Abstract - Non-linear phenomena occurrence in DC-DC power converters makes them difficult to analyse and to control. We propose to refine the boost converter map-based model by including all converter elements in order to obtain a more accurate model. It is then used to synthesise a fuzzy logic controller ensuring the converter current regulation in wide range of parameters variation and to shift the non-linear phenomena outside the operating domains.

Keywords: Hybrid system, boost converter, map-based model, fuzzy logic controller, non-linear phenomena, bifurcation.

I. INTRODUCTION

Hybrid dynamical systems are defined as an interaction of continuous and discrete subsystems, associating continuous and discrete control as well as continuous and discrete dynamics [ANT, 00], [PET, 99], [SCH, 00].

DC-DC converter can be regarded as a hybrid dynamical system. Indeed, obtaining a desired output requires the switching between various systems configurations, according to the conduction modes (continuous and discontinuous). In the discontinuous conduction mode (DCM), three topologies can be distinguished whereas only two can be identified in the continuous conduction mode (CCM).

A modelling step of the power converter is necessary to synthesise a suitable controller that guarantees desired performances. The obtained model must accurately describe the system behaviour by considering all converter elements and reducing the simplifying assumptions.

Among developed converter models, we can mention the small signals modelling approach which allows to describe the system behaviour around a given operating point [ERI, 99], [BAN, 01] and to design the adequate controller [AHM, 03], [DIO, 03]. Nevertheless, this approach is not sufficient to explain the observed abnormal behaviours of the converters. To solve this problem, the iterated non-linear mapping method can be used [CHE, 98], [TSE, 02]. This method allows obtaining a non-linear iterative function (map) that expresses the state variables at one sampling instant in terms of those at an earlier sampling instant. Due to the non-linearity of the DC-DC converter, the mapping technique is most accurate to explore the non-linear phenomena of this class of systems. Indeed, [DEA, 90] use this modelling approach to show the existence of sub-harmonic oscillations, quasi periodicity and chaotic phenomena in DC-DC converters. In [HAM, 92] the discrete mapping method is used to analyse DC-DC converters abnormal behaviour and stability conditions are given for the resulting model. In [BAN, 98], the authors analyse the different non-linear phenomena in boost converter. Even though the previous papers explored the different abnormal behaviours in converters, they relied mainly on ideal elements (neglect internal resistance, ideal switch, instantaneous commutation) which, in most cases, affect the model accuracy.

Using the discrete mapping technique, several controllers can be found in the literature [POD, 98] [CHA, 95]. Feedback controllers [BAN, 98], are widely used due to their low cost and simplicity of implementation. However the main inconvenient of such controllers is that non-linear phenomena may appear under circuit parameters variation in the operating domain making system analysis difficult. To overcome these limitations, other approaches can be used. PID controller, as an example, can provide better performances but for a restricted domain around the operating point. In this paper, we propose to use a fuzzy controller to uphold performances in a wide range of operating points and to shift non-linear phenomena out of the operating domain.

This paper deals with the control of a boost converter operating in CCM. We, thus, propose to refine the boost converter discrete model given in [BAN, 98] by including the internal resistors of the switching elements. First, the converter is controlled by a feedback law to explore its behaviours and to present the effects of the internal resistors. Then, based on Takagi-Sugeno system a fuzzy logic controller is designed to regulate the current in the converter inductance element and to ensure the non-linear phenomena shifting.

II. BOOST CONVERTER DISCRETE MODEL

This section is dedicated to the elaboration of a refined discrete model of the boost converter shown in figure 1, where $r_L$, $r_{SW}$, $r_{VD}$ and $r_C$ denote the resistors of inductor $L$, switch $sw$, diode $VD$ and capacitor $C$, respectively. $R$ is the load, $v_s(t)$ the supply voltage, $u_r(t)$ the output voltage and $i_L(t)$ the current in inductance $L$. The clock period is given by $T$ and the desired current is $I_{ref}$.
In CCM i.e., \( i_e(t) > 0 \), we have two topologies related to the switch \( sw \) position, as shown in figure 2. From this point of view, the converter can be regarded as a hybrid system. Indeed, in each topology the system can be described by a set of continuous differential equations and the transition (discrete event) from the first configuration to the second is conditioned by \( (i_e=I_{ref}) \) then the condition \((t \mod T)=0\) is required to return back to the first configuration (Fig. 3).

In the first configuration the switch \( sw \) is closed (Fig. 2-a) and the current \( i_L \) increases continuously until it reaches a reference current value \( I_{ref} \). At this point, the switch \( sw \) will be opened (Fig. 2-b) by the controller to enable the inductance to discharge and the capacitor to charge until the next clock pulse.

\[
\begin{align*}
\frac{d}{dt}i_e(t) &= v_e(t) - (r_e + r_{sw})i_e(t) \\
\frac{d}{dt}v_c(t) &= -\frac{1}{(R + r_c)}v_c(t)
\end{align*}
\]

where \( v_c(t) \) is the voltage across the capacitor \( C \).

In the second topology (Fig. 2-b), we have:

\[
\begin{align*}
\frac{d}{dt}i_e(t) &= v_e(t) - (r_e + r_{sw})i_e(t) \\
C\frac{d}{dt}v_c(t) &= -\frac{1}{(R + r_c)}(R\cdot i_e(t) - v_c(t))
\end{align*}
\]

The different possibilities of system states \((i_e(t), v_c(t))\) evolution are given by figure 4. At the \( n^{th} \) clock cycle \((t \in [nT, (n+1)T))\), the current \( i_L \) increases in the first configuration according to:

\[
i_e(t) = i_e(n)e^{\frac{-(t - nT)}{\tau_e}} + v_f\left(1 - e^{\frac{-(t - nT)}{\tau_e}}\right)
\]

and the voltage \( v_c(t) \) decreases as follows:

\[
v_c(t) = v_c(n)e^{\frac{-(t - nT)}{\tau_c}}
\]

Using (2-a) and (2-b), the voltage across the capacitor will be given by the solution of the following equation:

\[
\frac{d^2}{dt^2}v_c(t) + a_1\frac{d}{dt}v_c(t) + a_2v_c(t) = a_i
\]

with \( a_1 = \frac{Rv_f}{LC(R + r_c)} \), \( a_2 = \frac{R + r_c + r_{sw}}{LC(R + r_c)} \) and

\[
a_i = L + C\left[Rc_{1} + (R + r_c)(r_c + r_{sw})\right]/LC(R + r_c)
\]

when \( t_n < T \), for example the \( n^{th} \) clock cycle of Fig. 4 \((t \in [nT, (n+1)T))\), we can distinguish three roots for (4) according to the sign of the discriminant \( \Delta \) of the corresponding characteristic equation.

If \( \Delta > 0 \), the solution is

\[
v_c(t) = C_1e^{\lambda_1t} + C_2e^{\lambda_2t} + v_c
\]

where \( \lambda_1 \) and \( \lambda_2 \) represent the roots of the corresponding characteristic equation of (4), \( C_i = v_c(t_n) - v_c - C_1 \),

\[
C_2 = \frac{1}{\lambda_2 - \lambda_1} \left[ \frac{Rv_f(\tau_1)}{C\tau_1} - v_c\left(1 + \frac{1}{C(R + r_c)}\lambda_1\lambda_2\right)\right]
\]

and \( v_c = a_i/a_2 \)
is the particular solution of (4).
At the end of the $n^{th}$ clock cycle (Fig. 4), we have:

$$v_c(n+1) = C_e e^{j\omega \tau} + C_i e^{j\omega \tau} + v_c$$  \hspace{1cm} (5-a)

Using (2-b), the current $i_L$ is given by:

$$i_L(n+1) = \frac{1}{R} C_e e^{j\omega \tau} [\lambda C(R + \tau) + 1] + \frac{1}{R} C_i e^{j\omega \tau} [\lambda C(R + \tau) + 1] + \frac{\nu}{R}$$  \hspace{1cm} (5-b)

In the same way, if $\Delta = 0$, the system can be described by:

$$v_c(n+1) = (C_e + C_i) e^{j\omega \tau} + v_c$$  \hspace{1cm} (6-a)

$$i_L(n+1) = \frac{1}{R} e^{j\omega \tau} \left[ C_e e^{j\omega \tau} [\lambda C(R + \tau) + 1] + C_i e^{j\omega \tau} [\lambda C(R + \tau) + 1] + \frac{\nu}{R} \right]$$  \hspace{1cm} (6-b)

where

$$C_2 = \frac{1}{C(R + \tau)} \left[ R (i_L(t_c) - v_c) + C_i \left( \frac{\nu}{2} C(R + \tau) - 1 \right) \right]$$

and

$$C_1 = v_c(t_c) - v_c.$$  \hspace{1cm} (7-a)

If $\Delta < 0$, the system states become oscillatory and are given by:

$$v_c(n+1) = e^{j\frac{\pi}{\Delta \omega}} \left[ C_e \cos(\omega \tau) + C_i \sin(\omega \tau) \right] + v_c$$

$$i_L(n+1) = e^{j\frac{\pi}{\Delta \omega}} \left[ C_e \cos(\omega \tau) + C_i \sin(\omega \tau) \right]$$

$$+ e^{j\frac{\pi}{\Delta \omega}} \left[ C_e \cos(\omega \tau) + C_i \sin(\omega \tau) \right] + \frac{\nu}{R}$$  \hspace{1cm} (7-b)

with

$$\omega = \left( \frac{u_i - u^2}{4} \right), \hspace{1cm} C_1 = v_c(t_c) - v_c,$$

and

$$C_2 = \frac{1}{C(R + \tau)} \left[ R (i_L(t_c) - v_c) + C_i \left( \frac{\nu}{2} C(R + \tau) - 1 \right) \right].$$

In the case where $t_c \geq T$ (Fig. 4, $t \in [(n+1)T, (n+2)T]$), the system can be described by:

$$v_c(n+1) = v_c(n) e^{j\frac{\pi}{\Delta \omega \tau}}$$

$$i_L(n+1) = \frac{V_e}{i_c + s_{\omega}} \left( \frac{V_e}{i_c + s_{\omega}} - i_L(n) \right) e^{j\frac{\pi}{\Delta \omega \tau}}$$  \hspace{1cm} (8-a)

In the next section, this refined discrete model is used to explore the different system behaviours using a feedback controller first, then we'll synthesise a suitable fuzzy logic controller in order to ensure the current mode control of the converter in wide range of parameters variation.

**III. CONTROLLER SYNTHESIS**

With the aim to regulate the converter current $i_L$, as shown in figure 4, the controller has to determine the value of the "dwell time" $t_d$ or the duty cycle $0 \leq d = t_d/T \leq 1$ to ensure that the current value in inductance $L$ is as close as possible to a reference current value $I_{ref}$ across a wide range of system parameters variation. We propose in this section to synthesise a fuzzy logic controller (FLC) to achieve this objective.

Several structures similar to a classic PID (fuzzy PID, fuzzy pseudo-PID) can be found in the literature [YIN, 00], [MAN, 99], having the error ($e$), its variation ($\Delta e$) and its rate of variation ($\Delta^2 e$) as inputs. However, the rule base will be of three dimensions which makes the controller design difficult. Several solutions can be considered to overcome this problem [MAN, 99]. Among which the use of two fuzzy blocks with two inputs in which controller output is the weighted sum of their outputs. Based on this idea as well as the desire to reduce the computing time, only one fuzzy system is used to synthesise the controller as shown in figure 6. This fuzzy system (FLS) is a Takagi-Sugeno type, having the current error $e(n) = I_{ref} - i_L(n)$ and its variation $\Delta e(n) = e(n) - e(n-1)$ as inputs and $\delta d$ as output.

Based on human expert information, the rule base of the FLS is a collection of fuzzy rules where the $j^{th}$ component is given by:

**IF** $E$ **is** $E_1^{j}$ **AND** $\Delta E$ **is** $E_2^{j}$ **THEN** $\delta d = C_j(E, \Delta E)$

with $E = G_e e$, $\Delta E = G_e \Delta e$, and $E_1^{j}, E_2^{j}$ are respectively the membership sets of the scaled error $E$ and its variation $\Delta E$. $C_j$ is an output singleton.

![Fig. 5: Boost converter control scheme](image)

![Fig. 6: Proposed fuzzy logic controller structure](image)

Using the product as inference engine and the centre average for defuzzification, the FLS output can be formulated as follows:

$$\delta d = \frac{\sum_{j=1}^{N} C_j(E, \Delta E) \mu_j(E) \mu'_j(\Delta E)}{\sum_{j=1}^{N} \mu_j(E) \mu'_j(\Delta E)}$$  \hspace{1cm} (9)

where $N$ is the number of used rules and $\mu'_j(x)$ is the membership degree of input element $x$ to the set $E_1^{j}$. Using the weighting gains $G_1$ and $G_2$, the proposed controller output is given by:

$$\delta d = \frac{\sum_{j=1}^{N} C_j(E, \Delta E) \mu_j(E) \mu'_j(\Delta E)}{\sum_{j=1}^{N} \mu_j(E) \mu'_j(\Delta E)}$$  \hspace{1cm} (9)
\[ d = G_0 \delta d + G_2 \sum \delta d \]  

We note that the use of gains \( G_1 \) and \( G_2 \) gives more flexibility to the controller and allows determining the participation rate of each action in the controller output to achieve the desired performances.

### IV. SIMULATION AND RESULTS

This section is subdivided in two parts. Different behaviours of the boost converter, the internal resistors effect and the feedback controller limitations are presented in the first part. The proposed fuzzy logic controller is used in the second part, to ensure the converter current regulation in a wider range of the system parameters variation. For this, we use the boost converter of figure 1, with the following characteristics \( r_{VD}=0.24 \Omega, r_{SW}=0.3 \Omega, r_p=1.2 \Omega, r_C=0.1 \Omega, C=120 \mu F \) and the switching frequency \( f_{sw}=1/500Hz \).

#### A. Internal resistors effects

In this part, the internal resistors effects will be examined, and the feedback controller limits will be shown using the bifurcation diagram in the four cases of parameters variation given in table I.

\[
\begin{array}{|c|c|c|c|}
\hline
\text{Param.} & V_g[V] & R[\Omega] & L[mH] & L_{ref}[A] \\
\hline
V_g & [7, 50] & 20 & 27 & 4 \\
R & 30 & [8, 50] & 27 & 4 \\
L & 20 & [1, 30] & 4 \\
L_{ref} & 30 & 20 & 27 & [1.4, 7] \\
\hline
\end{array}
\]

In case where \( t_n < T \), the feedback action is computed using (1-a) and given by:

\[
t_n = \frac{L}{(L + r_{SW})} \ln \left( \frac{V_g - r_{VD}I_n}{V_g - (L + r_{SW})I_{ref}} \right) \tag{11}
\]

and when \( t_n \geq T \) the feedback controller output will be limited to \( I_n = T \).

Using this controller with the simplified model (\( r_{SW} = r_{VD}=0 \)) [BAN, 98] the quantification of the obtained results is given in the upper part of table II, whereas the lower part is dedicated to the refined model (\( r_{VD}=0.24 \Omega, r_{SW}=0.3 \Omega \)).

Figure 7 gives the bifurcation\(^1\) diagrams for the refined model. The different behaviours (periods\(^3\)) of the system (P1,...,P8 and chaos\(^3\)) appear clearly in figure 7-b.

\(^1\) Bifurcation: is an abrupt change in the qualitative behaviour of the system as a parameter is varied; for example, period doubling [DEA, 90].

\(^2\) Period: a system with a Pn (Period - n) behaviour means that the system steady state period \( P=nT \).

\(^3\) Chaos: can be described as a bounded oscillation with an infinite period, found in nonlinear, deterministic systems and characterized by extreme sensitivity to initial conditions, i.e., infinitesimal perturbation of initial condition lead to diverging solutions [DEA, 90]. In regulation, this behaviour is undesired due to the fact that the system behaviour becomes unpredictable.

<table>
<thead>
<tr>
<th>Variation of</th>
<th>P1</th>
<th>P2</th>
<th>P4</th>
<th>P8</th>
<th>Chaos</th>
</tr>
</thead>
<tbody>
<tr>
<td>Variation of</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Simplified model behaviour</td>
<td>F</td>
<td>35.1, 50</td>
<td>[24, 35.1]</td>
<td>[22.3, 24]</td>
<td>[21.7, 22.3]</td>
</tr>
<tr>
<td>Variation of</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>R</td>
<td>8</td>
<td>12.8</td>
<td>[12.8, 31.6]</td>
<td>[11.6, 37.4]</td>
<td>[37.4, 38.9]</td>
</tr>
<tr>
<td>Variation of</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>P</td>
<td>1.43</td>
<td>4.3</td>
<td>11.5</td>
<td>16.1</td>
<td>16.1</td>
</tr>
<tr>
<td>Variation of</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>L_{ref}</td>
<td>1.4, 3.4</td>
<td>3.4, 4.96</td>
<td>4.96, 5.37</td>
<td>[5.37, 5.52]</td>
<td>[5.53, 5.62]</td>
</tr>
</tbody>
</table>

From table II, we can remark that in period-one (P1) region, the refined model (Fig. 7) has the same behaviour as the simplified one with small broadening in the case of inductance variation (table II). But if it is required to design a system that operates in P2, P4 or in P8 region, the internal resistors effect becomes more and more important. For example, if it is necessary that the system operates in P2 region with the refined model, the load must be less than 30 \( \Omega \) otherwise the P4 region will be reached. However, for the simplified model this limit value is shifted to 31.6 \( \Omega \) (table II) which is situated in the P4 region of the refined model. Therefore, it is suitable, in these regions to use the refined model.

As illustrated previously, the feedback controller can not ensure a large-enough P1 region in view to simplify the analysis and prediction of system's behaviour. Thus based on the refined model that describes the system accurately, we propose in the next paragraph to synthesise a suitable fuzzy logic controller that ensures boost current regulation, non-linear phenomena shifting and P1 region widening.

#### B. Non-linear phenomena shifting

The proposed controller is designed as follows: the FLS block (Fig. 6) must be constructed first in which the inputs \( e \) and \( \Delta e \) are scaled respectively by \( G_1=0.2 \) and \( G_2=1 \). For each input, we define five fuzzy sets: Negative Large (NL), Negative (N), Zero (Z), Positive (P) and Positive Large (PL), using triangular membership functions, uniformly distributed over the normalised universe of discourse. The fuzzy reasoning part in this structure is formed by all possible combinations depicted in table III (25 fuzzy rules with 17 output singletons). To obtain a converter current close to the reference at each sampling instant, we reinforce the integral (summation) action by choosing the gains \( G_1=0.1 \) and \( G_2=150 \).

<table>
<thead>
<tr>
<th>Variation of</th>
<th>P1</th>
<th>P2</th>
<th>P4</th>
<th>P8</th>
<th>Chaos</th>
</tr>
</thead>
<tbody>
<tr>
<td>Variation of</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Simplified model behaviour</td>
<td>F</td>
<td>35.1, 50</td>
<td>[24, 35.1]</td>
<td>[22.3, 24]</td>
<td>[21.7, 22.3]</td>
</tr>
<tr>
<td>Variation of</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>R</td>
<td>8</td>
<td>12.8</td>
<td>[12.8, 31.6]</td>
<td>[11.6, 37.4]</td>
<td>[37.4, 38.9]</td>
</tr>
<tr>
<td>Variation of</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>P</td>
<td>1.43</td>
<td>4.3</td>
<td>11.5</td>
<td>16.1</td>
<td>16.1</td>
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<tr>
<td>Variation of</td>
<td></td>
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<td></td>
</tr>
<tr>
<td>L_{ref}</td>
<td>1.4, 3.4</td>
<td>3.4, 4.96</td>
<td>4.96, 5.37</td>
<td>[5.37, 5.52]</td>
<td>[5.53, 5.62]</td>
</tr>
</tbody>
</table>
The fuzzy nature of the proposed controller permits the elimination of the system response variation in steady state. It allows also the widening of the P1 region (non-linear phenomena shifting). Nevertheless, a steady state error persists and can be eliminated by the insertion of an integral action.

Figure 8 illustrates the system steady state response for the point \( V_g = 20v \), corresponding to a chaotic behaviour of the system (figure 7-a). The application of the FLS output, \( \delta d \), results in having the same current value at each sampling instant (regularity) as shown in figure 9. The desired response as illustrated in figure 10 is obtained by adding the integral element and using the complete action, \( d \) (eq. 10).

To assess the performances of the proposed controller, we use larger operating domains, given by table IV.

### TABLE III

**INFEERENCE MATRIX**

<table>
<thead>
<tr>
<th></th>
<th>NL</th>
<th>N</th>
<th>Z</th>
<th>P</th>
<th>PL</th>
</tr>
</thead>
<tbody>
<tr>
<td>PL</td>
<td>0.25</td>
<td>0.36</td>
<td>0.49</td>
<td>0.81</td>
<td>1</td>
</tr>
<tr>
<td>P</td>
<td>0</td>
<td>0.04</td>
<td>0.16</td>
<td>0.36</td>
<td>0.64</td>
</tr>
<tr>
<td>Z</td>
<td>-0.16</td>
<td>-0.04</td>
<td>0</td>
<td>0.04</td>
<td>0.16</td>
</tr>
<tr>
<td>N</td>
<td>-0.64</td>
<td>-0.36</td>
<td>-0.16</td>
<td>-0.04</td>
<td>0</td>
</tr>
<tr>
<td>NL</td>
<td>-1</td>
<td>-0.81</td>
<td>-0.49</td>
<td>-0.36</td>
<td>-0.25</td>
</tr>
</tbody>
</table>

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Fig. 7 Bifurcation diagrams using the refined model with variation of: (a) supply voltage, (b) load, (c) inductance, (d) reference current

Fig. 8: System chaotic behaviour

Fig. 9: System response using FLS \( (\delta d) \)

Fig. 10: System response using the proposed controller
TABLE IV
SYSTEM PARAMETERS WHEN USING FUZZY LOGIC CONTROLLER

<table>
<thead>
<tr>
<th>Bif. Par.</th>
<th>( V_g )</th>
<th>( R )</th>
<th>( L )</th>
<th>( I_{\text{ref}} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>([3, 50])</td>
<td>20</td>
<td>27</td>
<td>4</td>
<td></td>
</tr>
<tr>
<td>([8, 150])</td>
<td>27</td>
<td>4</td>
<td></td>
<td></td>
</tr>
<tr>
<td>([1, 200])</td>
<td>4</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>([1.4, 14])</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Figures 11-a, b, c, d, give the bifurcation diagrams corresponding to the operating domains, described in table IV, using the proposed fuzzy logic controller. Table V gives a quantification of the obtained results.

From figure 11-a, we can remark the non-linear phenomena elimination and the P1 region widening (P1: \( V_g \in [6.2, 50] \)) compared to that obtained by the feedback controller (P1: \( V_g \in [35.1, 50] \)).

For \( V_g = 4 \), we have a zero voltage across the capacitor (Fig. 11-a) because the supply voltage is not enough to charge the capacitor and we can also note that the reference current is not reached (Fig 12-a). For a supply voltage drop higher than 6.2, the capacitor is charged and discharged progressively and the reference current is reached with an overshoot and a response time that decreases with increasing supply voltage, as shown in figures 12-a and 12-b.

Fig. 11: Bifurcation diagrams using the fuzzy logic controller with variation of: (a) supply voltage, (b) load, (c) inductance, (d) reference current

Fig. 12: System response for five operating points according to \( V_g \) values
So, we can conclude that in $V_p \in [6.2, 50]$ V, the proposed controller can ensure current regulation and non-linear phenomena suppression for a wide range of supply voltage variation.

| TABLE V |
|------------------|-----------------|-----------------|-----------------|
| SYSTEM BEHAVIOUR WITH FUZZY LOGIC CONTROLLER | **P1** | **P2** | **P4** |
| Refined model | - | - | - |
| Fig. 11-a | [6.2, 50] (Vp) | - | - |
| Fig. 11-b | [8, 94] (R) | - | - |
| Fig. 11-c | [6.2, 50] (L) | - | - |
| Fig. 11-d | [1.4, 10.4] (Iref) | - | [94, 150] | [120, 200] | [10.4, 14] |

Figure 11-b shows the enhancement achieved with the use of a fuzzy logic controller. Indeed, a load variation entails a shift from a P1 region of [8, 12.8]Ω (table II) with a feedback controller to a P1 region of [8, 94]Ω (table V) with the proposed controller. For the feedback controller, the chaotic behaviour takes place for $R \geq 37.5\Omega$ whereas the fuzzy controller only exhibits quasi periodicity$^4$ for $R \geq 94\Omega$.

Figure 11-c gives the bifurcation diagram in the case of inductance variation from 1mH to 200mH. The converter operates in discontinuous conduction mode for inductance values less than 3mH, the P1 behaviour domain is [3, 120]mH, which is larger than the one obtained by the feedback controller given in table II.

The proposed controller ensures, in the case of the reference current variation (Fig. 11-d), a large P1 region of [1.4, 10.4]A whereas this region is limited to [1.4, 3.4]A if we use the feedback controller. The non-linear phenomena are shifted by the fuzzy controller and the observed quasi periodicity behaviour will appear only for $I_{ref} \geq 10.4A$.

VI. CONCLUSION

In this paper, we have presented a refined discrete model for boost converter and analysed the effects of internal resistors of the switching elements. After having explored the non-linear behaviours in this converter, the proposed fuzzy logic controller ensures the converter current regulation in wide range of system parameters variation, shifts the non-linear phenomena outside the operating domains, reduces the number of these phenomena and enlarges the period-one operating region.

REFERENCES


$^4$ Quasi-periodicity: is an oscillation that contains more than one component at incommensurable frequencies (i.e., irrational frequency ratios) [DEA, 90].