given by:

$$U_L = R_L I_L = \frac{U_s\,\delta}{N} \tag{6}$$

From the power balance and assuming 100% efficiency we can write:

$$P_L = U_L I_L = \hat{V}_1 I_r \tag{7}$$

From the above equations we get:

$$U_L = \frac{\sqrt{6}\,V\,\delta}{2\,N} + \frac{V^2\delta^2\,T}{2\,L_i I_L} \tag{8}$$

Even if this analysis has been carried out considering a particular time instant, it was already shown [2] that the accuracy of the output voltage value is better than 1% for any condition.

d) Output current

By substitution of (6) into (8) one gets:

$$I_L = \frac{\delta\,V\sqrt{6}}{4\,R_L N}\cdot\left[1+\sqrt{1+\frac{4\,R_L T\,N^2}{3\,L_i}}\right] \tag{9}$$

$$I_L = K\cdot\delta \tag{10}$$

$$K = \frac{V\sqrt{6}}{4\,R_L N}\cdot\left[1+\sqrt{1+\frac{4\,R_L T\,N^2}{3\,L_i}}\right] \tag{11}$$

Another interesting result is gotten substituting (10) into (6):

$$U_S = K\,R_L N \tag{12}$$

As compared to the voltage source converter [2,3], once the load current I_L is taken as controlled variable, we gain two main advantages: first the relationship between I_L and duty-cycle is linear, thus simplifying the control design; second the switch voltage stress is independent of the load current, making easier selection of the active device.

e) Minimum transformer secondary magnetizing inductance L_s

There is no current in the transformer primary side during the interval $(T-t_2)$. Considering $U_b=U_L$ (steady-state condition), the transformer secondary current varies linearly with a slope equal to U_b/L_s. If it rises to the output current level, the diode current goes to zero. In this situation an undesirable resonance between L_L+L_s and C_b imposes a low-frequency current component to the output. To avoid this a suitable minimum secondary inductance must be selected.

$$L_S \geq R_L T\cdot\left[\frac{2U_S(1-\delta)-V\sqrt{6}}{2U_S-V\sqrt{6}}\right] \tag{13}$$

f) Input inductances

These inductances must be selected in order to allow the DICM operation. The worst case is when the line current is maximum, which occurs when two of the three phase voltages are equal. The critical situation is when $T=t_2$. From (3), (4) and (7), the maximum input inductance is:

$$L_i \leq \frac{V^2\delta T}{2P_L} \tag{14}$$

g) Transformer turns ratio

From (8) it is possible to determine the transformer turns ratio, using the duty-cycle and the output current as parameters. As both secondary inductance and turns ratio are calculated, all the data needed to design the transformer are determined.

$$N = \frac{\delta I_L \sqrt{6}\, V}{2 R_L I_L^2 - \delta^2 V^2 \frac{T}{L_i}} \tag{15}$$

h) Energy transfer capacitor

The secondary side capacitance C_b reflected to the primary allows to define an equivalent capacitance:

$$C_o = \frac{C_a C_b}{C_a N^2 + C_b} \tag{16}$$

This capacitance is determined considering its peak-to-peak ripple voltage ΔU_s at full load:

$$C_o \geq \frac{I_L \delta T}{N \Delta U_s} \tag{17}$$

III. Auxiliary Circuit Operation: Ideal Case

The voltage stress typical of the Cuk converter imposes the use of a high-voltage switch. The use of MOSFETs would produce excessive conduction losses. On the other hand, IGBTs have significant turn-off losses, caused by the tail current effect.

The use of an auxiliary circuit for zero-voltage turn-off results in important reduction of switching losses [4]. This allows the use of IGBTs, even at high frequency (50 kHz), maintaining high converter efficiency. However, the turn on is still a hard-commutation even if the DICM operation reduces the switch current at turn on. A soft turn on could be obtained allowing discontinuous operation at the output [5], but for this specific application, as explained previously, it is not desirable.

Let us refer to figure 3 for an explanation of the auxiliary circuit operation, where L_d is the leakage inductance. The corresponding waveforms are shown in figure 4.

Let us suppose that voltage U_a does not vary appreciably during a switching period.

At the end of the T_{off} interval, rectified current i_r and resonant current i_1 are zero, voltage U_C is equal to the output voltage reflected to the primary side and output current flows through the freewheeling diode D.

When transistor S is turned on, capacitor C_C resonates with inductor L_C and the switch carries the sum of rectified current i_r, load current reflected to the primary side I_L/N and resonant inductor current i_l.

At instant T_1, voltage U_C reaches the value $-U_a$, diode D_1 starts to conduct and current i_l, which in the interval T_0-T_1 flowed through D_2 and S, now decreases to zero with a slope equal to $-U_a/L_C$ flowing almost entirely through diodes D_1 and D_2 ($C_a >> C_C$). At the end of T_{on} interval (instant T_2), the switch is turned off under zero voltage condition and capacitor C_C charges almost linearly to the value NU_L (instant T_3). Afterwards, freewheeling diode D starts to conduct. In the interval T_3-T_6 the rectified current i_r decreases to zero in a piecewise linear mode [2].

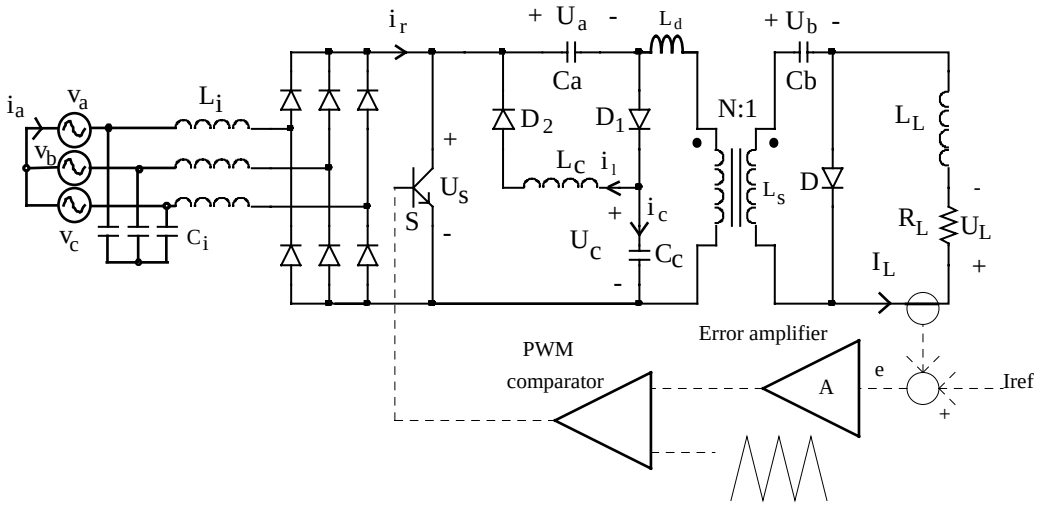

Fig. 3. Proposed topology with auxiliary circuit for soft-turn-off

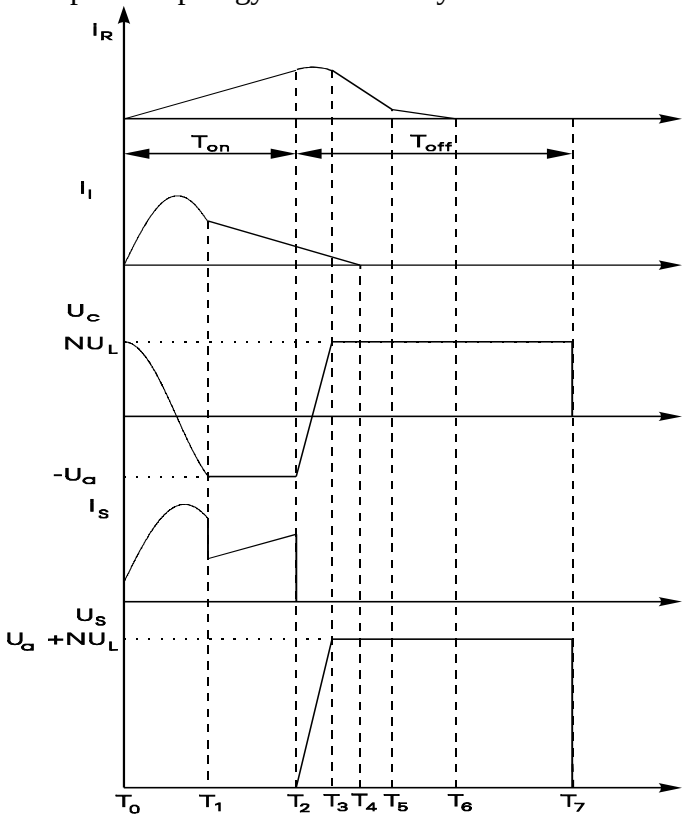

Fig. 4. Main converter waveforms: ideal case (without transformer leakage inductance)

Of course, the zero-voltage turn-off condition requires:

$$NU_L > U_a \tag{18}$$

which in terms of duty-cycle means: $\delta > 50\%$ (derived from 1) and 2)).

In practice, however, the transformer leakage inductance makes possible soft commutation also at lower duty-cycle, as shown in Figure 5, since capacitor C_C is charged at values higher than NU_L.

The actual voltage U_c depends on the impedance of the resonant circuit formed by the leakage inductance L_d and C_c:

$$U_C = U_L N + Z_c\left(\frac{I_L}{N} + \hat{I}_r\right) \tag{19}$$

$$Z_c = \sqrt{\frac{L_d}{C_c}} \tag{20}$$

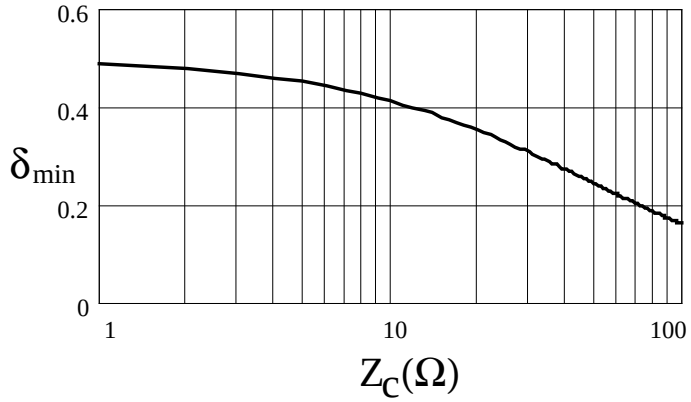

Fig. 5. Minimum duty-cycle for soft-turn-off.

IV. Control Strategy

The circuit uses a PWM controller. In order to ensure the desired mean current precision, the gain of the current error amplifier has to be higher than the inverse of the required precision. Variations in input voltage or load resistance will cause changes in the duty-cycle, in order to maintain constant the load average current. Due to the three-phase rectification the output current contains a ripple component at six time the line frequency which must be eliminated by the control. The small duty-cycle modulation needed to compensate it does not affect significantly the input current waveform and the power factor, as confirmed by the experimental results.

It is worthy to note that the DICM operation makes easier the controller design, simplifying the converter transfer function.

V. Simulation and Experimental Results

Figure 6 shows the measured static characteristic of the converter compared to the theoretical expectation. The small error at high current is due to the fact that the duty-cycle "seen" by the load is wider than the transistor one. The reason is that the output diode starts to conduct only after the voltage on C_c reaches NU_L. This phenomenon is less important at low current.

Experimental results are taken on a 400W prototype with the following parameters:
Input voltage: 220 V (line-to-line RMS value)
Switching frequency: 50 kHz
Output current: 10 A
Load: 4 mH, 4 Ω
C_a: 1 μF, C_b: 56 μF, C_c: 20 nF, C_i: 330nF

L_i: 330 µH, L_s: 50 µH, L_c: 160 µH
N: 7.4

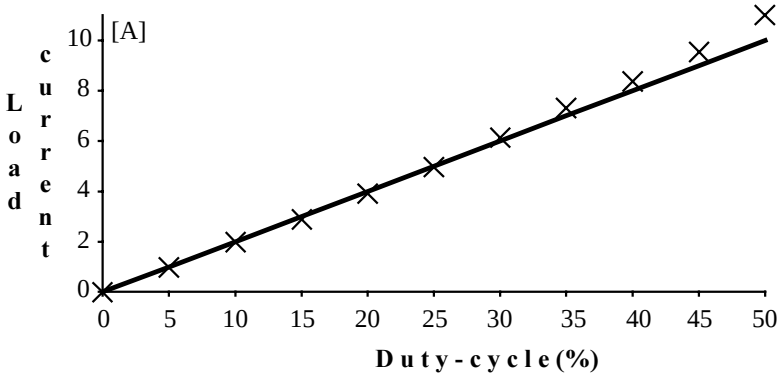

Fig. 6 Static characteristic of the converter

In Figure 7 is shown the average and RMS current through the switch. This current is composed by three components: the input current, the load current reflected to the primary side and the resonant current.

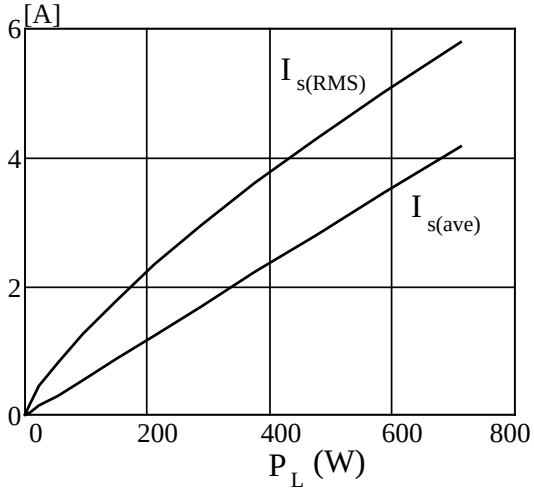

Figure 7. Average and RMS current through switch S.

As mentioned, due to DICM operation, each phase current zeroes every switching cycle. An unfiltered line current waveform is shown in figure 8, together with the corresponding phase voltage in the case of constant duty-cycle (open loop). As stated above, DICM operation ensures high power factor.

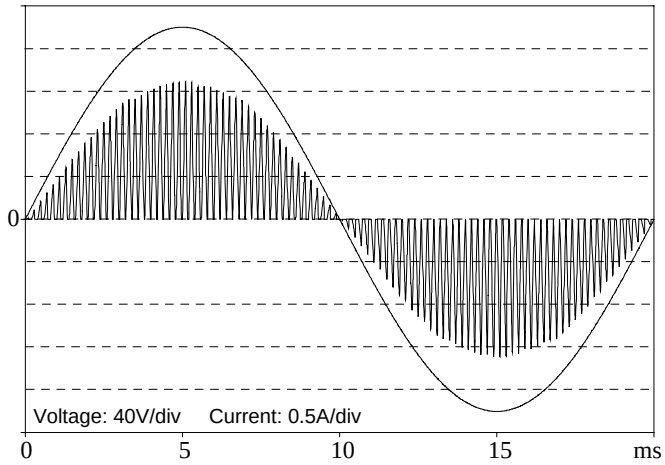

Fig. 8. Simulated input phase voltage and current (unfiltered)

Figure 9 shows transistor current and voltage. Soft turn-off is achieved even at duty-cycle equal to 40%. The 25% overvoltage is expected considering the transformer leakage inductance and the auxiliary capacitor selected. The IGBT tail current is clear but the respective low U_S voltage reduces significantly the associated loss.

Figure 10 shows the current and voltage on the auxiliary capacitor, C_c. An almost constant current charges the capacitor, while the discharge obeys a resonant law.

Figure 11 shows the effect of the control on the output current ripple. The PWM control changes the duty-cycle in order to maintain constant the output voltage (and the output current), rejecting the 300 Hz ripple.

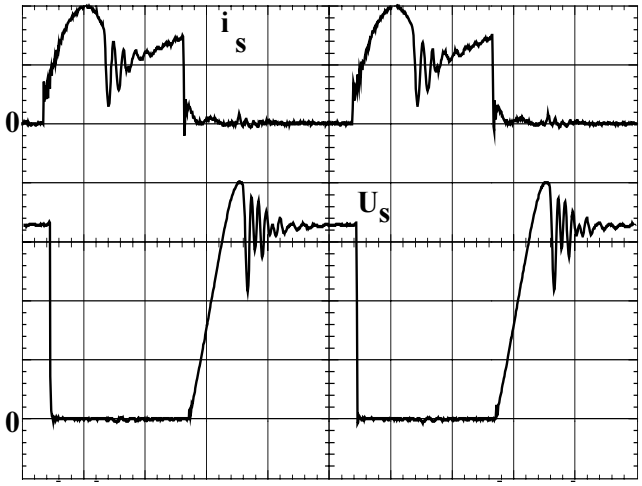

Fig. 9. Top trace: Transistor current (2A/div.)
Bottom trace: Transistor Vce voltage (200V/div.). Horz.: 4μs/div

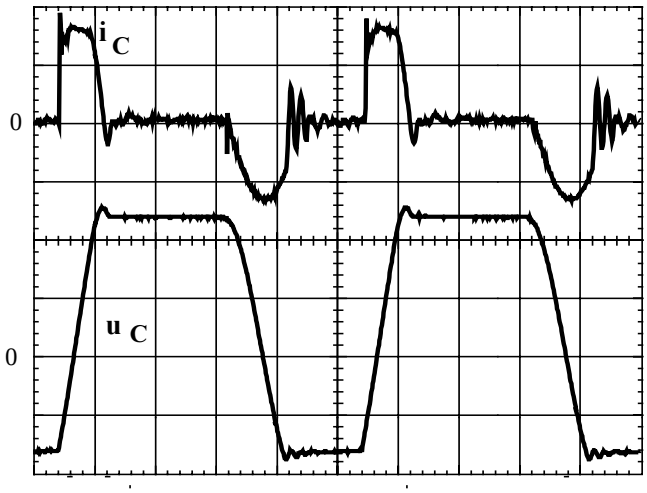

Fig. 10. Capacitor C_c current (top) (5A/div) and voltage (bottom) (200V/div) Horiz.: 4μs/div

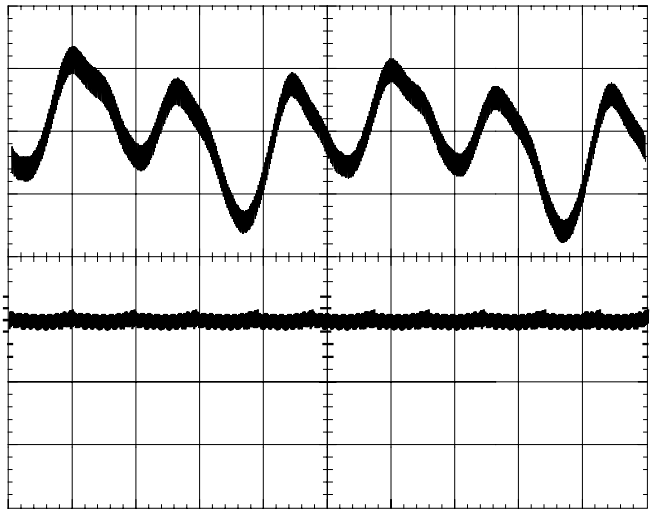

Fig. 11. Output current ripple (20 mA/div)
Top trace: constant duty-cycle; Bottom trace: PWM control. Horz.: 2 ms/div

Figure 12 shows one phase voltage and current (filtered). The duty-cycle modulation produced by the PWM controller do not affect significantly the current waveform, allowing a high power factor.

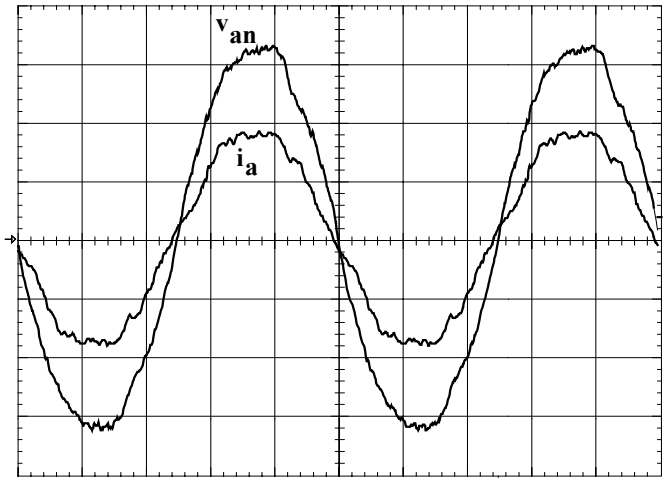

Fig. 12. Line voltage (50V/div) and current (1A/div) Horz.: 4ms/div

The relative ripple is $\pm$ 0.1%. Considering the additional attenuation caused by the magnet core and metallic vacuum chamber (where the beam is enclosed) the expected magnetic field ripple is about tens of ppm. The gain of the error amplifier is 1000, ensuring the maximum average current error of 0.1%.

The measured power factor is 0.98, including the input capacitive filter. The current THD is 5.8% for a line voltage THD of 3.4%. The efficiency is 90%.

VI. CONCLUSIONS

The proposed solution for feeding inductive loads has some interesting properties: limited voltage stress on the main switch; linear current over duty-cycle ratio; and high power factor.

The auxiliary circuit needed to avoid voltage spikes caused by the transformer leakage inductance was designed in order to additionally allow transistor zero-voltage turn-off, resulting in high efficiency even at high switching frequency.

Lastly, PWM control compensates very quickly for line disturbances and also provides fast dynamic response, rejecting almost completely the voltage ripple. The duty-cycle modulation does not affect significantly the input current waveform.

Measurements taken on a prototype confirmed theoretical expectations.

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Fig. 1. Isolated Cuk converter with inductive load.
Fig. 2. Input current waveform.
Fig. 3. Proposed topology with auxiliary circuit for soft-turn-off
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Fig. 7. Average and RMS current through switch S.
Fig. 8. Simulated input phase voltage and current (unfiltered).
Fig. 9. Top trace: Transistor current (2A/div.). Bottom trace: Transistor Vce voltage (200V/div.). Horz.: 4μs/div.
Fig. 10. Capacitor C_c current (top) (5A/div) and voltage (bottom) (200V/div) Horiz.: 4μs/div.
Fig. 11. Output current ripple (20 mA/div). Top trace: constant duty-cycle; Bottom trace: PWM control. Horz.: 2 ms/div.
Fig. 12. Line voltage (50V/div) and current (1A/div) Horz.: 4ms/div.

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Biographies

José Antenor Pomilio (M'93) was born in Jundiaí, Brazil, in 1960. He received the B. Sc. (1983), the M. Sc. (1986) and the D. Sc. (1991) degrees in Electrical Engineering from the University of Campinas (UNICAMP), Campinas, Brazil.

He began work as Instructor at the Electrical Engineering Department at UNICAMP in 1984. From 1988 to 1991 he was head of the Power Electronics Group at the Brazilian Synchrotron Laboratory. Since 1991 he is an assistant professor at the School of Electrical Engineering of the University of Campinas. In 1993/1944 he held a post-doctoral position at the Electrical Engineering Department of the University of Padova, Italy.

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Dr. Pomilio is a member of IEEE PELS, IES and IAS. He is currently coordinator of the Power Electronics Committee of the Brazilian Society of Automatic Control (SBA) and member of the Brazilian Society of Power Electronics (SOBRAEP).

Giorgio Spiazzi (S'92) was born in Legnago (province of Verona, Italy) in 1962. He graduated with honors in Electronic Engineering at the University of Padova in 1988.

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Mr. Spiazzi is a member of IEEE.