

APPLICATION OF SLIDING MODE CONTROL TO SWITCH-MODE POWER SUPPLIES

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Switch-mode power supplies represent a particular class of variable structure systems (VSS). Thus, they can take advantage of non-linear control techniques developed for this class of systems. In this paper the so called sliding mode control is reviewed and its application to switch-mode power supplies is discussed.

Sliding mode control extends the properties of hysteresis control to multi-variable environments, resulting in stability even for large supply and load variations, good dynamic response and simple implementation.

Application to dc-dc converters, as well as rectifiers and inverters, is analyzed and provisions to overcome the inherent drawbacks of sliding mode control, i.e. variable switching frequency and possible steady-state errors, are described. Experimental results are also reported, which allow a comparison between the sliding mode approach and other standard control techniques, e.g. current-mode control, showing its effectiveness.

1. Introduction

Control of switch-mode power supplies can be difficult, due to their intrinsic non linearity. In fact, control should ensure system stability in any operating condition and good static and dynamic performances in terms of rejection of input voltage disturbances (audiosusceptibility) and effects of load changes (output impedance). These characteristics, of course, should be maintained in spite of large input voltage, output current, and even parameter variations (robustness).

A classical control approach relies on the knowledge of a linear small-signal model of the system to develop a suitable regulator.¹ The design procedure is well known, but is generally not easy to account for the wide variation of system parameters, because of the strong dependence of small-signal model parameters on the converter operating point. This aspect becomes even more problematic in rectifiers and/or inverters in which the operating point moves continuously, following the periodic input/output voltage variations.

Multiloop control techniques, like current-mode control, have greatly improved power converters behavior but the control design remains difficult for high-order topologies, like those based on Cuk and Sepic schemes.^{2,3}

A different approach, which complies with the non-linear nature of switch-mode power supplies, is represented by the sliding mode control, which is derived from the variable structure systems (VSS) theory.^{4,5} This control technique offers several advantages: stability even for large supply and load variations, robustness, good dynamic response and simple implementation. Its capabilities emerge especially in application to high-order converters, yielding improved performances as compared to classical control techniques.

In this paper, the sliding mode control is reviewed and its applications to switching power supply are investigated. Experimental results of different converter structures demonstrate the superior performances of this non-linear control technique as compared to standard control approaches.

2. Control of Variable Structure Systems: The Sliding Mode Approach

VSS are systems whose physical structure is changed intentionally during the time in accordance with a preset structure control law. The instants at which the changing of the structure occur are determined by the current state of the system.

From this point of view, switch-mode power supplies represent a particular class of VSS, since their structure is periodically changed by the action of controlled switches and diodes.⁶

2.1. An example: buck converter

To the purpose of explanation, consider the simple buck converter shown in Fig.1.⁷ In the continuous conduction mode operation, the converter structure periodically changes by the action of the controlled switch S, giving rise to the two substructures shown in Fig.1. Considering as state variables the output voltage error δu_o and its derivative, the system behavior can be described by the following equations

$$\begin{cases} \dot{x}_1 = x_2 \\ \dot{x}_2 = -\frac{x_2}{RC} - \frac{x_1}{LC} + \frac{1}{LC}(U_i \cdot \sigma - U_o^*) \end{cases} \quad (1)$$

in which σ is a discontinuous variable equal to 1 when the switch is ON and zero when the switch is OFF, and U_o^* is the output voltage reference.

The phase trajectories corresponding to different values of control variable σ are shown in Fig.2. In both cases, the system evolution is a damped oscillation which starts from a point representing the system initial conditions and reaches an equilibrium point given by $x_1 = U_i - U_o^*$ and $x_2 = 0$ for the case $\sigma = 1$, and given by $x_1 = -U_o^*$ and $x_2 = 0$ for the case $\sigma = 0$. Note, however, that in this latter condition, the inductor current cannot become negative due to the presence of the freewheeling diode. Thus, when this current reaches zero, the output capacitor is discharged to zero by the load resistance only, giving rise to a linear phase trajectory.

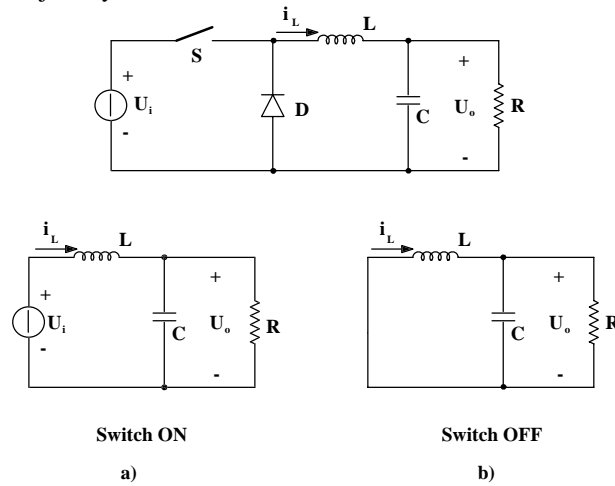


Fig. 1 - DC-DC buck converter; a) subtopology during switch on-time, b) subtopology during switch off-time.

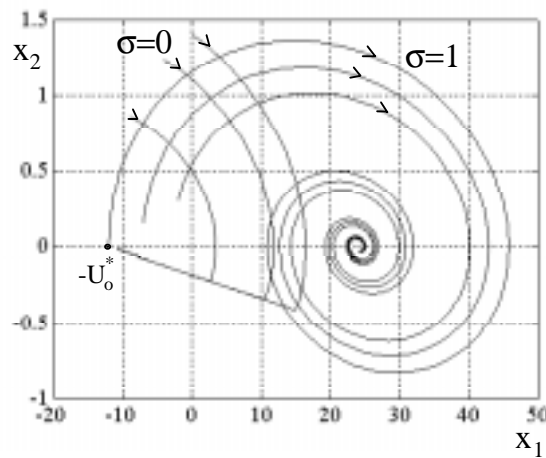


Fig. 2 - Phase trajectories corresponding to eq. (1) for different initial conditions.

2.2. Sliding mode control

Let us define the following function

$$\psi = x_1 + \tau \cdot x_2, \quad (2)$$

which is a linear combination of the two state variables. In the phase plane, equation $\psi=0$ represents a line, called *sliding line*, passing through the origin (which is the final equilibrium point for the system) with a slope equal to $-1/\tau$.

We now define the following control strategy

$$\begin{aligned} \text{if } \psi > +\beta &\Rightarrow \sigma = 0 \\ \text{if } \psi < -\beta &\Rightarrow \sigma = 1, \end{aligned} \quad (3)$$

where β defines a suitable hysteresis band. In this way, the phase plane is divided in two regions separated by the sliding line, each associated to one of the two subtopologies defined by the switch status σ . Let us suppose that the system status is in P, as shown in Fig.3. Since we are in the region $\psi < -\beta$, the switch is close and the motion occurs along a phase trajectory corresponding to $\sigma=1$. When the system status crosses the line $\psi=+\beta$, according to (3), the switch is turned off and the system status follows a phase trajectory corresponding to $\sigma=0$. Observing that the phase trajectories, in proximity of the sliding line, are directed toward the line itself, the resulting motion is made by continuous commutations around the sliding line, so that the system status is driven to the final equilibrium point.

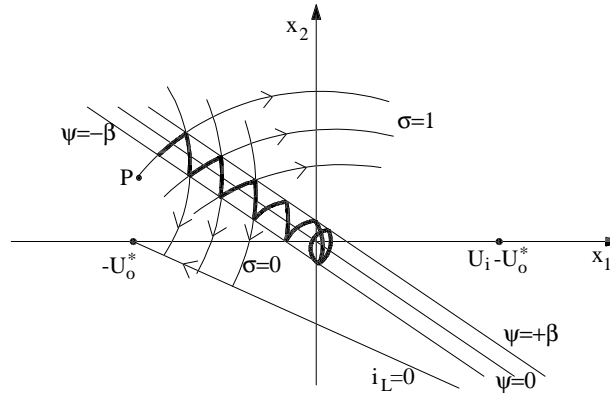


Fig. 3 - Sliding motion for the buck converter.

From this example, in the hypothesis of a suitable small value of β , two important conclusions arise

- I: when the system is in the sliding mode, its evolution is independent of the circuit parameters. It depends only on the sliding line chosen. In the example shown in Fig.3 the dynamic is of the first order with a time constant equal to τ .
- II: if N is the order of the original system, the dynamic of the controlled system in sliding mode has order $N-1$, since the state variables are constrained by the equation $\psi=0$.

Note that the switching frequency is determined by the amplitude of the hysteresis band β .

The potentialities of this control technique in the application to switch-mode power supplies are now evident: it exploits the intrinsic non-linear nature of these converters and it is able to provide dynamic behaviors that are different from that of the substructures composing the system, and correspond to that of a reduced order system.

2.3. Conditions for sliding motion

In the simple case of the second-order system considered in the previous section, sliding mode control design requires only selection of parameter τ . Selection, must be done in order to ensure the following three constraints

- (i) the *hitting condition*, which requires that the system trajectories cross the sliding line irrespective of their starting point in the phase plane;
- (ii) the *existence condition*, which requires that the system trajectories near the sliding line (in both regions) are directed toward the line itself;

- (iii) the *stability condition* of the system motion on the sliding line (i.e. the motion must be toward the equilibrium point).

Another important consideration concerning physical systems is that not all points in the phase plane are reachable. In the previous example, since inductor current and output voltage are always non negative, the regions under the line $x_2 = -(x_1 + U_o^*)/RC$ and to the left of line $x_1 = -U_o^*$ are left out.

3. Application of Sliding Mode Control to DC-DC Converters

3.1. Principle scheme

A basic sliding mode control scheme of dc-dc converters is shown in Fig.4.

Let U_i and u_{CN} be input and output voltages, respectively, i_{L_i} and u_{C_j} ($i=1-r, j=r+1-N-1$) the internal state variables of the converter (inductor currents and capacitor voltages), and N the system order. According to the theory, all state variables are sensed, and the corresponding errors x_i (defined by difference to the steady-state values) are multiplied by proper gains K_i and added together to form the sliding function ψ . Then, hysteretic block HC control the switch so as to maintain function ψ near to zero, thus we can write

$$\psi = \sum_{i=1}^N K_i \cdot x_i = \mathbf{K}^T \mathbf{x} = 0. \quad (4)$$

where $\mathbf{K}=[K_1, K_2, \dots, K_N]^T$ is the vector of sliding coefficients (T means transposition).

This equation represents a hyperplane in the phase space passing through the origin (note that (2) is a particular case of (4) for $N=2, K_1=\tau/C$ and $K_2=1$).

This general scheme, although interesting in theory, is not practical. In fact, a control approach representing an effective alternative to standard control techniques must have the following characteristics

- (i) simple design and implementation even for high-order converter topologies,
- (ii) constant switching frequency,
- (iii) no steady-state errors,
- (iv) possibility to implement a current limitation,

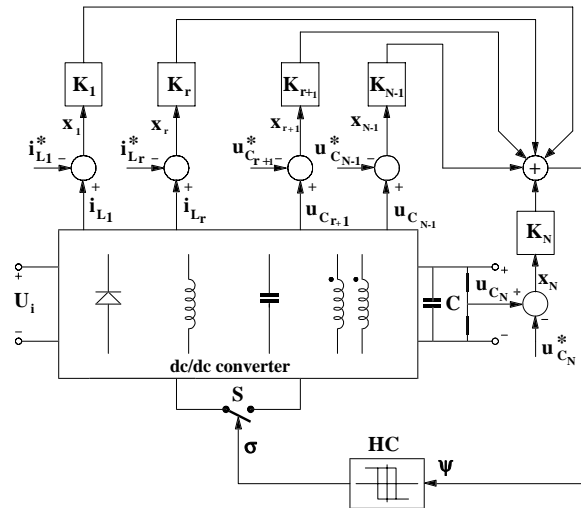


Fig. 4 - Sliding mode control of dc-dc converters: principle scheme.

which are not satisfied in the scheme of Fig.4. In fact, for fourth order converters like Cuk and Sepic, it requires sensing of four state variables, which is not acceptable from practical point of view. Moreover, the hysteretic operation causes a variable switching frequency, while steady-state errors can arise from the fact that, in a practical implementation, all state variable errors, except for the output voltage, are computed using high-pass filters,⁸ thus they have zero average value. If sliding function ψ , due to the hysteretic control, has non-zero average value, a steady-state output voltage error necessarily appears. Lastly, switch current limitation is not implemented in the scheme shown.

As far as control complexity is concerned, it has been proved that excellent performances can be obtained even with reduced-order controllers, i.e. by sensing only one inductor current (the only inductor current in 2nd-order schemes like

buck, boost and buck-boost, and the input inductor current for 4th-order schemes, like Cuk and Sepic) and the output voltage.⁸⁻¹⁰ This corresponds to choose zero value for some coefficients K_i in (4).

3.2. Practical sliding mode controller of dc-dc converters

The general-purpose sliding mode controller scheme is shown in Fig.5.⁹ As mentioned above, only one inductor current and the output voltage are sensed.

3.2.1. Elimination of steady-state errors

As already mentioned, current error x_i is computed by means of a high-pass filter, while output voltage error x_u is obtained by comparison with reference signal U_o^* . A PI action is introduced on sliding function ψ in order to eliminate its dc value, thus reducing the dc value of the output voltage error to zero. In practice, the integral action of this regulator is enabled only when the system is on the sliding surface; in this way, the system behavior during large transients, when ψ can have values far from zero, is not affected, thus maintaining the fast response of sliding mode control.

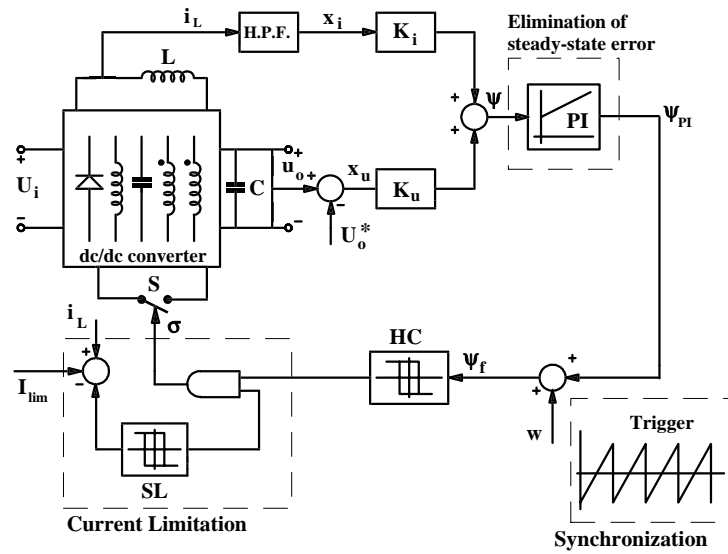


Fig.5 - General purpose sliding mode controller scheme

3.2.2. Switching frequency stabilization

In order to provide stabilization of the switching frequency, a proper ramp signal w at the desired frequency f_w is added to function ψ . If, in the steady state, the amplitude of w is predominant in ψ_f , a commutation occurs at any cycle of w , thus making the switching frequency equal to f_w . This also allows converter synchronization to an external trigger. Instead, under dynamic conditions, error terms x_i and x_u increase, w is overridden, and the system retains the excellent dynamic response of the sliding mode.

Note that an alternative way to obtain constant switching frequency is to modulate the hysteresis band amplitude, e.g. by means of a phase locked loop.

3.2.3. Current limitation

The current limiter shown in Fig.5 overrides sliding mode control when the inductor current exceeds I_{lim} . If this happens, the hysteretic control maintains this current around the value I_{lim} . This corresponds to perform a sliding mode on the inductor current only.

It is important to note that the controller design requires selection of two parameters, the time constant of the high-pass filter used to compute the inductor current error, and the ratio K_i/K_u of the sliding coefficients (their absolute value is not important as long as eq. (4) is verified). However, the general theory does not take into account the additional

state variable of the filter. In order to overcome this problem, it is possible to derive a small-signal model of the sliding mode control, which takes into account also the filter time constant, so that the design parameters can be chosen from small-signal analysis as in current-mode control.¹¹

3.3 Example of design procedure: Sepic converter

The Sepic converter topology is reported in Fig.6. It is a fourth-order system with the same voltage conversion ratio as flyback converter. Controller structure follows the general scheme of Fig.5. In order to give design criteria for selection of sliding mode controller parameters, the system must be represented in a suitable mathematical form.^{9,10} To this purpose, the converter equations related to the two subtopologies corresponding to the switch status are written as

$$\dot{\mathbf{v}} = \mathbf{A}_{\text{on}} \mathbf{v} + \mathbf{F}_{\text{on}} \quad \text{switch on,} \quad (5.a)$$

$$\dot{\mathbf{v}} = \mathbf{A}_{\text{off}} \mathbf{v} + \mathbf{F}_{\text{off}} \quad \text{switch off,} \quad (5.b)$$

where $\mathbf{v} = [i_1, i_2, u_1, u_2]^T$ is the state variable vector.

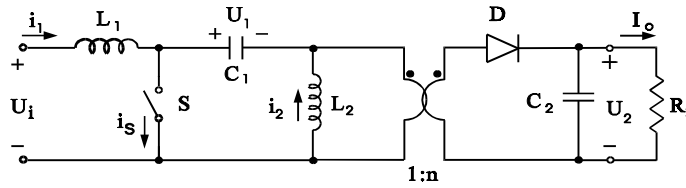


Fig. 6 - Dc/dc Sepic converter.

These equations are combined in the following form (VSS)

$$\dot{\mathbf{v}} = \mathbf{A} \mathbf{v} + \mathbf{B} \sigma + \mathbf{F}, \quad (6.a)$$

where σ is the discontinuous variable corresponding to the switch status and matrices \mathbf{A} , \mathbf{B} , \mathbf{F} are given by

$$\mathbf{A} = \mathbf{A}_{\text{off}}, \quad \mathbf{F} = \mathbf{F}_{\text{off}}, \quad (6.b)$$

$$\mathbf{B} = (\mathbf{A}_{\text{on}} - \mathbf{A}_{\text{off}}) \mathbf{v} + (\mathbf{F}_{\text{on}} - \mathbf{F}_{\text{off}}). \quad (6.c)$$

It is convenient to write the system equations (6.a) in terms of error state variables x_i

$$\mathbf{x} = \mathbf{v} - \mathbf{V}^*, \quad (7)$$

where $\mathbf{V}^* = [I_1^*, I_2^*, U_1^*, U_2^*]^T$ is the vector of state variable references. Accordingly, (6.a) becomes

$$\dot{\mathbf{x}} = \mathbf{A} \mathbf{x} + \mathbf{B} \sigma + \mathbf{G}, \quad (8)$$

where $\mathbf{G} = \mathbf{A} \mathbf{V}^* + \mathbf{F}$.

3.3.1. Existence condition

Assuming that the switch is kept on ($\sigma=1$) when ψ is negative and off ($\sigma=0$) when ψ is positive, we may express the existence condition in the form

$$\begin{aligned} \frac{\partial \psi}{\partial t} &= \mathbf{K}^T \mathbf{A} \mathbf{x} + \mathbf{K}^T \mathbf{G} < 0 & 0 < \psi < \xi \\ \frac{\partial \psi}{\partial t} &= \mathbf{K}^T \mathbf{A} \mathbf{x} + \mathbf{K}^T \mathbf{B} + \mathbf{K}^T \mathbf{G} > 0 & -\xi < \psi < 0, \end{aligned} \quad (9)$$

where ξ is an arbitrary small positive quantity.^{4,5} Note that (9) simply tell us that when the switch is *off* function ψ must decrease, while when it is *on* ψ must increase.

Inequalities (9) are useful only if state variable errors x_i are bounded, otherwise (9) must be analyzed under small-signal assumption. In this latter case, satisfying (9) means to enforce existence condition in a small volume around the operating point and this is equivalent to ensure stability condition as demonstrated in reference 11.

3.3.2. Hitting condition

If sliding mode exists, a sufficient hitting condition is

$$\mathbf{K}^T \mathbf{A}_4 \leq 0 \quad (10.a)$$

where \mathbf{A}_4 is the fourth column of matrix \mathbf{A} .^{4,5} This yields the following constrain

$$-\frac{K_i}{nL_1} - \frac{K_u}{R_L C_2} \leq 0 \quad (10.b)$$

3.3.3. Stability condition

This issue must be addressed taking into account the effect of the time constant τ_{HPF} of the high-pass filter needed to extract the inductor current error. To this purpose the small-signal analysis carried out in reference 11 allows to derive precise constraints on controller parameters for the three basic converter structures (buck, boost, buck-boost). For higher order converters, like Sepic, the approach outlined in reference 11 is valuable to observe the effect of controller parameter variations on the system dynamic. For example, Fig.7 shows the root locus of the closed-loop Sepic converter resulting from different values of the filter time constant. As we can see, too low values cause real poles P_1 and P_2 to become complex or even to cross the imaginary axis.

3.3.4. PI compensator

As already mentioned in paragraph 3.2.1, the integral action of PI regulator on variable ψ is enabled only when the system is on the sliding surface. Since its aim is to cancel the steady-state error on the output voltage, its time constant τ_{PI} can be chosen big enough not to affect the system stability. Moreover, it does not affect system dynamic during large transients.

3.3.5. Ramp signal w

As far as the ramp signal w is concerned, its amplitude is selected by taking into account the slope of function ψ_{PI} and the hysteresis band amplitude, so that function ψ_f hits the lower part of the hysteresis band at the end of the ramp, causing the commutation. The aim of this external ramp is to force a constant switching frequency in spite of duty-cycle variations (i.e. input and/or output voltage changes) and load variations.

As an example, Fig.8 shows simulated waveforms of ramp w , and ψ_{PI} , ψ_f signals. From waveform analyses, we can find that the slope S_e of the external ramp must satisfy the following inequality

$$S_e > \frac{\Delta B}{\delta T_s} - S_r, \quad (11)$$

in which ΔB represent the hysteresis amplitude and S_r is the slope of function ψ_{PI} during switch on-time. This latter, is given by

$$S_r = \left(\frac{U_i}{L_1} K_i - \frac{I_o}{C_2} K_u \right) \cdot K_p \quad (12)$$

where K_p is the gain of the proportional part of PI regulator.

It is worthy to note that, in the presence of an external ramp, signal ψ_{PI} must have a non zero average value in order to accommodate for the desired converter duty-cycle (see Fig.8).

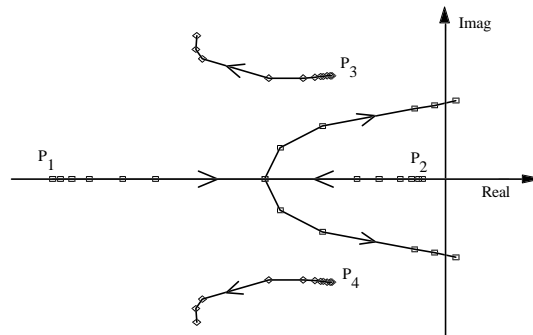


Fig. 7 - Root locus of closed-loop system for variation of low-pass filter time constant of the Sepic converter.

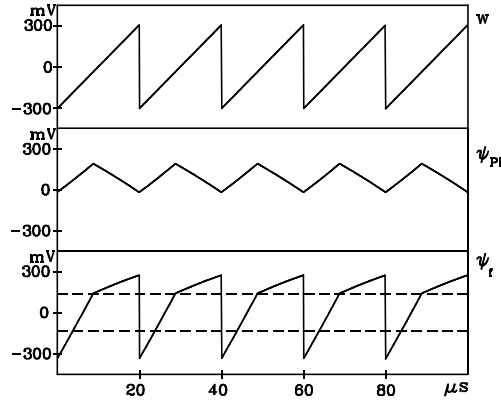


Fig. 8 - simulated waveforms of ramp w , and ψ_{PI} , ψ_f signals.

4. Application of Sliding Mode Control to AC-DC Converters

With the ever increasing demand for power quality from the utility grid, power factor correction (PFC) is becoming a basic requirement for switch-mode power supplies. High quality rectifiers are used to interface ac line and dc load and appear to the line as a resistive load, so achieving unity power factor even in presence of distorted line voltage.¹²

An inherent limitation of these converters is their slow dynamic response. In fact, sinusoidal input current means large input power fluctuations at twice the line frequency, resulting in low-frequency output voltage ripple, which cannot be corrected by the control, otherwise input current waveform is affected. Big output filter capacitors must therefore be adopted in order to limit the voltage ripple, which further reduces the bandwidth of the voltage control loop. Accordingly, these converters are normally used as pre-regulators, their output performances being below the requirements of standard power supplies.

In practice, this bandwidth limitation can be overcome if some input current distortion is accepted (within the limits posed by the standards, e.g. IEC 555-2), provided that this results in smaller power fluctuation. From this point of view, sliding mode control is very powerful, since it can provide an optimal trade-off between the needs for increasing response speed and reducing input current distortion and output voltage ripple.

A high-quality rectifier based on Cuk topology with sliding mode control is shown in Fig.9.

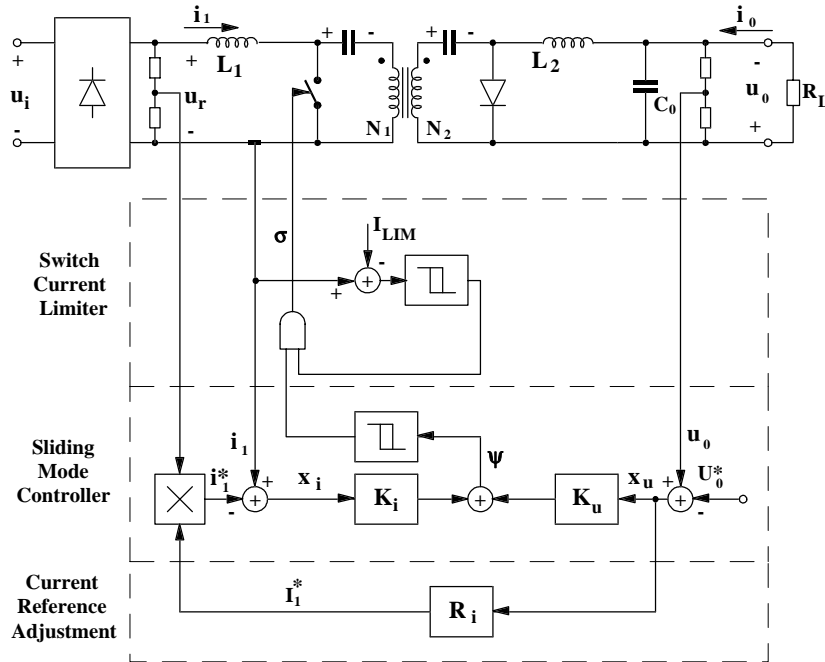


Fig. 9 - High-quality rectifier based on Cuk converter with sliding mode control.

In this scheme, including an insulation transformer, reference values (i_1^* , U_o^*) are required for both state variables. However, while U_o^* is an input variable for the control, i_1^* must be evaluated as a function of voltage error x_u . For this

purpose, similarly to the case of current-mode control, an error amplifier R_i is used to determine reference amplitude I_1^* , which is then multiplied by rectified voltage signal u_R to obtain current reference i_1^* . As usual, the band of regulator R_i is kept well below the line frequency to avoid input current distortion. However, during transient conditions, if the system remains in sliding mode, the large output voltage error forces, via block K_u , a large current error, causing a temporary loss of the sinusoidal shape of input current. In this way, a fast output voltage recovery to nominal conditions is obtained.

If coefficients K_i and K_u are chosen properly, we gain additional advantages. First, in the steady-state the controller tends to maintain the minimum errors x_i and x_u which are consistent with energy balance condition. Second, the value of the coefficients also determines the relative amplitudes of voltage and current errors. This means that varying the ratio K_i/K_u we can choose the best trade off between input current distortion and fast output voltage response.

The input current limiter is implemented in the usual manner as for dc-dc converter controllers.

5. Application of Sliding Mode Control to Single-Phase DC-AC Converters

The increasing demand in the performance of ac power supplies in terms of tight voltage regulation, low harmonic content, wide range of load variation, overload protection and fast response, requires the use of high-frequency PWM inverters. Moreover a robust control technique is needed in order to ensure proper operation even with reactive and/or non-linear loads. For these reasons sliding mode control seems suitable also for these converters.¹³⁻¹⁷

The classical single-phase ac power supply is shown in Fig.10. It includes a bridge inverter feeding the insulating transformer and a filter made up by smoothing inductor L and output capacitor C .

As it can be seen, this topology resemble that of a buck converter except for the filter input voltage, which can be two- or three-level depending on the inverter modulation strategy, and for the presence of the insulating transformer. Thus, the implementation of the sliding mode control follows the guidelines given for dc-dc converters, with the important difference represented by the sinusoidal variation of voltage and current references. This fact influences the existence conditions of the sliding mode, because the rate of change of these references must be taken into account. Moreover, in order to improve the steady-state performance, an exact current reference is needed, which can be accomplished by direct measurement of the load current¹³⁻¹⁵ or by a suitable observer.¹⁶

It is important to note that the presence of an insulating transformer modifies the system phase trajectories introducing a term linearly dependent on time, which causes their translation in the phase plane. This phenomenon does not modify appreciably the shape of the phase trajectories as long as the translation is much slower than the oscillatory term.

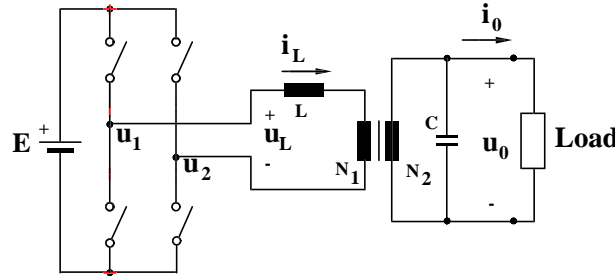


Fig. 10 - Single-phase ac power supply scheme.

6. Experimental Results

In this section we report some experimental results illustrating the performance of sliding mode control applied to switch-mode power supplies. A first application considered is a dc-dc converter based on the Sepic topology shown in Fig.6, whose parameters are listed in table 1.¹⁰ In order to compare the performance of sliding mode control with standard control technique, a current mode control was also implemented. This latter requires a R-C damping network across energy transfer capacitor C_1 in order to stabilize the current loop.

Fig.11 reports the output voltage and input current behavior in the case of load step variation from full-load to no-load and vice-versa (the damping network was maintained for both controllers). Sliding mode control shows better performances, both in terms of overshoot and response speed. Moreover, it remains stable even without the damping network and in presence of wide input voltage variations.

Table 1. Parameter values of the DC/DC Sepic converter.

$U_i = 15V$	$U_2 = 20V$	$f_s = 50kHz$	$n = 1.5$	$R_L = 20\Omega$
$L_1 = 0.7mH$	$L_2 = 0.38mH$	$C_1 = 6.8\mu F$	$C_2 = 100\mu F$	
$K_i = 1.1$	$K_u = 1$	$\tau_{HPF} = 0.5ms$	$\tau_{PI} = 0.5ms$	

As an example of what we can obtain in terms of response speed of an ac-dc converter with power factor correction, the scheme of Fig.9 was implemented and tested. Its parameters are reported in table 2.

Table 2. Parameter values of the AC/DC converter of Fig.9.

$U_i = 100V_{peak}$	$U_o = 15V$	$f_s = 60kHz$	$N_1/N_2 = 10$	$R_L = 5\Omega$
$L_1 = 4mH$	$L_2 = 40\mu H$	$C_1 = 0.22\mu F$	$C_2 = 100\mu F$	$C_o = 10000\mu F$

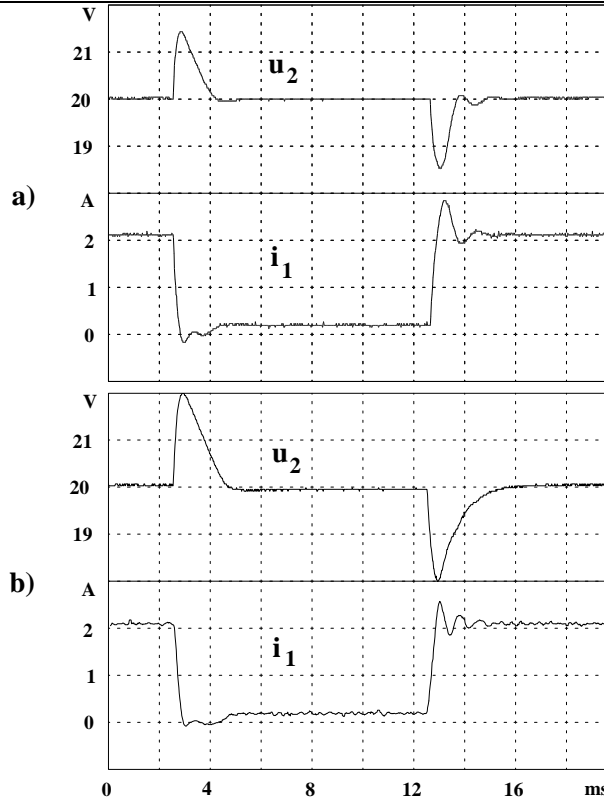


Fig. 11 - Output voltage and input current waveforms of the Sepic converter for load step variation:
a) sliding mode control, b) current mode control.

Fig.12 shows line current and output voltage at start-up: the current distortion during the transient, caused by the high error voltage x_u , is the price for the faster output dynamic. In the steady state, however, a good current waveform is achieved.

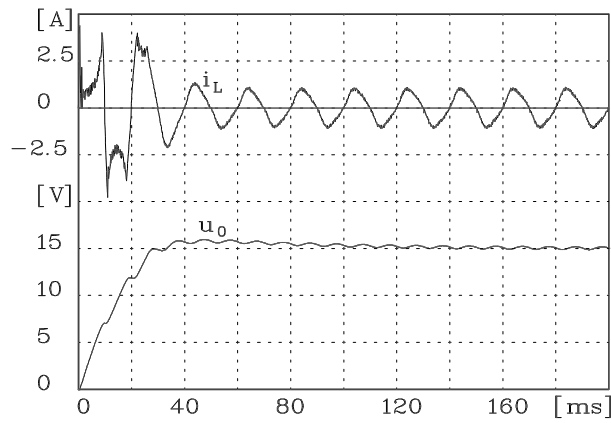


Fig. 12 - Line current and output voltage of the AC/DC converter at start-up.

Lastly, a prototype of the ac power supply shown in Fig.10 was built and experimentally tested with the parameters listed in table 3. In Fig.13 the start-up process is illustrated. A step voltage reference is applied, and the transient is very fast and smooth, with no need of soft-start provisions. Fig.14 reports the converter behavior in overload condition, for an inductive load. The current limit acts at every half period, but the recovery transient remains well limited, thus allowing quasi-regular operation.

Table 3. Parameter values of the DC/AC converter of Fig.10.

$E = 500V$	$U_o = 290V_{rms}$	$f_o = 50Hz$	$N_1/N_2 = 1/1.2$	$P_L = 5kVA$
$L = 3.2mH$	$L_\mu = 1.17H$	$C_o = 48.6\mu F$	$f_s = 20kHz$	

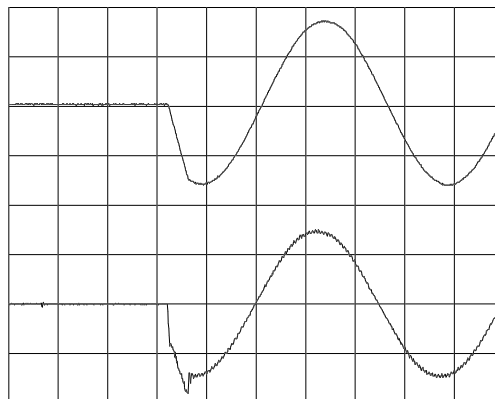


Fig. 13 - Top to bottom: output voltage (250V/div., 4ms/div.), inverter current (25A/div.).

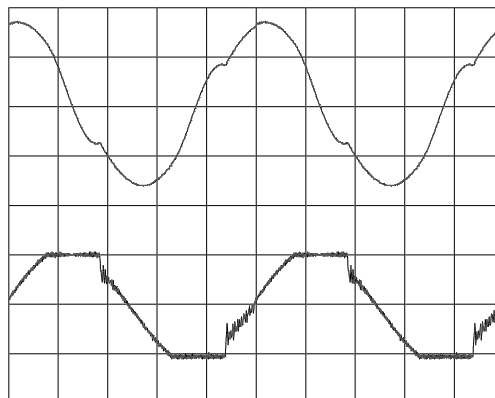


Fig. 14 - Top to bottom: output voltage (250V/div., 4ms/div.), inverter current (25A/div.).

7. Conclusions

Control techniques of variable structure systems find a natural application to switch-mode power supplies. In particular, the sliding mode control represents a powerful tool to enhance performances of power converters.

Sliding mode control is able to ensure system stability even for large supply and load variations, good dynamic response and simple implementation, even for high-order converters. These features make this control technique a valid alternative to standard control approaches like current-mode control.

Application to dc-dc converters, as well as rectifiers and inverters, is analyzed and provisions to overcome the inherent drawbacks of sliding mode control, i.e. variable switching frequency and possible steady-state errors, are described.

Experimental results are also reported, showing the effectiveness of this control technique.

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