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FIXED FREQUENCY CONTROL LAWS FOR MULTI-CELL CHOPPER

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Abstract

In this paper, the authors present two fixed frequency control laws for the multicell chopper. The first is based on the partial linearizing feedback method associated with a linear control in series. The second is a fixed frequency control law involving a sliding mode approach. A broad part of this work is devoted to the characterisation of these two control laws in steady state and in transient state for a three-cell chopper.

Introduction

The multi-cell converters have been developed to increase the converted power. With an appropriate control, the performances in transient can be exceptional [1]. The goal of the control is to balance the flying capacitor voltages to a fraction of the voltage source (kE/p) and to adjust the current to a reference called I_{ref} with the largest possible bandwidth. The fixed frequency control laws for this converter are in general based on an average state equation [2], [3]. It is a non-linear state equation, with an input-output coupling. In this paper, two controls laws are presented. The first is based on linearizing feedback based on the average model, leading to a linear and decoupled transformed model. The linearizing function is computed and the linear control of the new model is defined in the general case. Some results of simulation are presented in the case of a three-cell chopper. In order to improve the bandwidth of the output current, a fixed frequency control law is defined : the fixed frequency sliding mode control. It is based on the use of switching functions defined for a variable frequency control associated with a modulator and PI regulator. Simulation will show the very good performances obtained with this control, in transient and in steady state.

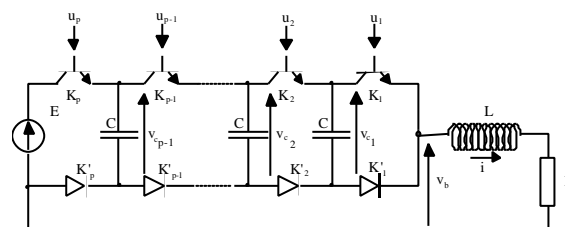


Fig. 1 - Structure of a p cells chopper

Modelling

Figure 1 presents a p -cell chopper. The instantaneous state equation is the one of the equation (1). It is a non linear equation with an input-output coupling.

$$\begin{cases} \dot{x} = f(x) + G(x)U = Ax + \sum_{i=1}^p g_i(x)u_i \\ y = h(x) \end{cases} \quad (1)$$

It is a square system because the number of input is equal to the number of state. The state vector is defined as (2).

$$x^T = [v_{C1}, \dots, v_{C(p-1)}, i] \quad (2)$$

The output of the system are equal to its states (h is the identity function). The inputs are the states of the switches (u_i). They can take two values: one if K_i is on (K'_i is off), and zero when K'_i is on (K_i is off). The matrix A , and $G(x)$ are defined by the equation (3).

$$A = \begin{bmatrix} 0 & \dots & \dots & 0 \\ \vdots & \ddots & \ddots & \vdots \\ \vdots & \ddots & 0 & 0 \\ 0 & \dots & 0 & -\frac{R}{L} \end{bmatrix} \quad (3) \quad G(x) = \begin{bmatrix} -\frac{i}{C} & \frac{i}{C} & 0 & \dots & 0 \\ 0 & -\frac{i}{C} & \frac{i}{C} & 0 & \vdots \\ \vdots & 0 & \ddots & \ddots & 0 \\ 0 & \dots & 0 & -\frac{i}{C} & \frac{i}{C} \\ \frac{V_{C1}}{L} & \dots & \dots & \frac{V_{Cp-1}-V_{Cp-2}}{L} & \frac{(E-V_{Cp-1})}{L} \end{bmatrix}$$

From this model it is possible to generate an average model that represents an average state variable evolution. It is the same type of equation as (1) but in this case, it is not the instantaneous value which are involved but the average value of each state (4). The inputs are now the values of the duty cycles applied to each cell. This model is always non-linear and the output is still coupled to the duty cycle.

$$\begin{cases} \langle \dot{x} \rangle = f(\langle x \rangle) + G(\langle x \rangle)\langle u \rangle \\ \quad = A\langle x \rangle + \sum_{i=1}^p g_i(\langle x \rangle)a_i \\ \langle y \rangle = h(\langle x \rangle) \end{cases} \quad (4)$$

Using the average model of the chopper causes a loss of information. Indeed, the natural balancing of the floating voltage v_{ck} could not be explained by [4].

Linearizing feedback

The goal of the linearizing function is to transform (4) in a linear state equation. We want to control the chopper with large variations of the output current, and for this reason we choose to use a decoupling feedback. Figure 2 presents its structure. By using the two functions \mathbf{a} and \mathbf{b} , we obtain a linear system comprising $p-1$ parallel integrator and one first order system. The partial linearizing feedback is interesting because the decoupling function does not depend on the load which may vary in a very large range.

In our case, the matrix \mathbf{b} is equal to the inverse of matrix G and α is equal to zero. The linearizing feedback exists if the rank of G is equal to p . The equation (5) defines the value of the determinant of G .

$$\det(G(\langle x \rangle)) = (-1)^{p-1} \frac{i^{p-1} E}{\prod_{k=1}^p C_k L} \quad (5)$$

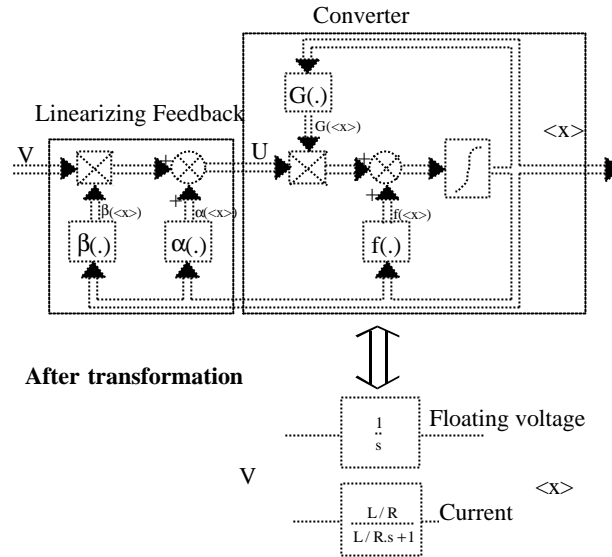


Fig. 2 - Structure of the partial linearizing feedback

The rank of G will be of p if the output current i and the input voltage E are different from zero. It is always true in the chopper mode. In this case, the system is also controllable. So, we can define a linear control based on the new system of state equation (6).

$$\dot{\langle x \rangle} = A \langle x \rangle + V \quad (6)$$

It should be noted that the new inputs have units. Those controlling the flying voltages are written in volts per second, and those adjusting the current are in amps per second.

Linear control

Let's consider that the values of passive element are known exactly. In open loop, we can write the transfer functions that link the input v_i to the output x_i as :

$$\begin{cases} \text{if } k \neq p & H_{OL}^k(s) = \frac{v_{Ck}}{v_k} = \frac{1}{s} \\ \text{if } k = p & H_{OL}^p(s) = \frac{i}{v_p} = \frac{t_0}{t_0 s + 1} \end{cases} \quad (7)$$

with $t_0 = L/R$

The flying voltages can be controlled with a proportional regulator (K_k), and the output current with an IP (K_p, τ_i) regulator. In closed loop, the system will be constituted of $p-1$ first order transfer function (for the flying voltages) with the same dynamic (τ_v) and a second order transfer function for the current (m, ω_n).

$$\begin{cases} \text{if } k \neq p & H_{CL}^k(s) = \frac{v_{Ck}}{v_{refk}} = \frac{1}{t_v s + 1} \\ \text{if } k = p & H_{CL}^p(s) = \frac{i}{I_{ref}} = \frac{(w_n)^2}{s^2 + 2m.w_n.s + (w_n)^2} \end{cases} \quad (8)$$

The parameters of the regulator are easy to calculate and they are defined in table I.

	Gain	Time
if $k \neq p$	$K_k = \frac{1}{t_v}$	
if $k = p$	$K_p = 2.m.w_n - \frac{1}{t_0}$	$t_i = \frac{K_p}{w_n^2}$

Table I : Values of the regulator's parameters

Control of a three-cell chopper

The function \mathbf{b} is defined by equation (9). The nominal value of the components are: $C=10 \text{ mF}$, $L=1.5 \text{ mH}$, $R= 20 \text{ W}$. The frequency of switching is equal to 16 kHz .

$$\mathbf{b} = \begin{bmatrix} \frac{\langle v_{c1} \rangle - E}{\langle i \rangle \cdot E} \cdot C & \frac{\langle v_{c2} \rangle - E}{\langle i \rangle \cdot E} \cdot C & \frac{L}{E} \\ \frac{\langle v_{c1} \rangle}{\langle i \rangle \cdot E} \cdot C & \frac{\langle v_{c2} \rangle - E}{\langle i \rangle \cdot E} \cdot C & \frac{L}{E} \\ \frac{\langle v_{c1} \rangle}{\langle i \rangle \cdot E} \cdot C & \frac{\langle v_{c2} \rangle}{\langle i \rangle \cdot E} \cdot C & \frac{L}{E} \end{bmatrix} \quad (9)$$

Let us first examine the voltage loop. The worst case for the proportional regulator is when the capacitor is known with the bigger uncertainty. We want to have the more dynamic control of the voltages. So, the gain must be calculated when the capacitor is the smallest. For the control of the current loop, the problem is identical. The regulator must be calculated for the smallest value of the inductor.

The dynamic of the system in closed loop must be slower than the switching period to respect the average model assumption used for this law. We chose to have in closed loop the transfer functions defined by (8) with:

$$t_v = 0.1 \text{ ms} \quad \text{and} \quad w_n = 20000 \text{ rd/s} \quad m = 0.7 \quad (10)$$

This control has been tested on a sequence with variation of I_{ref} , E and the load.

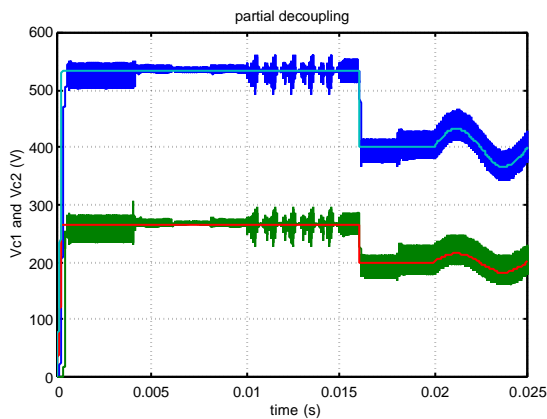


Fig. 3 Floating voltages Vc1, Vc2 and references

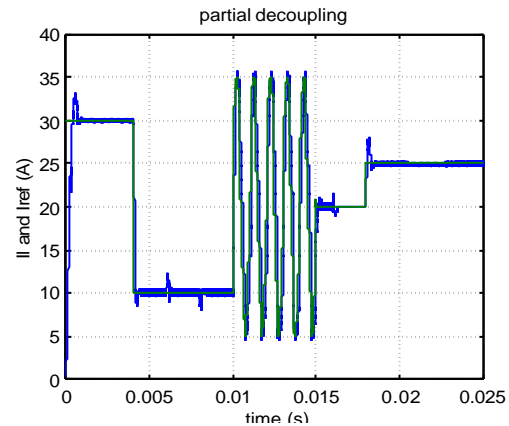
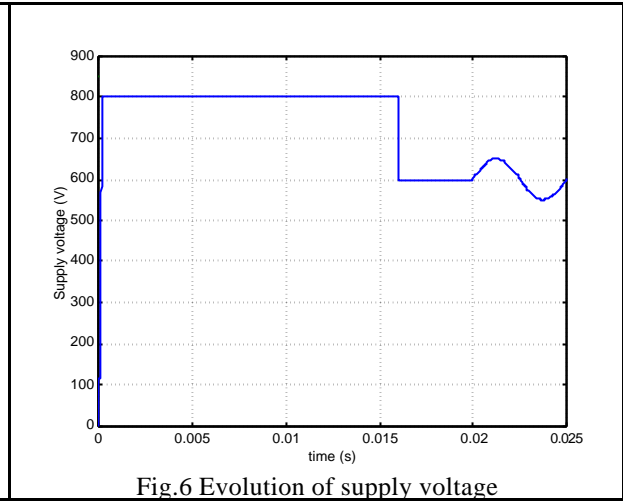
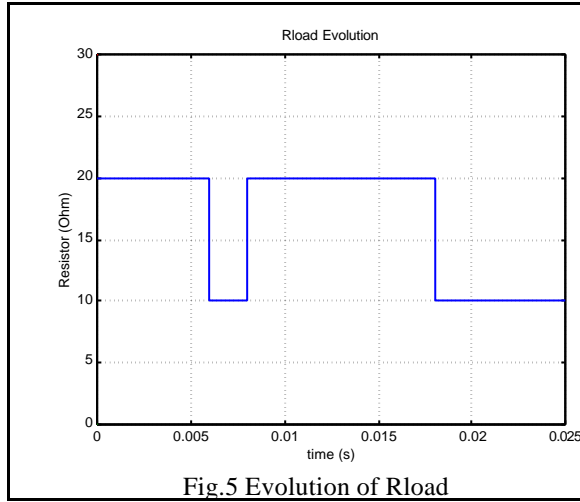


Fig.4 Output current and reference



The load changes from +100% to -100% of the nominal value. The supply voltage undergoes a sharp variation then a sinusoidal disturbance.

The output current is always conform to the reference I_{ref} with any error. This control is slightly sensitive to the variation of the resistor. Generally, the implantation of this control needs to use numerical hardware. We now present a technique of control with a simple design and easy implantation: this is a sliding mode control at fixed frequency.

Sliding mode control with fixed frequency

In a former paper [5], the authors have already presented some variable frequency sliding mode control. The performances of those controls are excellent in transient. The insensitivity to parametric variations is also very good. In steady state, a drawback appears. The phase-shift between the cells is not fixed directly, and it is necessary to use a specific mode of switching in steady state. In this paper, the authors present a technique to transform a variable frequency control into a fixed frequency control law. For this, we are going to use the switching functions defined by using the Lyapunov criterion, the stability of the system in closed loop is preserved. They are defined by equation (11).

$$\begin{aligned}
 S_p(x) &= \frac{p \cdot I_{ref}}{E} \cdot (E - V_{cp-1}) - i \\
 \text{if } k \in [2, p-1] S_k(x) &= \frac{p \cdot I_{ref}}{E} \cdot (V_{ck} - V_{c(k-1)}) - i \\
 S_1(x) &= \frac{p \cdot I_{ref}}{E} \cdot V_{c1} - i \quad (10)
 \end{aligned}$$

In steady state, the functions S_k represent the error on the current. By adding a modulator, and a PI regulator, the duty cycle will be fixed in steady state. Figure 7 presents the structure of the control of the cell number k . The complete structure is composed of p identical blocks. Its outputs are the duty cycles, and its inputs are the switching functions. Let's note that the modulator is a function that can give p control signals for the cells with a fixed phase shift ($2 \cdot \pi / p$) between them.

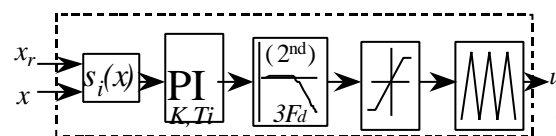


Fig.7 Structure of the control law

The regulator is calculated by studying the ripple of the switching function. The value of the gain can be calculated according to equation (12).

$$K_k < \frac{1}{1+V_{sat}} \quad (12)$$

The value V_{sat} is the height of the saturation block connected after the integrator, chosen in adequation with the sawtooth amplitude. The last parameter to define is the integration time constant. The goal of this is to create a duty cycle and to cancel the error in steady state, normally introduced by the frequency constant operating mode. So, the constant must be chosen to be slower than the switching period, in order to converge towards the state of balance without losing the properties of the sliding mode.

With this type of control, the performances in transient are good but not as good as those obtained with a free-frequency control. However by fixing the phase-shift between the control signal, the waveforms in steady state are always optimal. Let's now apply this type of control to a three-cell chopper using the same benchmark profile as for the partial linearizing feedback control.

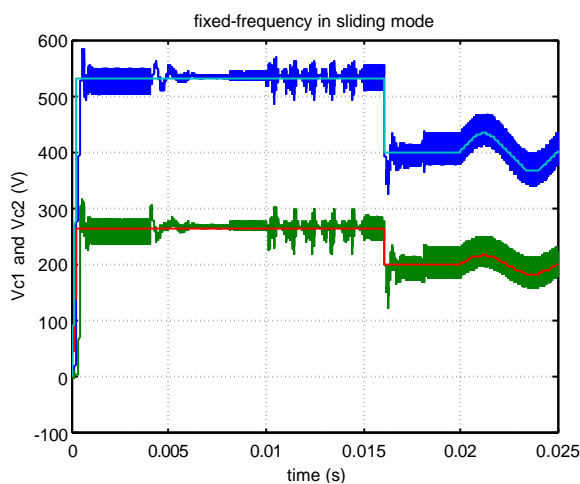


Fig.8 Flying voltages Vc1, Vc2 and references

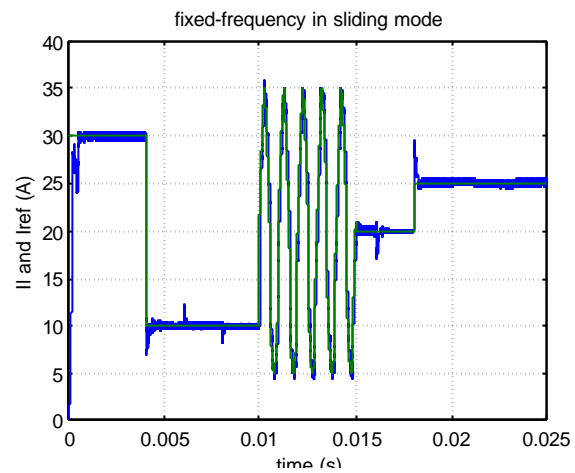


Fig. 9 Output current and reference

The results are slightly different from those obtained with the partial decoupling control law. In the transient, the performances are good, and for sinusoidal evolution of the current reference the behaviour is interesting.

By way of comparison we give here the results obtained with a control law in sliding mode in free-frequency operation. The switching functions are identical to the preceding case.

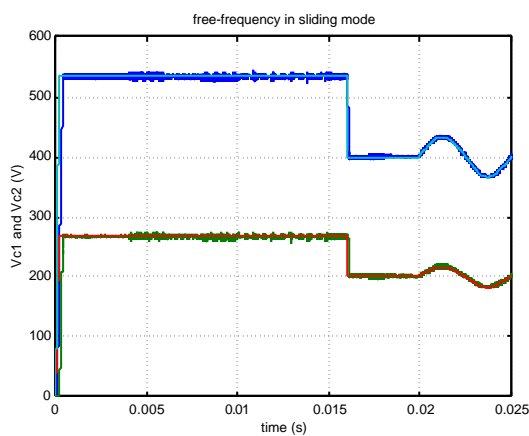


Fig.10 Flying voltages Vc1, Vc2 and references

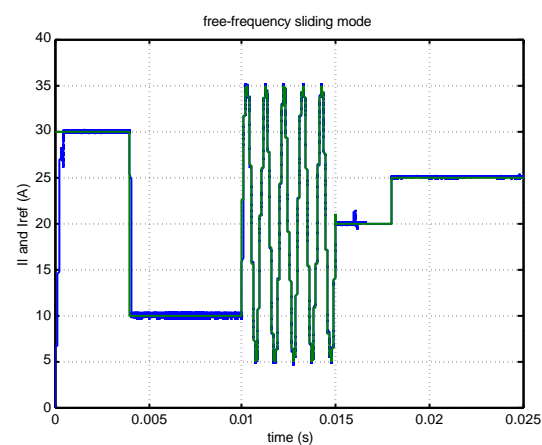


Fig. 11 Output current and reference

This operation is optimal for the dynamic consideration with a very good robustness compared to the load but the phase shift between the cells is not always guaranteed [5].

Conclusion

In this paper, two control strategies with fixed switching frequency are defined for the control of a multicell chopper. The first is based on partial linearizing feedback and the second uses the switching functions according to a sliding mode approach with fixed frequency. Both controls are very good. The first is better for step reference variations (supply voltage and reference current) and the second is better for sinusoidal reference and parametric variations. The control law in the sliding mode approach at constant frequency is easily achieved with analog circuits and the control law with partial decoupling is more fit for a numerical implementation. These control laws are simply generalizable when the number of cells increases, they constitute a whole set of solutions making it possible to answer various problems.

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