HYBRID CONTROLLERS BASED ON SLIDING MODE CONTROL FOR AC-DC CONVERTER WITH POWER FACTOR CORRECTION

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ABSTRACT

This paper deals with hybrid controllers based on sliding mode control, a polynomial RST and a standard PI controllers. The hybridization consists to start of the SMC topology and using the RST and PI controllers to build switching signal term. The total harmonic distortion (THD) is taken into account as one criterion to evaluate the performances. Simulation results and their comparison between different schemes based on SMC show that these new controllers are realizable and lead to good performances as test tracking, disturbance rejection and robustness.

Keyword: - *AC-DC Converter, boost PFC, SMC, RST, PI controller, hybridization, THD.*

1. INTRODUCTION

The considerable evolution of electronics involved a proliferation of the apparatuses in informatics and domestic electricity devices. It should be said however that these apparatuses contribute to the deterioration of the grid quality because of the harmonic pollution which leads, for example, to a poor efficiency. Although of lower coast, classic supplies provided with AC-DC converter using capacitor filters are not any more recommended. Effectively, the rules become increasingly severe. Several solutions are already proposed whose principal goals can be summarized as follows:

- Obtaining a sinusoidal current network and in phase with the voltage
- Ensuring the smallest THD as possible in order to respect the standard norm (for example, IEC 61000-3-2 for systems of class D)
- Ensuring voltage output constant

Several topologies are presented and there are those which are rather expensive but with increased performances [1-7]. The rated of re-injection of the currents harmonics can be quantified by this total harmonics distortion (THD). The power factor (FP) is defined as:

$$F_p = \frac{P}{S} = \frac{V.I_1.\cos\varphi_1}{V.I} \qquad (1)$$

With S, P indicate respectively the apparent and active powers; I, I_1, φ_1, the effective values of AC currents and the current fundamental and the dephasing between current and tension. The effective value of current is,

$$I = \sqrt{\left(\sum_{k=1} I_k^2\right)} = \sqrt{I_1^2 + \sum_{k=2} I_k^2} \qquad (2)$$

I_k is the current of rank k.

The THD is usually defined as follows:

$$THD = \sqrt{\frac{\sum_{k=2} I_k^2}{I_1^2}} = \frac{1}{I_1} \cdot \sqrt{\sum_{k=2} I_k^2} \qquad (3)$$

According relations (1), (2) and (3),

$$F_p = \frac{\cos\varphi_1}{\sqrt{1+THD^2}} \qquad (4)$$

The power factor F_p is thus related to the THD. With a current purely sinusoidal and in phase with the voltage, the power factor approaches the unit value ($F_p \approx 1$).

In this paper, a classic scheme of a converter AC-DC without filter side network is chosen because it presents severe conditions about the THD even it has low coast. The basic scheme taken into account in this paper is showed by Fig. 1.

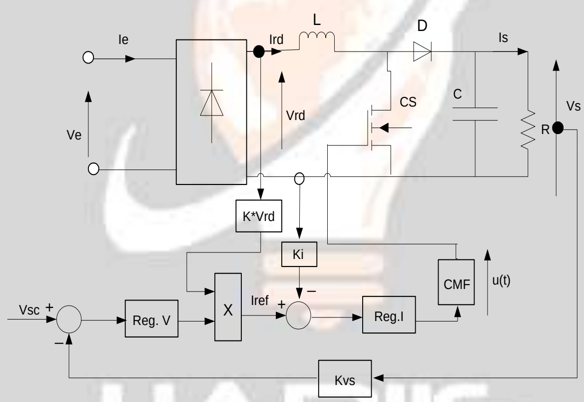


Fig.-1: Basic scheme of boost PFC

Two loops are here highlighted: the voltage loop and the current one. The reference of current (I_{ref}) is obtained by multiplying the output of the voltage regulator by a party ($K*V_{rd}$) of rectified voltage. The output of the current controller is treated in a shaping circuit CMF to obtain the signal u(t) used to control the static converter CS. In this paper, some schemes using the sliding mode control (SMC) are given and then a new controller based on hybridization of the SMC and polynomial RST controllers is proposed. First, the current loop is studied to obtain some conditions to have a perfect loop in comparison with loop voltage. Modeling of the loop voltage is then done for the RST and PI controllers' synthesis. Some schemes using SMC controllers follow this generality. The new controller named RST-SMC is finally proposed.

2. CURRENT LOOP ANALYSIS

Usually, two types of command are used the current control: the PWM command and the hysteresis control. Because of the nonlinear model of the converter static, the control by hysteresis is here chosen. The inductance boost L must be dimensioned according the chopping frequency. The expression of the frequency F_d is given according the relation (5) [6-7],

$$F_d = \frac{1}{T_d} = \frac{V_{rd}(V_s - V_{rd})}{2.\Delta I.L.V_s} \qquad (5)$$

The Fig. 2 shows curves giving F_d according L for imposed ΔI. In this case, V_s = 400 [V], V_{rd} =235 [V], ΔI = ± 0,1[A], ± 0,2[A], ± 0,3[A].

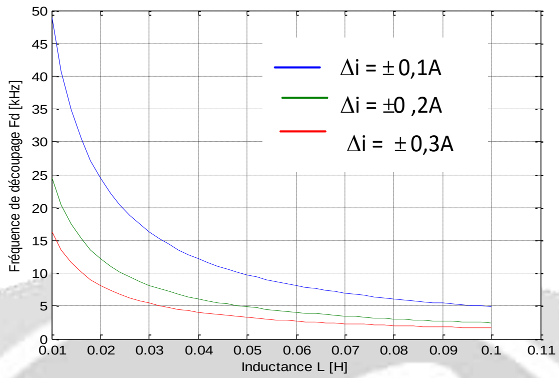


Fig.-2: F_d according the inductance L

3. LOOP VOLTAGE

It is here assumed that the current loop is faster than the voltage one and every time $I_{red} = I_{ref}$. The following approximation about the transfer function may be taken [8-9]]:

$$\begin{cases} \dfrac{V_s(p)}{I_{rd}(p)} = \dfrac{V_s(p)}{I_{ref}(p)} \\[2ex] \dfrac{V_s(p)}{I_{ref}(p)} = \dfrac{K}{1+pT} \end{cases} \qquad (6)$$

With, $$K = \frac{V_M.R}{4.V_s} \; ; \qquad T = \frac{R.C}{2} \qquad (7)$$

Here V_M, V_s, R and C denote respectively the effective value of the network voltage, the output voltage, the load resistance and the capacitor.

3.1 The polynomial controller RST

The RST controller is primarily a digital one. Its owes its name by the three polynomials which define it: *R(z)*, *T(z)* and *S(z)*. The command law is:

$$R(z).U(z) = T(z).Y_c(z) - S(z).Y(z) \qquad (8)$$

Here $Y_c(z)$ and *Y(z)* denote the set point and measured values. The basic idea for the RST synthesis is resumed in the Fig. 3.

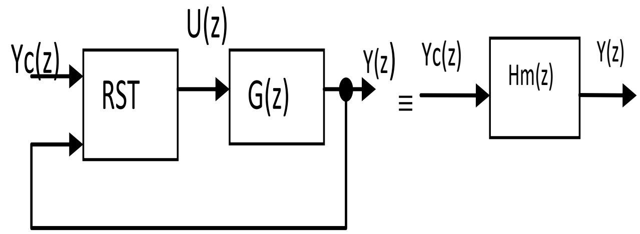


Fig -3: Basic scheme for RST synthesis

Having the function transfer *G(p)*given the second expression of the relation (7), the transfer function discrete *G(z)* is as follows:

$$G(z) = \frac{K.(1-e^{-h/T})}{z-e^{-h/T}} = \frac{b_0}{z-z_0} = \frac{B(z)}{A(z)} \tag{9}$$

With *h*, the sampling time.

In the Fig. 3, the closed loop transfer function $H_m(z)$ is chosen to respect the performances desired in closed loop. Its expression is given as:

$$H_m(z) = \frac{B_m(z)}{A_m(z)} \tag{10}$$

$H_m(z)$ must verify some conditions to respect all performances desired as said below. Using relations (8) and (10), the synthesis is obtained around the following expression:

$$\frac{T(z).B(z)}{A(z).R(z)+B(z).S(z)} = \frac{B_m(z)}{A_m(z)} \tag{11}$$

All steps to calculate *R(z)*, *S(z)* and *T(z)* are resumed in [9-10].

3.2 Sliding mode control

The basic idea of SMC is to bring a system in an area properly selected and then, design a control law to maintain the system in this area [11]. Usually, the SMC goes through three stages as follows:

- Switching surface choice
  The equation is given by the general form proposed in [12]:

$$S(X) = \left(\frac{d}{dt}+\lambda\right)^{n-1}.e \tag{12}$$

With $e = X_d - X$, the error, X_d and X the desired and the measured signals, λ a positive number and n the order system.

- Convergence condition

The convergence condition is defined by the Lyapunov equation (13); it is the condition to ensure the area to be attractive and invariant.

$$S(X).\overset{o}{S}(X) \leq 0 \tag{13}$$

- Control determination

The algorithm is given by the relation,

$$u = u^{eq} + u^{n} \tag{14}$$

Where u is the control signal, u^{eq} the equivalent control signal and u^{n} the switching control term.

The equivalent control signal is calculated with the conditions:

$$S(X) = 0 \quad \overset{o}{S}(X) = 0 \quad u^{n} = 0 \tag{15}$$

In the general cases, relation (16) gives the function for the switching control term,

$$\begin{cases} u^{n} = K.sign(S(X)) \\ u^{n} = K.sat(S(X)) \end{cases} \tag{16}$$

In [14], a proportional term is added with the first expression of the relation (16). It is made to increase the attractivity.

$$u^{n} = K.sign(S(X)) + K_1.S(X) \tag{17}$$

Where K and K_1 are constant positive.

In [15], the switching control term is replaced by a command resulting from fuzzy logic controller (FLC). It is showed by the relation (18):

$$\begin{cases} u = u^{eq} + u^{n} \\ u^{n} = u_{FLC} \end{cases} \tag{18}$$

3.3 Hybrid controller PI-SMC

The first hybrid controller is an extension of the relation (17) in which the proportional term is replaced by a PI controller. According relation (12) and for $n = 1$, the switching surface is:

$$S(X) = X_d - X = e \tag{19}$$

Fig. 4 shows how to built the control switching term u^{n}.

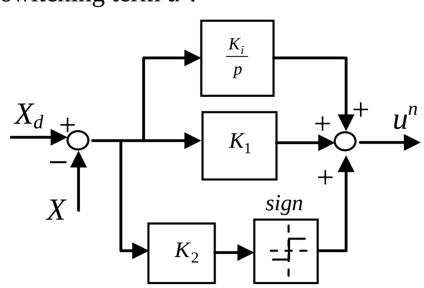


Fig-4: Scheme to build the control switching

The resulting controller is supposed to increase performances and to avoid the problem of chattering.

3.4 SMC-I controller

One possibility is to add an integrator with a gain at the output of the nonlinear element. It is not exactly the integral sliding mode method (ISM) because this last cited incorporates the integral in the surface. The switching term signal u^n is calculated by the usual way for the SMC. Fig-5 shows the proposal.

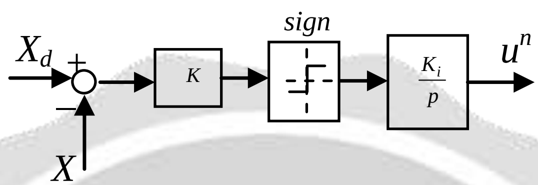


Fig-5: Scheme to build the switching term in SMC-I

3.5 The hybrid SMC-RST

This new hybrid controller is a derivative of the scheme proposed in [15]. Here the switching control term is issued of a polynomial RST controller. The relation (20) gives the expression:

$$\begin{cases} u = u^{eq} + u^n \\ u^n = u_{RST} \end{cases} \tag{20}$$

4. APPLICATIONS

It is already said that a hysteresis command is used for the current loop. To apply all methods cited above, controllers synthesis for the loop voltage may be done.

4-1 Determination if the equivalent control

The Fig. 6 shows the circuit modeling when the static converter (CS) is off.

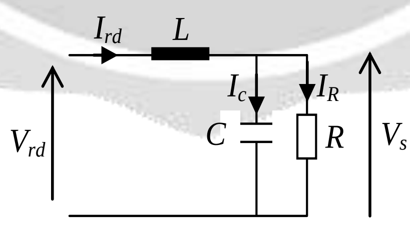


Fig.-6: Electric scheme when CS is off

The equations are resumed in relation (21).

$$\begin{cases} V_{rd} = L\dfrac{di_{rd}}{dt} + V_s \\ i_{rd} = i_C + i_R \\ i_C = C\dfrac{dV_s}{dt} \qquad i_R = \dfrac{V_s}{R} \end{cases} \tag{21}$$

For $n = 1$,

$$\begin{cases} S(V_s) = V_{sc} - Vs \\ \overset{o}{S}(V_s) = \overset{o}{V}_{sc} - \overset{o}{V}_s \end{cases} \tag{22}$$

With V_{sc}, the desired value. The two last expressions in the relation (21) give,

$$\overset{o}{V}_s = \frac{1}{C}\left[i_{rd} - \frac{V_s}{R}\right] \tag{23}$$

So,

$$\overset{o}{S}(V_s) = \overset{o}{V}_{sc} - \frac{1}{C}\left[i_{rd} - \frac{V_s}{R}\right] \tag{24}$$

Here,

$$i_{rd} = i_{rd}^{eq} + i_{rd}^{n} \tag{25}$$

Conditions $S(V_s) = 0$, $\overset{o}{S}(V_s) = 0$ $\quad i_{rd}^{n} = 0$ give,

$$i_{rd}^{eq} = C\overset{o}{V}_{sc} + \frac{V_s}{R} \tag{26}$$

4-2 PI controller synthesis

The transfer function is defined as: $\quad G_R(p) = \dfrac{1 + pT_n}{pT_i}$ (27)

Because of the expression of the transfer function system as, $G(p) = \dfrac{K}{1 + pT}$, given by relation (6), the method proposed in [16] is here adopted:

$$\begin{cases} T_n = a.T, & a \geq 0 \\ T_i = b.K.T, & b > 0 \end{cases} \tag{28}$$

For example, $a = 1$ means that the constant time T is cancelled.

4-3 RST synthesis

Because of the expression of $G(z)$, zero cancellation in not needed. In this case, a polynomial P(z) of degree 2 and one effect of perturbation compensation ($m = 1$) are chosen. For the application, computed results of the RST are as follows:

$$\begin{cases} h = 2\ [ms] \quad d^\circ P = 2 \quad m = 1 \quad \zeta = 0{,}3 \quad \omega_n = 300\ [rd.s^{-1}] \\ R(z) = z - 1 \\ S(z) = 0{,}349z - 0{,}170 \\ T(z) = 0{,}180 \end{cases} \tag{29}$$

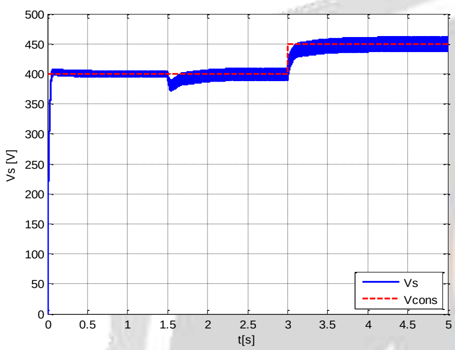
4-4 Simulation results

The condition tests are:

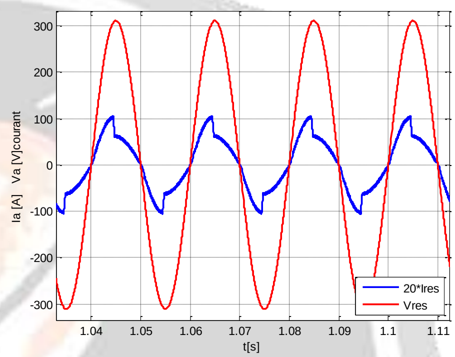
$$\begin{cases} t = 1{,}5\ [s] & R \to \dfrac{R}{2} \\ t = 3\ [s] & V_{sc} \quad 400[V] \to 450[V] \end{cases} \tag{30}$$

For the PI controller: $a = 1 \qquad b = 0{,}307$.

The simulation results are given by following figures.



(a)



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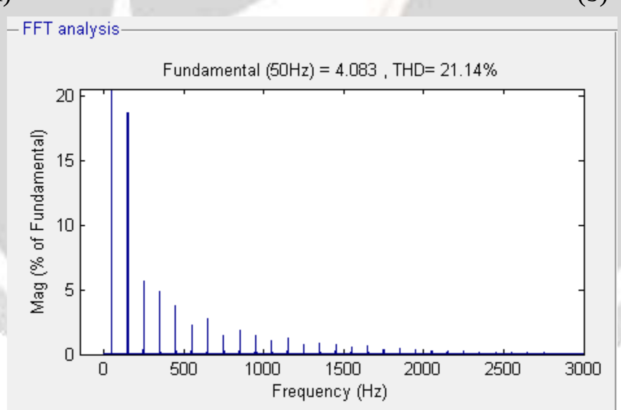
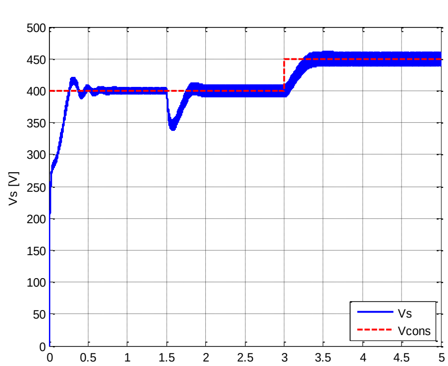
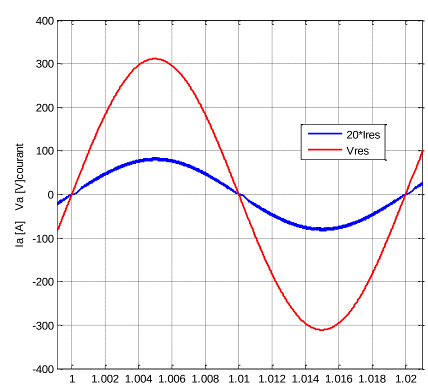


(c)

Fig.-7: Results with SMC-PI



(a)



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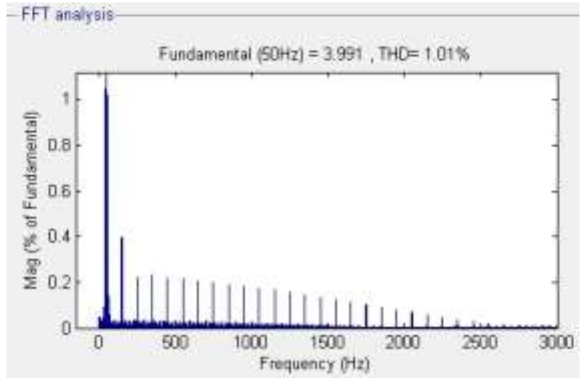

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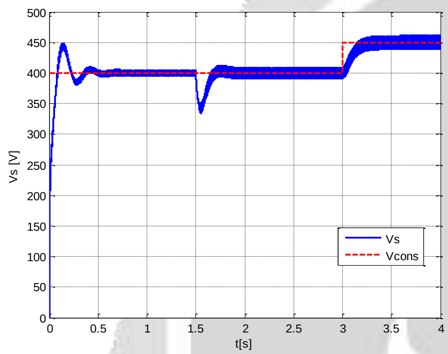
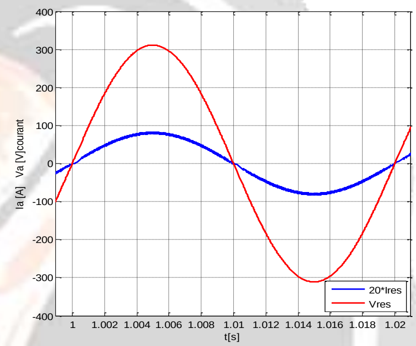
Fig.-8: Results with SMC-I



(a)



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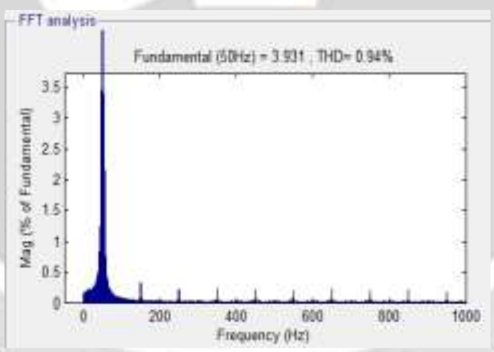

(c)

Fig.-9: Results with SMC-RST

For each case, output voltage *Vs*, the *TDH* , current and voltage curves (I_{rd},V_{rd}) are showed here. Effectively, they are sufficient to appreciate performances for the boost PFC. Table-1 resume the results.

Table-1:Comparison of the 3 controllers results

	PI-SMC	SMC-I	RST-SMC
D_1% t_p[s]	0 % -	5,2% 0,31[s]	12% 0,14[s]
Changing load (R)	$\begin{cases}\Delta P = 25[W] \\ \Delta t = 0,9[s]\end{cases}$	$\begin{cases}\Delta P = 60[W] \\ \Delta t = 0,6[s]\end{cases}$	$\begin{cases}\Delta P = 60[W] \\ \Delta t = 0,4[s]\end{cases}$
THD	21,14 %	1,01 %	0,94%

The PI-SMC is faster and doesn't present any overshoot (D_1 = 0%). It is less sensitive by resistance load variation but the reaction time to reach the reference is longer. Here, it is important to note that the THD is higher. It is not acceptable (*THD* = 21,14%). To reduce the THD, changing the constant ***b*** with higher value can be adopted. Effectively, the speed response depends of ***b.*** And it is well known, that the more the response is faster, the more the THD is higher.

RST-SMC presents an overshoot higher than the SMC-I one although the time peak (t_p) is shorter. With the same ΔP, during the load resistance change, the RST-SMC has a reaction much faster. The great difference resides in the THD.

5. CONCLUSION

In this paper, hybrid controllers based on SMC are presented. All these controllers are realizable. THD, set point tracking, load disturbance rejection and robustness are taken as criterion to compare the performances. Even, the overshoots with SMC-I and SMC-RST are higher, the THD are lower than with SMC-PI. The SMC-RST has good load disturbance rejection. We can conclude that the new hybrid controller, RST-SMC, presents the better performances. In accordance with this work, SMC-I and SMC-RST will be applied on SEPIC converter.

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