Soft-Commutated Cuk and SEPIC Converters as Power Factor Preregulators

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Abstract

Cuk and SEPIC converters operating in Discontinuous Current Mode (DCM), used as Power Factor Preregulators (PFP), present some advantages over other topologies: sinusoidal current shaping using constant frequency and duty-cycle, low input current harmonic content, low input filter requirements, isolation between line and load, etc. The main drawbacks are the voltage and current stresses that devices are submitted to. In this paper, an auxiliary circuit is proposed which limits the overvoltage on the main switch and allows its zero-voltage turn off. Zero-current turn on is provided by the DCM operation. In this way switch commutations are soft, thus improving efficiency and reducing RF noise. Design criteria and experimental results are reported.

INTRODUCTION

Recently, power converters with high power factor and sinusoidal input current have gained considerable attention, due to international standards regarding the line pollution. Various topologies can be used to perform this task, the most diffused being the boost one. When output isolation is required as well as over-current and start-up protection, Cuk and SEPIC converters are good solutions. Moreover, when operating in Discontinuous Current Mode (DCM) [1-3], they offer other advantages: low harmonic input current without filtering, operation at constant frequency and duty-cycle, no need of current loop, soft turn-on of the main switch. The main drawbacks are the high current and voltage stresses on the switches. Voltage stresses can be further worsened by the effects of transformer leakage inductance, calling for devices with high-voltage capability. IGBTs can be used, which have less forward voltage drop and are cheaper than MOSFETs. But, because of the turn-off tail current, higher commutation losses can be expected, which limit the switching frequency.

In [4], a lossless clamper circuit was used in order to limit the overvoltage due to the transformer leakage inductance. Here, an improved auxiliary circuit is presented, which allows zero-voltage turn-off of the transistor. This means that all commutations are soft, including that of the switches in the auxiliary circuit. The use of soft-switching techniques also limits the RF noise generated, which is a non negligible aspect in preregulators [5].

Cuk converter

The converter shown in figure 1.a is the insulated version of the Cuk converter driven by a rectified sinusoidal voltage \( v_g \). All the converter components are rated on the switching frequency basis, except the output capacitor \( C_o \), which must be big enough to filter the low frequency ripple, at twice the line frequency, caused by the input power variation when operating as PFP [1]. The capacitor voltages are:

\[
\begin{align*}
V_C &= V_{V_g} - V_L \\
V_L &= V_C - V_{V_g} \\
V_{V_g} &= V_{V_l} = V_L + V_C \\
V_C &= V_{V_l} = V_{V_g} - V_L
\end{align*}
\]

When operating in DCM, the current waveforms of both inductors look as shown in figure 2 [6]. The circuit behavior is as follows: when the switch is turned on, the freewheeling diode is reverse biased and the currents rise linearly (assuming small

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high-frequency voltage ripple across capacitors). During turn-off interval, both inductor currents decrease until their sum becomes zero, causing the freewheeling diode turn off. After that, they remain constant for the rest of the switching period.

The average input current is given by [2]:

\[
i(\theta) = \frac{\delta^2}{2} \cdot T 
\]

where \( T \) is the switching period and

\[
L_e = \frac{N^2 \cdot L_1 \cdot L_2}{L_1 + N^2 \cdot L_2}
\]

This relation shows that, for constant switching frequency and duty-cycle, the average input current is proportional to the input voltage, ensuring unity power factor.

\[
\begin{align*}
\text{Limits for DCM operation. Let us define the following adimensional parameters \cite{1,3,6}:} \\
M &= \frac{V_i}{V_o} & \text{Voltage conversion ratio} \\
K_e &= \frac{2 \cdot L_e}{R_L \cdot T} \\
\text{The converter duty-cycle is given by:} \\
\delta &= M \cdot \sqrt{K_e} \\
\text{which shows that, like any other converter operating in DCM, the} \\
\text{voltage conversion ratio} M \text{ depends on the load.} \\
\text{In order to ensure the DCM operation in all conditions,} \\
\text{the converter parameters must satisfy the following inequality:} \\
K_e < \frac{N^2}{2 \cdot (M \cdot N + 1)^2} \\
\text{Once the value of equivalent inductance} L_e \text{ is known, from} \\
\text{the desired relative input current ripple} \tau \text{ (peak to peak),} \\
\text{the input inductance} L_1 \text{ is obtained as follows:} \\
L_1 &= \frac{2L_e}{\tau} \\
\text{With reference to figure 2, it is important to note that current} \tau' \text{ must be greater than zero, otherwise the converter operation changes:} \\
\text{in fact, current} i_{L1} \text{ cannot go negative for the presence of the diode} \\
\text{bridge. This latter constrain imposes a minimum value for} L_1, \text{ that is:} \\
L_1 > \frac{L_2 \cdot N}{M} \\
\text{SEPIC Converter} \\
\text{For this converter, whose scheme is reported in figure 1.b, the} \\
\text{analysis is analogous to the Cuk one, with the difference that the} \\
\text{current} i_{L1} \text{ is not affected by the transformer turns-ratio. The} \\
equations are the same except (4), and (10) which become respectively:} \\
L_e = \frac{L_1 \cdot L_2}{L_1 + L_2} \\
L_1 > \frac{L_2}{M \cdot N}
\end{align*}
\]

\[
\text{AUXILIARY CIRCUIT FOR SOFT TURN-OFF}
\]

\[
\text{In the above PFPs the insulation transformer causes severe} \\
\text{voltage spikes and ringing due to its leakage inductance, when} \\
\text{the current in the primary side reverses.} \\
\text{Figure 3 shows Cuk and SEPIC converters with the proposed} \\
auxiliary circuit, which allows zero-voltage turn-off of the main} \\
\text{switch while limiting its overvoltage.}
\]

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\text{Auxiliary Circuit}
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\text{Auxiliary Circuit}
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\text{Auxiliary Circuit}
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\[
\text{Figure 2. Inductor voltages and currents in DCM for the Cuk} \\
\text{converter}
\]

\[
\text{Figure 3. Cuk (a) and SEPIC (b) converters with auxiliary circuit for} \\
\text{soft-commutation}
\]

\[
\text{Auxiliary circuit behavior} \\
\text{Figure 4 shows the simulated converter waveforms during one} \\
\text{switching period. The same driving signal is applied to both} \\
\text{switches.} \\
\text{At the end of} t_{\text{off}}, \text{the voltage on} C_c \text{ is equal to the} \\
\text{reflected output voltage plus the overvoltage caused by the} \\
\text{transformer leakage inductance. The freewheeling diode} D \text{ is off.} \\
\text{At instant} t_0, \text{both switches are turned on under zero-current condition.}
\]
Figure 4. Converter and auxiliary circuit waveforms during a switching period

Interval $t_0$-$t_1$. Capacitor $C_c$ starts resonance with inductor $L_c$. At instant $t_1$, its voltage reaches the value $-v_a$ and $D_{c1}$ starts conducting.

Interval $t_1$-$t_3$. The inductor current $i_{Lc}$ divides between $C_a$ and $C_c$, flowing almost entirely through $S_c$, $D_{c2}$, $S$, $C_a$ and $D_{c1}$ ($C_a > C_c$), and decreases linearly to zero (instant $t_2$) with a slope equal to $-v_a/L_c$. Note that the auxiliary circuit does not cause increased voltage or current stresses in the main switch. On the contrary, the total current carried by the main switch is lowered.

Interval $t_3$-$t_4$. At $t_3$, the main switch is turned off under zero-voltage condition, and the auxiliary one under zero-current condition. The current flows through $D_{c1}$ to recharge capacitor $C_c$. As the current during this small interval is almost constant, the voltage varies linearly. When $V_c$ equals the output voltage reflected to the primary side, $D$ starts to conduct (interval $t_4$).

Interval $t_4$-$t_5$. Voltage $V_c$ increases above the reflected output voltage in a resonant manner due to the oscillation between $L_d$ and $C_c$. The corresponding overvoltage depends on the inductor current value and the characteristic impedance of the resonant circuit. The duration of this interval is one fourth of the resonant period.

Interval $t_5$-$T$. At $t_5$, diode $D_{c1}$ stops conduction and current $i_{L1}$ flows through the transformer until the current in the freewheeling diode becomes zero (operation in discontinuous mode).

Interval $t_6$-$T$. The output diode opens at zero current. The voltage reflected to the primary becomes zero and the voltage across the main switch drops to $v_a$.

The value of this capacitor is chosen considering the allowed switch voltage stress. The maximum voltage across $C_c$, taking into account the effect of transformer leakage inductance, can be estimated as:

$$V_c = NV_L + \frac{V}{L_c} \delta T \cdot Z_d$$

Accordingly, the switch voltage stress is:

$$\bar{V}_s \approx V_c + \bar{V}_c$$

It is important to note that, due to the charging process of $C_c$, output diode turn on is delayed (interval $t_3$-$t_4$). This means an effective increase in the duty-cycle "seen" by the output. The delay increases at light load and near the input voltage zero-crossing. This phenomenon does not affect the converter behavior as power factor preregulator but imposes an upper limit for the value of $C_c$.

Inductor $L_c$

Inductance $L_c$ is a critical parameter because it must be chosen in order to allow zeroing of its current during the minimum switches on time, i.e., with reference to figure 4:

$$t_a + t_b < \delta \cdot T$$

(15)

It is not possible to obtain an exact analytical expression for this inductance, but an approximate worst-case result can be obtained. During interval $t_d$, $C_c$ and $L_c$ resonate until the voltage across $C_c$ reaches $-v_a$. Knowing the voltage $V_{cmax}(\theta)$ across $C_c$ before the switch turn-on, interval $t_a$ is given by:

$$t_a = \frac{1}{\omega_c} \arccos \left( -\frac{1}{\gamma(\theta)} \right)$$

(16)

where

$$\gamma(\theta) = \frac{V_{cmax}(\theta)}{V_{c}(\theta)}$$

(17)

At $t_1$, current $i_{Lc}$ starts to decrease almost linearly to zero under the action of voltage $v_a$. Thus, we can write:

$$t_b = \frac{L_c \cdot I_{c}(\theta)}{v_{c}(\theta)}$$

(18)

where $I_{c}(\theta)$ is the value of current $i_{Lc}$ in the instant $t_1$, and is given by:

$$I_{c}(\theta) = \frac{V_{cmax}(\theta)}{Z_c} \sin(\omega_c t_1) = \frac{v_{c}(\theta)}{Z_c} - \sqrt{\gamma(\theta)^2 - 1}$$

(19)

As we can see, in order to find the maximum value of interval $t_a$+$t_b$, we must know the behavior of voltage $V_{cmax}(\theta)$. For this purpose note that near the line voltage zero crossing, since both $i_{L1}$ and $i_{L2}$ are low, there is not enough energy to charge $C_c$ to the value $NV_L$, so no resonance occurs between the transformer leakage inductance and $C_c$ (the output diode does not enter conduction). Thus, we must find the interval, during the line period, in which this situation occurs. The equivalent circuit during interval $t_3$-$t_4$ is shown in figure 5.

![Figure 5. Equivalent circuit at turn off (interval $t_3$-$t_4$)](image-url)

The initial conditions are:

Before to go into detail, let us define the following parameters:

$$\omega_d = \frac{1}{\sqrt{L_d C_c}}, \; Z_d = \frac{L_d}{\sqrt{C_c}}$$

$$\omega_c = \frac{1}{\sqrt{L_c C_c}}, \; Z_c = \frac{L_c}{\sqrt{C_c}}, \; \omega_e = \frac{1}{\sqrt{L_e C_c}}$$

**Selection of Auxiliary Circuit Parameters**
\[ i_e(0) = \frac{v_g(\theta)}{L_c} \delta T, \quad v_c(0) = -v_g(\theta) \]

Voltage \( v_c \) reaches its maximum when \( i_e \) zeroes:

\[ v_{c_{\text{max}}}(\theta) = v_g(\theta) \left[ \delta T \omega_c \cdot \sin(\alpha) - \cos(\alpha) \right] \]

where \( \alpha = \pi + \arctg(-\delta T \omega_c) \).

From the relation

\[ V_{c_{\text{max}}}(\theta) = NV_L \]

we find the value of angle \( \theta_1 \) below which \( V_{c_{\text{max}}}(\theta) \) is given by (20):

\[ \theta_1 = \arcsin \left[ \frac{N \cdot V_L}{V_i} \left( \delta T \cdot \omega_c \cdot \sin(\alpha) - \cos(\alpha) \right) \right] \]

For \( \theta_1 < \theta < \pi - \theta_1 \), the freewheeling diode D starts conduction during the \( t_{\text{off}} \) interval allowing resonance between \( L_d \) and \( C_c \) to occur. Note however, that for angles \( \theta \) slightly greater than \( \theta_1 \) the overvoltage on \( C_c \) is not maximum, since the overshoot is interrupted by the end of turn-off period (interval \( t_4-t_5 \) is less than one forth of resonant period). At higher angles, \( v_c \) is maximum and is given by:

\[ V_{c_{\text{max}}} = \frac{V_{N}L}{L_c} + \frac{V_g(\theta)}{L_c} \delta T \cdot Z_d \]

Figure 6 shows typical behavior of voltage \( V_{c_{\text{max}}} \), normalized to the reflected output voltage, versus input voltage angle.

For \( \theta_1 < \theta < \pi - \theta_1 \), the zero-voltage turn-off condition requires a voltage across \( C_c \) always greater than \( V_a \), i.e.:

\[ V_{c_{\text{max}}}(\theta) \geq v_g(\theta) \]

Using these results, the typical variation of interval \( t_{\text{a}+t_{\text{b}}} \), normalized to the duty-cycle, is shown in figure 7. Note that the worst case occurs just after \( \theta_1 \) (or before \( \pi - \theta_1 \)). This fact suggests a method for estimating \( L_c \), considering that the most critical situation occurs at angle \( \theta_1 \) and, at this point, there is the total overvoltage across \( C_c \).

DESIGN PROCEDURE

In the converter design procedure, the following input data are considered:

- peak input voltage \( V_i \)
- output voltage \( V_L \)
- maximum and minimum load resistance \( R_{L_{\text{max}}}, R_{L_{\text{min}}} \)
- switching frequency \( f=1/T \)
- maximum switch voltage \( V_S \)
- relative input current ripple \( r_i \)

Power stage design

In the following, the detailed power stage design procedure is outlined using the results derived in the previous paragraphs.

1) Choose a suitable maximum value for the duty-cycle \( \delta_{\text{max}} \);
2) calculate M from (5);
3) calculate N taking the equality in (25);
4) find the value of \( K_e \) from (7) using the minimum load resistance and verify condition (8). If it is not satisfied, a different value for \( \delta_{\text{max}} \) must be chosen;
5) knowing \( K_e \), calculate \( L_e \) from (6);
6) calculate \( L_1 \) from (9);
7) calculate \( L_2 \) from (4);
8) check condition (12); if it is not satisfied, a lower input current ripple must be considered.

Capacitor \( C_a \) and \( C_b \) are selected taking into account the desired high frequency voltage ripple. As far output capacitor is concerned, it must be rated at line frequency instead of switching frequency. So, given the desired relative voltage ripple \( r_v \) (peak to peak), we can write:
\[ C_o = \frac{1}{\alpha_1 r_i R_{L_{\text{min}}}} \]  

(27)

**Auxiliary circuit design**

From the above considerations, the choice of auxiliary circuit parameters can be summarized as follows:

1) from the maximum voltage stress allowed on the switch and (13-14), calculate the value of capacitance \( C_c \) and verify inequality (26);
2) calculate the minimum duty-cycle in correspondence of maximum load resistance from (7);
3) calculate \( \theta_1 \) from (21) at minimum duty-cycle;
4) find the value of parameter \( \gamma(\theta_1) \) from (17) and (22);
5) calculate the value of inductance \( L_c \) from (23) using the minimum duty-cycle.

**EXPERIMENTAL RESULTS**

A 250W converter prototype was built according to the following specifications:

\[ V_i = 120\sqrt{2} \text{ V} \]
\[ V_L = 50 \text{ V} \]
\[ R_{L_{\text{max}}} = 62.5 \text{ \Omega}, \quad R_{L_{\text{min}}} = 10 \text{ \Omega} \]
\[ f = 50\text{kHz} \quad (T = 20 \mu\text{s}) \]
\[ V_S = 800\text{V} \]

Input current ripple: 30%

Following the design procedure outlined above, the converter parameters are:

- Transformer turns ratio: 5:1
- Measured transformer leakage inductance: \( L_d = 13 \mu\text{H} \).
- Nominal duty-cycle: \( \delta = 0.5 \) \( (P_o = 250\text{W}) \)
- Minimum duty-cycle: \( \delta = 0.2 \) \( (P_o = 40\text{W}) \)
- \( L_1 = 1.9 \text{ mH} \)
- \( L_2 = 6.2 \mu\text{H} \)
- \( L_c = 30 \mu\text{H} \)
- \( C_a = 500 \text{ nF} \)
- \( C_b = 50 \mu\text{F} \)
- \( C_o = 1000 \mu\text{F} \)
- \( C_c = 20 \text{ nF} \)

Figure 8 shows line current and voltage. The distortion in the current near the zero crossing is due to the limitation imposed by the inductance \( L_1 \). As the inductance is relatively high, at low input voltage the current rate of change is limited. Nevertheless, a high power factor is obtained. At full power, the measured value is 0.99.

Figure 9 displays the rectifier input current showing its high-frequency components. The expected ripple of 30% is got. The spectrum shows the high-frequency harmonics.

Figure 10 shows the behavior of the auxiliary circuit. Note that the charge process happens at constant current and the inversion of the voltage obey a sinusoidal variation.

Figure 11 shows the main switch current and voltage waveforms. The zero-current turn-on and zero-voltage turn-off are clear. The IGBT current tail is present and the use of zero-voltage turn-off contributes to reduce the losses [7].

Figure 12 shows the variation of the efficiency and power factor. The measured power factor at minimum load is 0.98 and higher than 0.99 at nominal load. The efficiency is relatively low specially because the high RMS currents present in the secondary side, which increase the resistive losses. Figure 13 reports the calculated converter losses at nominal power: as we can see, the auxiliary circuit losses are 24% of total ones. They weight substantially on the converter efficiency, but it is important to note that they are not proportional to output power. Thus, higher efficiency can be expected at higher power rating.
The use of the proposed auxiliary circuit, besides limiting the overvoltage, allows a full soft-commutation converter. A higher switching frequency is therefore possible also with IGBTs, maintaining a reasonably high efficiency. In spite of some low-frequency distortion in the line current, the power factor is next to unity and the current harmonics are small.

REFERENCES


