Comparative study of two sliding surfaces to control a double boost converter

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Abstract

In this paper, a double cascade boost converter is studied and controlled. The system dynamics have been described by the averaged 4th order nonlinear state space model (1). Based on such a model and taking account the non-minimum phase nature of the converter, two robust non-linear controllers are synthesized using the sliding mode approach, where each controller use a different surface and a different structure. The first controller is elaborate with two nested loops, the inner currant loop is designed with the sliding mode approach and the outer voltage loop is designed with the famous PI controller, the second controller is also designed with the sliding mode approach, but this time with different sliding surface, this surface allows to use a single control loop to regulate the currant and the voltage in the same time. The simulation results have been performed through Matlab/SimPowerSystems environment and show that the designed controllers meet their objectives.

Keywords
Sliding mode approach, DC-DC converter, double cascade boost, non-minimum phase.

1.Introduction

To keep up with the progress of the technology in invention's world and electric installation, there is a vital part which is needed to be developed and improved, this one is the conversion of energy or more precisely the power electronics which steel pose many problems in term of stability, response time and circuit's harmonics etc. To overcome those problems, the vital aim is to focus on the automatic control of the electronic systems with chosen the most appropriate methods and parameters. About the methods there is plenty, as backstepping, passivity, flatness and sliding mode approach, whose give good and satisfying results.

This work focus on the DC-DC converters, those converters have many features as boosting or decreasing the output voltage, when this paper is interested in improving and increasing the voltage gain, to get this purpose several strategies are proposed, and between the more interesting strategies there is the cascading of two or more elementary converters. Some previous works have presented a study of cascaded converters, such as (Rensburg et al., 2008) who presented a new boost topology which ensures a significant improvement on boost factor, and(Zhao and Lee, 2003) who presents a family of high-efficiency, high step-up clamp-mode converters without extreme duty ratios. In this article, a study of two cascaded boost converters will be presented.

In this paper, a double cascade boost converter is studied and controlled, this system is just a cascading of two identical elementary simple boosts which allows to improve the voltage gain without using systems with high frequency. The system dynamics have been described by the averaged 4th order nonlinear state s pace model (1). Based on such a model and taking account the non-minimum phase nature of the converter, two robust non-linear controllers are synthesized using the sliding mode approach; this one is suitable to our system thanks to its appropriateness for the nonlinearity aspect of power system converter and also because of its robustness and capability of system order reduction (Slotine et al., 1991), (FLOQUET, 2000) ; where each controller use a different sliding surface and a different structure, and as the converter comprise two stages, each stage will be controlled separately. The first controller is elaborate with the traditional surface proposed by (Utkin, 2013), moreover because of the non-minimum phase of the double boost converter the regulator must comprise two nested loops, where the inner currant loop is designed with the sliding mode approach and the famous PI linear controller is used to design the outer voltage loop, on the other hand a second controller is also designed with the sliding mode approach, but this time with different sliding surface, this surface which is proposed by(Hijazi et al., 2009) allows to use a single control loop to regulate the currant and the voltage in the same time. The simulation results have been performed through Matlab/SimPowerSystems environment and show that the designed controllers meet their objectives.

The paper is organized as follows: in Section 2, the Double Cascade Boost converter is described and modeled; when the section 3 is devoted to designing the Sliding mode controller with two methods, and whose performances are illustrated in section 4; in the last the section 5 gives a conclusion and a reference list end the paper.

2. Double Boost Converter Presentation and Modeling

A double boost systemis a non-isolated DC-DC converter with high voltage ratio, it comprises two identical elementary boosts placed in cascades. The figure below illustrates the topology of a double boost converter, wherein each stage is controlled by a binary signal μ .

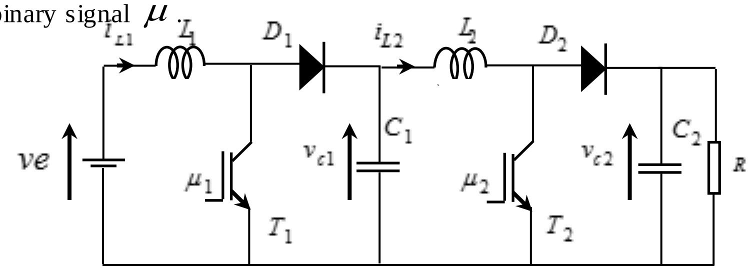


Figure 1. Double Boost converter

Where:

- i_{L1} and i_{L2} are respectively the currents in inductors L_1 and L_2 .
- v_{c1} denotes the voltage in capacitor C_1 and v_{c2} is the output voltage.
- μ_1 and μ_2 are denoted the duties ratio functions:

$$\mu_j(j=1,2)=\begin{cases}1 \quad if \quad T_j \quad is \quad ON \quad and \quad D_j \quad is \quad OFF \\ 0 \quad if \quad T_j \quad is \quad OFF \quad and \quad D_j \quad is \quad ON\end{cases}$$

2.1 Switched model

By applying standard Kirchhoff Laws to the circuit, it can be represented by the following differential equations:

$$L_1\frac{di_{L1}}{dt}=ve-(1-\mu_1)v_{c1} \qquad (1.a)$$

$$C_1\frac{dv_{c1}}{dt}=(1-\mu_1)i_{L1}-i_{L2} \qquad (1.b)$$

$$L_2\frac{di_{L2}}{dt}=v_{c1}-(1-\mu_2)v_{c2} \qquad (1.c)$$

$$C_2\frac{dv_{c2}}{dt}=(1-\mu_2)i_{L2}-\frac{v_{c2}}{R} \qquad (1.d)$$

2.2 Averaged model

The above equations represent the instantaneous model of a double boost converter; although the control signal u (the switch position) is a binary variable changing at a high frequency and some variables of the boost converter have discontinuous derivative. Therefore, it is convenient to establish the control objective and some other relationships in terms of average variables.

The averaged model of such a converter is shown to be the following:

$$\dot{x}_1 = \frac{ve}{L_1} - \frac{(1-u_1)}{L_1} x_2 \quad (2.a)$$

$$\dot{x}_2 = \frac{(1-u_1)}{C_1} x_1 - \frac{x_3}{C_1} \quad (2.b)$$

$$\dot{x}_3 = \frac{x_2}{L_2} - \frac{(1-u_2)}{L_2} x_4 \quad (2.c)$$

$$\dot{x}_4 = \frac{(1-u_2)}{C_2} x_3 - \frac{x_4}{RC_2} \quad (2.d)$$

where x_1, x_2, x_3 and x_4 denote the average values over cutting periods, of the signals i_{l1}, v_{c1}, i_{l2} and v_{c2}. The controls inputs for the above model is the functions u_1, u_2 called duty ratio function, which is the average values over cutting periods of signals μ_1, μ_2.

2.3 Equilibrium points

The equilibrium point can be obtained by forcing the time derivative of the average model (2) to be null $\dot{x} = 0$.

$$x_1^e = \frac{ve}{R(1-u_{10})(1-u_{20})} \quad (3.a)$$

$$x_2^e = \frac{ve}{(1-u_{10})} \quad (3.b)$$

$$x_3^e = \frac{ve}{R(1-u_{10})(1-u_{20})^2} \quad (3.c)$$

$$x_4^e = \frac{ve}{R(1-u_{10})^2(1-u_{20})^2} \quad (3.d)$$

3. Controller design

Double Boost converter being a variable structure system, where, switching over changes the structure of the system. Sliding mode control is the most suitable control strategy for such systems, because of its low sensitivity to the system parameters variations and disturbances which makes the control robust.

3.1 Control objectives

- Controlling the voltage x_2 to make it track a given reference signal x_2^*.
- Regulating the output voltage x_4 to pursuit a reference value x_4^*.

x_2^* and x_4^* denote the corresponding voltage's references signals.

3.2 Controller design

The double cascade boost converter has a non-minimum phase nature (Byrnes and Isidori, 1988), this nature causes a stability and control problems because when we regulate the output voltage directly the current phase grow unstably which causes a damage of the converter, and as it shown in (Sira-Ramirez, 1987), the currant must increase continuously so that the surface $v - v^*$ tend to zero, this will include other objectives which are:

- the input current x_1 must track, as closely as possible, a given reference signal x_1^*.
- And the current x_3 in inductor L_2 also must track, as closely as possible, a given reference signal x_3^*.

x_1^* and x_3^* denote the corresponding currant's references signals.

To solve the problem causes by the non-minimum phase nature, two nested controllers must be used for each stage of the convertor: an inner current control loop and an outer voltage control loop, but also a single control loop can be used thanks to the sliding surface proposed in(Hijazi et al., 2009) this surface allow to regulate the current and the voltage in the same time. In this section, the both solutions will be introduced.

3.2.a Controller design with two loops

as mentioned before; two nested loops will be used, the first loop which is the voltage controller is usually realized with standard linear control techniques, whereas the current control is implemented using the sliding mode approach for the control of inductors current. The following figure show the system structure which comprise four blocs:

- the **PI** controller which is a linear regulator used to generate the current references x_1^* and x_3^* this bloc is controller's outer loop.
- the second bloc which is **SMC with S1** is the inner controller synthetized with the sliding mode approach using the first proposed sliding surface (4).
- the third bloc the **PWM** is the famous pulse-with modulation generator.
- the last bloc is the **DBC** which is the double boost converter illustrate on figure (1). Where the currant's error z_1, $z_1 = \begin{pmatrix} z_{11} \\ z_{12} \end{pmatrix}$, and voltage's error z_2, $z_2 = \begin{pmatrix} z_{21} \\ z_{22} \end{pmatrix}$.

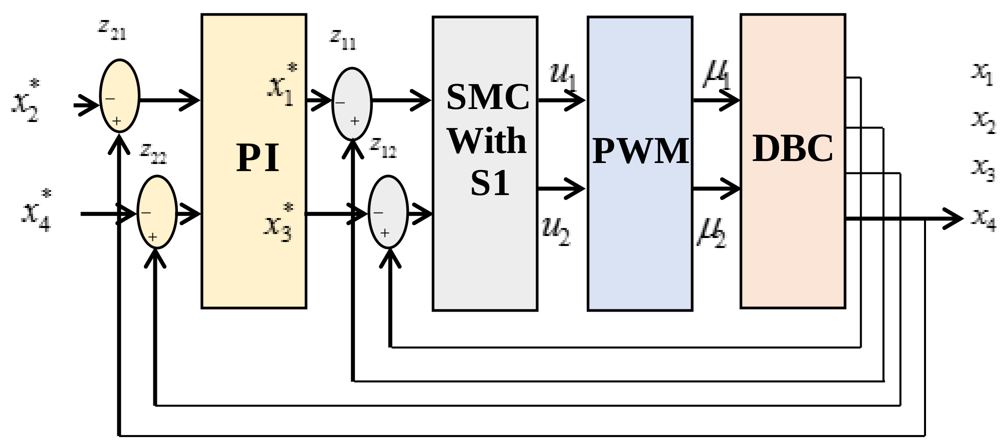


Figure 2. structure of the system with a regulator designed with two loops

a.1 The choice of the sliding surface

To start an appropriate sliding function s must be chosen to guarantee the system stability in the sliding modes $s = 0$ (Slotine et al., 1991):

$$s(x) = \left(\frac{\partial}{\partial t} + k_x \right)^{n-1} z(x) \qquad (4)$$

$z(x)$ the deviation of the controlled variable, k_x a positive constant and n the system order.

The Double Boost Converter comprise two stages where each stage can be considered as a subsystem, each subsystem is a second order system $n = 2$ then we get two sliding surfaces:

$$s = \begin{pmatrix} s_1 \\ s_2 \end{pmatrix} = \begin{pmatrix} \dot{z}_{11} + k_1 z_{11} \\ \dot{z}_{12} + k_2 z_{12} \end{pmatrix} \qquad (5)$$

Where the currant's error z_1 :

$$z_1 = \begin{pmatrix} z_{11} \\ z_{12} \end{pmatrix} = \begin{pmatrix} L_1\left(x_1 - x_1^*\right) \\ L_2\left(x_3 - x_3^*\right) \end{pmatrix} \qquad (6)$$

a.2 Existence condition

To ensure the existence of the sliding mode; a dynamic of defined Lyapunov function must be negative $\dot{V} \prec 0$. The Lyapunov function is defined as:

$$V = \frac{1}{2}s^2 \qquad (7)$$

its dynamic is:

$$\dot{V} = s\dot{s} \qquad (8)$$

And the dynamic of the sliding surface S is:

$$\dot{s} = \begin{pmatrix} \dot{s}_1 \\ \dot{s}_2 \end{pmatrix} = \begin{pmatrix} \dot{u}_1 x_2 - (1-u_1)\dot{x}_2 - L_1\ddot{x}_1^* + k_1\, ve - k_1(1-u_1)x_2 - k_1 L_1 \dot{x}_1^* \\ \dot{x}_2 + \dot{u}_2 x_4 - (1-u_2)\dot{x}_4 - L_2\ddot{x}_3^* + k_2\ x_2 - k_2(1-u_2)x_4 - k_2 L_2 \dot{x}_3^* \end{pmatrix} \qquad (9)$$

In sliding mod control, we usually determine u as following (Hu, 2016):

$$u = \begin{cases} 1 & if \quad s \prec 0 \\ 0 & if \quad s \succ 0 \end{cases} \qquad (10)$$

Using (10) to ensure the existent condition, $\dot{s}$ must be defined as flowing:

$$\begin{cases} \dot{s} \succ 0 & if \quad s \prec 0 \\ \dot{s} \prec 0 & if \quad s \succ 0 \end{cases} \qquad (11)$$

Thanks to (11) there is no need to calculate the product $S\dot{S}$ and the sliding region can be easily deduced, this region is limited by the following inequalities:

$$\dot{s} = \begin{cases} \dot{s}_1 = \begin{cases} -L_1\ddot{x}_1^* + k_1\, ve - k_1 L_1 \dot{x}_1^* \succ 0 \\ -\dot{x}_2 - L_1\ddot{x}_1^* + k_1\, ve - k_1 x_2 - k_1 L_1 \dot{x}_1^* \prec 0 \end{cases} \\ \dot{s}_2 = \begin{cases} \dot{x}_2 - L_2\ddot{x}_3^* + k_2\ x_2 - k_2 L_2 \dot{x}_3^* \succ 0 \\ \dot{x}_2 - \dot{x}_4 - L_2\ddot{x}_3^* + k_2\ x_2 - k_2 x_4 - k_2 L_2 \dot{x}_3^* \prec 0 \end{cases} \end{cases} \qquad (12)$$

At last the condition will be satisfied if:

$$k_1 \succ \frac{\dot{x}_2 + L_1\ddot{x}_1^*}{ve - L_1\dot{x}_1^* - x_2} \qquad (13.a)$$

$$k_2 \succ \frac{\dot{x}_4 - \dot{x}_2 + L_2\ddot{x}_3^*}{x_2 - \dot{x}_3^* - x_4} \qquad (13.b)$$

a.3 Current loop design

where $u_{eq}(t)$ and $u_n(t)$ are the equivalent control and the switching control, respectively. The equivalent control, $u_{eq}(t)$, proposed by (Utkin, 2013) is based on the nominal (estimated) plant parameters and provides the main control action, while the switching control, $u_n(t)$, ensures the discontinuity of the control law across sliding surface. If the initial error is not on the sliding surface $s(t)$ or there is a deviation of the representative point from $s(t)$ due to parameter variations and disturbances, the controller must be designed such that it can drive the error to the sliding surface. The error under the condition that will move toward and reach the sliding surface is said to be on the reaching phase. The control's law defined by:

$$u(t) = u_{eq}(t) + u_n(t) \tag{14}$$

- the equivalent control

the equivalent control can be calculated from $\dot{s} = 0$, and the obtained control's low is:

$$\dot{u}_{eq1} = \frac{(1-u_{eq1})(\dot{x}_2 + k_1 x_2) + L_1(\ddot{x}_1^* + k_1 \dot{x}_1^*) - k_1 ve}{x_2} \tag{15.a}$$

$$\dot{u}_{eq2} = \frac{(1-u_{eq2})(\dot{x}_4 + k_2 x_4) + L_2(\ddot{x}_3^* + k_2 \dot{x}_3^*) - k_2 x_2 - \dot{x}_2}{x_4} \tag{15.b}$$

- the switching control

The switching control law, is introduced as (Slotine et al., 1991):

$$u_n(t) = \begin{cases} u_{n1} = -\alpha_1 sign(s_1) \\ u_{n2} = -\alpha_2 sign(s_2) \end{cases} \tag{16}$$

Where α_1 and α_2 are positive constants and sign function is defined as:

$$sign(s) = \begin{cases} 1 \rightarrow s \succ 0 \\ 0 \rightarrow s = 0 \\ -1 \rightarrow s \prec 0 \end{cases}$$

Then the find controller law is:

$$u = \begin{pmatrix} u_{eq1} + u_{n1} \\ u_{eq2} + u_{n2} \end{pmatrix} = \begin{pmatrix} u_{eq1} - \alpha_1 sign(s_1) \\ u_{eq2} - \alpha_2 sign(s_2) \end{pmatrix} \tag{17}$$

a.4 Voltage loop design

The aim now is to design a tuning law for x_1^*, x_3^*. And usually, the voltage control is achieved with standard linear control techniques, that's why the PI controller is used to generate the currant references. A low pass filter is adopted, then the PI controller structure is de fined by:

$$x_1^* = \left(k_{p1} z_{21} + k_{i1} \int_0^t z_{21}\right) \frac{a}{s+a} \tag{18.a}$$

$$x_2^* = \left(k_{p2} z_{22} + k_{i2} \int_0^t z_{22}\right) \frac{b}{s+b} \tag{18.b}$$

Where (k_{p1}, k_{p2}) the proportional gains, (k_{i1}, k_{i2}) integral gains, and (a, b) positive constants.

With $z_2 = \begin{pmatrix} z_{21} \\ z_{22} \end{pmatrix} = \begin{pmatrix} y_1^* - y_1 \\ y_2^* - y_2 \end{pmatrix}$, $y_1 = (x_2)^2$, $y_2 = (x_4)^2$, $y_1^* = (x_2^*)^2$ and $y_2^* = (x_4^*)^2$.

3.2.b Controller design with single loop

In this part, another sliding surface will be used for the controller design, this surface proposed by (Hijazi et al., 2009) allows to ensure the regulation of the currant and the voltage in the same time, which make the controller design less complicated and more performer.

The figure bellow illustrates the new general system structure using the second surface, which comprise this time three blocs:

- the first bloc is **SMC with S2** which is the inner controller synthetized with the sliding mode approach using the sliding surface (4) proposed by (Hijazi et al., 2009).
- **PWM**: pulse-with modulation generator.

- the last bloc is the **DBC** which is the double boost converter illustrate on figure (1). Where the voltage's error z_{11} and z_{12}.

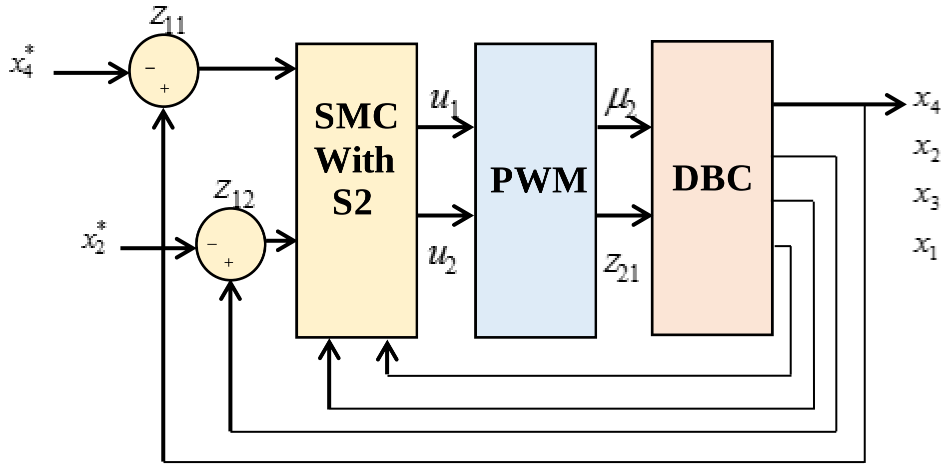


Figure 3. structure of the system with a regulator designed with single loop

b.1 The choice of the sliding surface

The surface used in this controller is proposed by (Hijazi et al.+, 2009), and can be introduced like:

$$s = \begin{cases} s_1 = k_3 z'_{21} + k_4 z_{11} \\ s_2 = k_5 z'_{22} + k_6 z_{12} \end{cases} \tag{19}$$

Where k_3, k_4, k_5 and $k_6 \in R^+$, $z'_2 = \begin{pmatrix} z'_{21} \\ z'_{22} \end{pmatrix} = \begin{pmatrix} C_1(x_2 - x_2^*) \\ C_2(x_4 - x_4^*) \end{pmatrix}$ and z_1 correspond to (6)

As reference current depends on the operating point, will be extracted from (3):

$$x_1^* = \frac{(x_2^*)^2}{Rve} \tag{20.a}$$

$$x_3^* = \frac{(x_4^*)^2}{Rx_2^*} \tag{20.b}$$

Sliding surface coefficients k_3, k_4, k_5 and k_6 must be chosen to ensure that the sliding mode exists at least around the desired equilibrium point, and the dynamics of the system will reach the surface and lead toward the equilibrium point.

b.2 Existence condition

As mentioned before, to ensure the existence of the sliding mode; a dynamic of Lyapunov function must be negative $\dot{V} \prec 0$. the same Lyapunov function defined before $V = \frac{1}{2}s^2$ and its dynamic $\dot{V} = s\dot{s}$ will be used.

the dynamic of the sliding surface s (20) is:

$$\dot{s} = \begin{cases} \dot{s}_1 = k_3 ve - k_3(1-u_1)x_2 - k_3 L_1 \dot{x}_1^* + k_4(1-u_1)x_1 - k_4 x_3 \\ \dot{s}_2 = k_5 x_2 - k_5(1-u_2)x_4 - k_5 L_2 \dot{x}_3^* + k_6(1-u_2)x_3 - k_6 \frac{x_4}{R} \end{cases} \tag{21}$$

Using (11) and (20), the sliding region which is limited by the following inequalities can easily deduced

$$\dot{s} = \begin{cases} \dot{s}_1 \begin{cases} k_3 ve - k_3 z'_{21} + k_3 x_2^* - k_3 L_1 \dot{x}_1^* + k_4 z_{11} - k_4 x_1^* - k_4 z_{12} + k_4 x_3^* \prec 0 \\ k_3 ve - k_3 L_1 \dot{x}_1^* - k_4 z_{12} + k_4 x_3^* \succ 0 \end{cases} \\ \dot{s}_2 \begin{cases} k_5 z'_{21} - k_5 x_2^* - k_5 z'_{22} + k_5 x_4^* - k_5 L_2 \dot{x}_3^* + k_6 z_{12} - k_6 x_3^* - k_6 \dfrac{z'_{22}}{R} + k_6 \dfrac{x_4^*}{R} \prec 0 \\ k_5 z'_{21} - k_5 x_2^* - k_5 L_2 \dot{x}_3^* - k_6 \dfrac{z'_{22}}{R} + k_6 \dfrac{x_4^*}{R} \succ 0 \end{cases} \end{cases} \quad (22)$$

From those inequalities and in order to ensure that the sliding mode exists at least around the equilibrium point (3) with z_1, z'_2 tend to zero at (3), the following condition must be satisfied:

$$\frac{k_3}{k_4} \prec \frac{x_1^* - x_3^*}{ve - \dot{x}_1^* + x_2^*} \quad (23.a)$$

$$\frac{k_5}{k_6} \prec \frac{x_3^* - \dfrac{x_4^*}{R}}{-x_2^* + x_4^* - L_2 \dot{x}_3^*} \quad (23.b)$$

b.3 Controller design

the same control law (14) defined before will be used:

$$u = \begin{pmatrix} u_{eq1} + u_{n1} \\ u_{eq2} + u_{n2} \end{pmatrix}$$

- the equivalent control

the equivalent control can be calculated from $\dot{s} = 0$, and the obtained control's low is:

$$u_{eq1} = 1 - \frac{k_4 x_3 + k_3 L_1 \dot{x}_1^* - k_3 ve}{k_4 x_1 - k_3 x_2} \quad (24.a)$$

$$u_{eq2} = 1 - \frac{k_6 \dfrac{x_4}{R} + k_5 L_2 \dot{x}_3^* - k_5 x_2}{k_6 x_3 - k_5 x_4} \quad (24.b)$$

- the switched control

The switching control low is defined to ensure the negativity of $\dot{V}$.

Using (8), (21) and (24):

$$\dot{V}_1 = u_{1n} \left[k_3 x_2 - k_4 x_1 \right] . s_1 \quad (25.a)$$

$$\dot{V}_2 = u_{2n} \left[k_4 x_4 - k_6 x_3 \right] . s_2 \quad (25.b)$$

We look for u_{1n} and u_{2n} as:

$$u_{1n} = \frac{1}{(k_3 x_2 - k_4 x_1)} \tilde{u}_{1n} \quad (26.a)$$

$$u_{2n} = \frac{1}{(k_4 x_4 - k_6 x_3)} \tilde{u}_{2n} \quad (26.b)$$

We have

$$\dot{V}_1 = s_1 \tilde{u}_{1n} \quad (28.a)$$

$$\dot{V}_2 = s_2 \tilde{u}_{2n} \quad (28.b)$$

The suitable choice of $\tilde{u}_{1n}$ and $\tilde{u}_{2n}$

$$\tilde{u}_{1n} = -\alpha_1 \, sin\, g(s_1) \quad (29.a)$$

$$\tilde{u}_{2n} = -\alpha_2 \, sin\, g(s_2) \quad (29.b)$$

To ensure the stability's condition $\dot{V} \prec 0$:

$$\dot{V}_1 = -\alpha_1 |s_1| \quad (30.a)$$

$$\dot{V}_2 = -\alpha_2 |s_2| \quad (30.b)$$

Then the final control's law:

$$u_1 = 1 - \frac{k_4 x_3 + k_3 L_1 \dot{x}_1^* - k_3 ve - \alpha_1 sgn(s_1)}{k_4 x_1 - k_3 x_2} \quad (31.a)$$

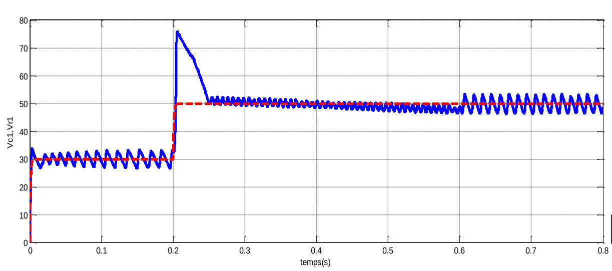
$$u_2 = 1 - \frac{k_6 \frac{x_4}{R} + k_5 L_2 \dot{x}_3^* - k_5 x_2 - \alpha_2 sgn(s_2)}{k_6 x_3 - k_5 x_4} \quad (31.b)$$

4. Simulation results

In this section, the simulation's results of the cascade boost converter with both controllers designed previously is illustrated by using SIMPOWER of MatlabSimulink, in order to verify the theoretical results using the parameters value shown in tables 1.

Table 1. systemparameters

Double boost parameters	Value	Controller 1 parameters	Value	Controller 2 parameters	Value
Inductors	$L_1 = 35\ \mu H$, $L_2 = 100 \mu H$	k_1, k_2	200000	k_3, k_5	$0,5 - 0,15$
Capacitors	$C_1 = 6\ mF$, $C_2 = 600\ \mu F$	α_1, α_2	0.03	k_4, k_6	$0,2 - 0,15$
Frequency	$10\ kHz$	k_{i1}, k_{i2}	10	α_3, α_4	0,5
Input voltage	12.8	k_{p1}, k_{p2}	40-2	x_2^*	$30 - 60$
Load resistors	$10\ \Omega$	x_2^*	$30 - 50$	x_4^*	$50 - 100$
Switchers	Ideal	x_4^*	$50 - 70$		



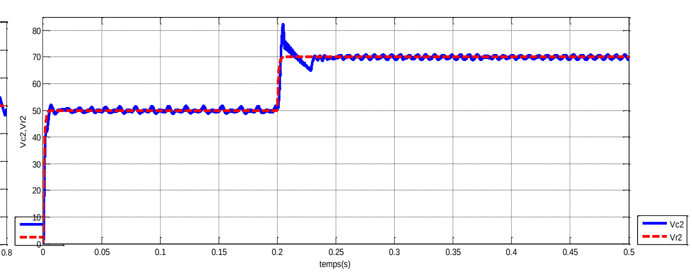


Figure 4. The intermediate voltage V_{c1} in C_1 Figure 5. The intermediate voltage V_{c2} in C_2

Figure 4 and 5 shown respectively the intermediate and the output regulated voltages (V_{c1}, V_{c2}) presented with the blue color, those signals are obtained thanks to the first controller which is synthetized by two nested loops, the regulated voltages pursuit the given references (the doted red signals) but the system is clearly instable specially the intermediate voltage (V_{c1}), then the simulation results show the weakness of the first regulator and how it need to be improved.

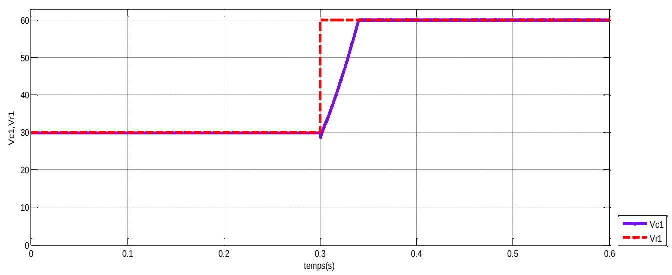


Figure 6. the intermediate voltage V_{c1} in C_1

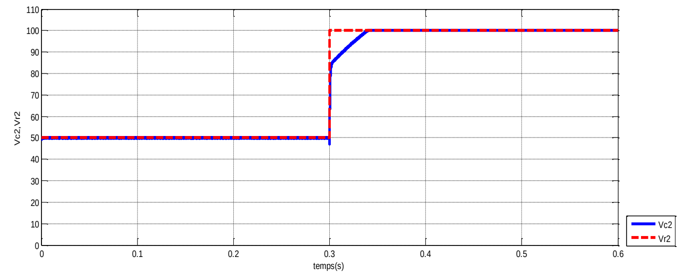


Figure 7. the intermediate voltage V_{c2} in C_2

Figure 6 and 7 shown respectively the intermediate and the output regulated voltages (V_{c1} , V_{c2}) obtained thanks to the second controller which is synthetized with a single loop using the (Hijazi et al., 2009) sliding surface, each figure illustrate two signals, the first one which is the doted red is the reference signal and the second signal which is bleu represent the voltage signal. After 0.3 s the reference signal changed and the voltage signal pursuit that change with a transition period that lasts 18 ms. Also, the two figures shown that the regulated voltages pursuit the given references as good as expected, the system is stable, then the simulation results show the efficiency of the second controller.

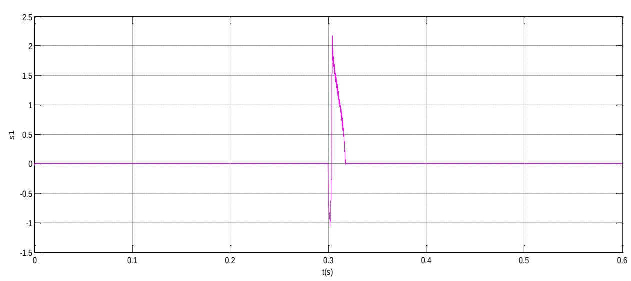


Figure 8. The sliding surface s1 (21.1)

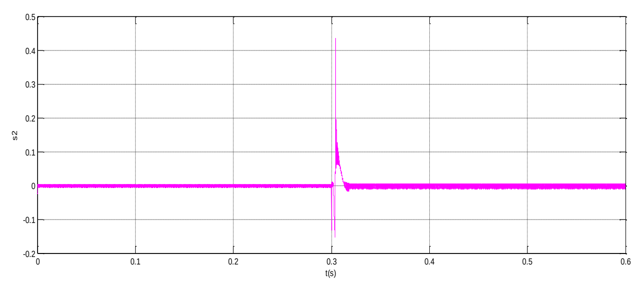


Figure 9. The sliding surface s2 (21.2)

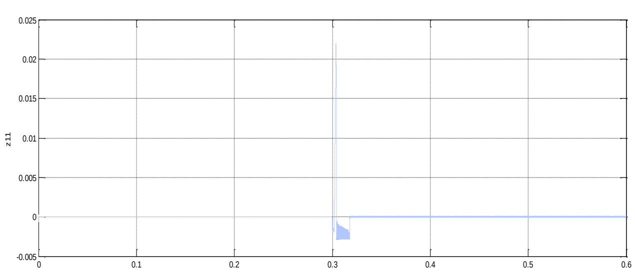


Figure 10. The currant's error z_{11} (6)

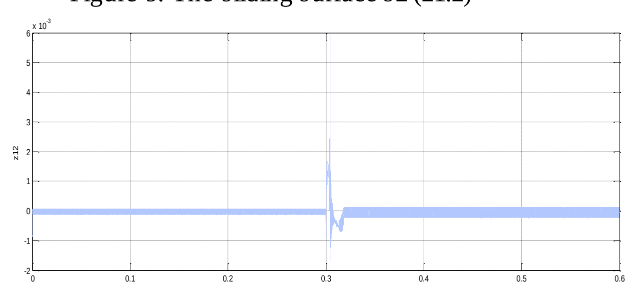


Figure 11. The currant's error z_{12} (6)

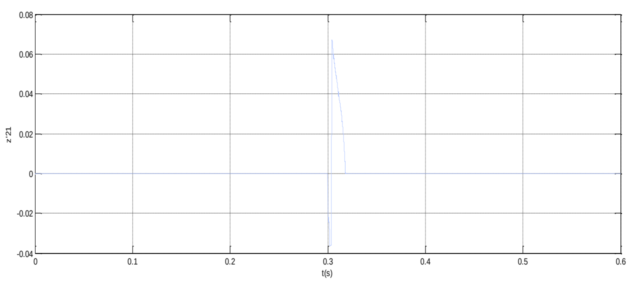


Figure 12. The voltage's error z'_{21} (19)

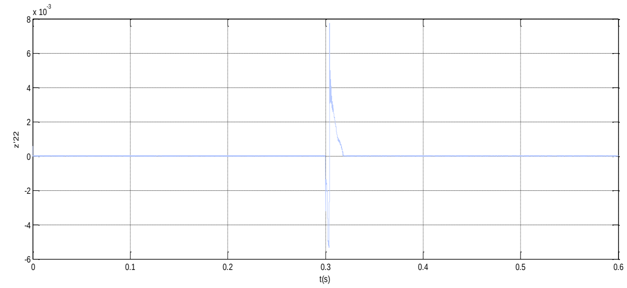


Figure 13. The voltage's error z'_{22} (19)

Figure 8 and 9 shown the sliding surfaces where $s_1 = 0$ and $s_2 = 0$, and show that the sliding mode is obtained, where the figure 10, 11, 12 and 13 shown respectively the currant's errors z_{11} and z_{12} , and the voltage's errors z'_{21} and z'_{22} , all the errors of the system vanishes (errors=0) which ensures the pursuit of the references and shows the reliability of the synthesized regulator.

5. Conclusion

To get high conversion ratio which is needed in plenty of applications, there is different topologies to obtain that aim. In this paper, we studied and controlled a double cascade boost converter, this converter has a satisfying high conversion ratio. To control this circuit, we used a sliding mode approach. This controller is designed with two different ways: firstly, the controller is designed using two nested loops utilizing the sliding mode approach for the inner currant loop and PI controller for the outer voltage loop, secondly we designed a controller using the sliding mode approach but with another sliding surface which allows us to regulate the currant and the voltage in the same time, the simulation results show that the controllers meet their objectives besides the second controller has a better results in the term of stability and tracking and a good minimization of regulator's blocs.

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Biography

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