Fuzzy Logic Based Sliding Mode Controlled For Active Clamp SEPIC Converter

Azam Salimi¹, Majid Delshad²
¹ Azad University, Khorasgan Branch, azam.salimi1365@yahoo.com
² Azad University, Khorasgan Branch, delshad.majid@gmail.com

Abstract: In this paper a modified design of a sliding mode controller based on fuzzy logic for a soft switching sepic converter is introduced. Here a proportional - integral (PI)-type current mode control is employed and a sliding mode controller is designed utilizing fuzzy algorithm. Sliding mode controller ensures robustness against all variations and fuzzy logic helps to reduce chattering phenomenon introduced by sliding controller. Therefore error, voltage and current ripples decrease. One of the advantages of proposed controller is better performance in comparison with conventional sliding mode controller in soft switching sepic converter. The proposed system is simulated by MATLAB / SIMULINK. The simulation results verify the good performance of proposed controller against variation of input and reference voltages.

Keywords: Switching mode power supplies, SEPIC active clamp converter, sliding mode control, robustness, fuzzy control, current mode control, non-linear behavior.

1. INTRODUCTION

Control of switching power supplies has always been a daunting task for nonlinear and time variant cases. DC-DC converters change state momentarily from one to another and especially the boost one contains non-minimum phase behavior, which makes control mechanism more difficult [1]-[5]. Linear control techniques don't show robustness against sudden load and input voltage variations. It is required for a DC-DC converter to provide regulated output despite all kinds of perturbations [2],[4]. The conventional proportional integral- derivative (PID) controller has been used in many control applications because of its simplicity and effectiveness. The disadvantage of PID controller is its poor capability of dealing with system uncertainty i.e. parameter variations and external disturbance. Since control action changes rapidly from one state to another changing converter topologies, so it makes sense that nearly all designed controllers for switching converters are indeed variable structured ones. Sliding mode controllers (SMC) for such systems prove to be very useful because of inherent robustness, capability of system order reduction and congenial to on-off switching of converters [2],[6]-[7]. Sliding mode being a discontinuous control, state trajectories move back and forth around a certain average surface in the state space and ripples come into being, which is called chattering.

Fuzzy logic systems don't need accurate mathematical models of the controlled system and hence, have been applied to many unknown nonlinear control problems [9]-[13]. Human expert knowledge or the trial-and-error tuning procedure determines the design of fuzzy rules in fuzzy control (FC). Theoretical FC designs based on the sliding mode control scheme known as fuzzy sliding mode control (FSMC) have been proposed to reduce the number of fuzzy rules [14]. Application of FSMC to a DC/DC buck boost converter has been proposed in [15] where the implementation needed two variables i.e. sliding surface and its derivatives. This technique takes huge computational time and may be prone to instability. In this paper, a single variable fuzzy sliding mode control technique is proposed for a DC/DC sepic active clamp converter. The proposed system's robustness is tested against input and reference voltage variations.

2. MODELING ACTIVE-CLAMP SEPIC CONVERTERS

Fig.1 shows an active-clamp SEPIC converter with transformer isolation which is studied in this section to demonstrate the application of the proposed modeling method.
Compared to a hard-switching SEPIC converter, the active-clamp version has an additional switch (S2), a clamping capacitor (Cs) and a resonant inductor (Lr). The resonant inductor is usually created by the transformer leakage inductance. The main switch, (S1) is modeled as an ideal switch and an anti parallel diode and its output capacitance which is absorbed in the resonant capacitor (Cr). The auxiliary switch also has output capacitance, but it is usually much smaller than Cr and can be ignored in the following analysis. The Lm represents the magnetizing inductor of the transformer. The model described in this section has been used to design feedback control for the single-phase PFC converter in [21].

**A. Basic Operation**

Steady-state operation of the active-clamp SEPIC converter is reviewed here to help understand the modeling steps. Fig. 2 shows the key waveforms of the converter in steady state operation. In order to simplify the steady state analysis, the following assumptions are made.

- All parasitic components are neglected except LK
- The clamp capacitor, coupling capacitor and input inductor are large enough, so they have constant value in a switching cycle
- The converter has six modes in each switching cycle [19] and the first mode begins when main switch is turned on.

**Mode 1)** The main switch and the secondary rectifier D are both on. The transformer magnetizing inductor is discharged by the output voltage, and the resonant inductor is charged by the coupling capacitor voltage. This mode ends when the resonant inductor current becomes equal to the magnetizing current.

**Mode 2)** The resonant inductor is in series with the magnetizing inductor, and both are charged by the coupling capacitor voltage. This mode ends when the main switch is turned off.

**Mode 3)** S1, S2 and D are all off in this mode. The output capacitor is charged by the input current and resonant inductor current. This mode ends when the secondary diode D turns on.

The voltage across CS is as following

\[ v_c + v_o(1 + \frac{L_r}{L_m}) \]

**Mode 4)** The voltage across the magnetizing inductor is clamped because of the conduction of D. The resonant inductor resonates with the output capacitor When resonant capacitor completely charges, the body diode of S2 conducts and this mode ends.

**Mode 5)** In this mode both S1 and D are on and the voltage across the magnetizing inductor is clamped by the output voltage. Hence both inductor currents will decrease, with different slopes. This mode ends when S1 is turned off.

**Mode 6)** When S2 is turned off this mode begins and Lr current starts to discharge Cr resonantly and when this capacitor discharge completely the body diode of S1 conducts and therefore after this instant, S1 can be turned on under zero voltage condition (ZVS).

**B. Reduced-Order Averaged Model Derivation**

From the steady-state operation analysis, it can be concluded that the resonant inductor current and the resonant capacitor voltage are fast state variables, while all other state variables (im , vs , vC , vo , and iL) can be considered slow variables. The time-scale separation method presented in [20] can be applied to develop a reduced-order averaged model for the slow state variables which describes low-frequency (up to half the switching frequency) dynamics of the converter. The procedure is to first determine the responses of the resonant inductor current and the capacitor voltage in each of the six
intervals over a switching cycle. The slow variables are assumed to be constant (and ripple free) in this calculation. The calculated responses, which are usually dependent of the time and the slow state variables, are substituted into the state space equations of the slow state variables such that these equations become self sufficient (decoupled), that is, they don’t involve the resonant inductor current and capacitor voltage any more. Standard averaging technique can then be applied to remove the time dependence of the decoupled equations, resulting in a fifth-order averaged models for the five slow state variables (im , vs , vc , vo , and iL). Although mathematically not necessary, we will make some approximations for the waveforms shown in Fig. 2(a) to simplify the model derivation. The approximations mainly involve ignoring the dead time between the conduction of and . As defined in Fig. 2(a), the dead time encompasses intervals[t1–t2], [t2–t3] , and [t3–t4] , and is usually very short. The approximated waveforms without these three intervals are shown in Fig. 2(b).

This leaves three intervals in each switching cycle:[0,αTs] , which coincides with interval[t0,t1] ; [αTs,t2], which corresponds to[t1,t2] ; and[dTs,Ts] , which corresponds to[t3,t4] . Note that the resonant capacitor voltage becomes a rectangular wave when the resonant transition intervals are ignored. (Similar approximations are used in [19] in the derivation of full-order averaged models.) Additionally, we assume that all components shown in Fig. 1 are ideal, and that the converter is lossless. The first step in the development of a reduced-order averaged model is to calculate the periodic responses of the fast variables, that is, the resonant inductor current and capacitor voltage. Since the resonant capacitor voltage assumes a rectangular wave under the assumptions made before, only the responses of the resonant inductor need to be calculated. In the procedure proposed in [18], this is calculated by treating all slow variables as constant and ignoring their switching ripple. Based on the previous steady-state operation analysis, responses of the resonant inductor current are found to bel

\[
\begin{aligned}
i_r(t) &= \begin{cases} 
im + \frac{v_c + n v_o}{L_r}(t - \alpha T_s) & t \in [0, \alpha T_s] \\ \nim & t \in [\alpha T_s, T_s] \\ \nim + \frac{v_c + n v_o - v_s}{L_r}(t - d T_s) & t \in [d T_s, d T_s] \end{cases}
\end{aligned}
\]  \tag{1}

where is n the turns ratio of the transformer. Note that parameter represents the length of the first interval as defined in Fig. 2 and is dependent of other state variables. To eliminate this dependent variable, we note that the resonant inductor current is periodic under the assumption made before, such that

\[
i_r(0) = i_r(T_s) \quad \text{ (2)}
\]

With (1) and (2), a solution can be found for

\[
\alpha = \frac{(-1+d)(n v_o + v_c - v_s)}{n v_o + v_c} \quad \text{ (3)}
\]

In the next section, we will present a different approach to the calculation of by considering the switching ripple of the magnetizing current as a way to improve the model accuracy when large switching ripple is present. The next step is to substitute and in the state equations of the slow variables by the calculated responses of the resonant inductor current and capacitor voltage. Following are the state equations of the five slow state variables after such substitution.

1) Input inductor current

\[
i_L(t) = \begin{cases} v_{in} & t \in [0, \alpha T_s] \\ \frac{v_{in}}{L} & t \in [\alpha T_s, d T_s] \\ v_{in} - v_s & t \in [d T_s, d T_s] \end{cases}
\]

2) Clamping capacitor voltage

\[
v_s(t) = \begin{cases} 0 & t \in [0, \alpha T_s] \\ \frac{1}{C_s} \left[ i_L + \nim + \frac{v_c + n v_o - v_s}{L_r}(t - d T_s) \right] & t \in [d T_s, d T_s] \end{cases}
\]

3) Transformer magnetizing current

\[
im(t) = \begin{cases} -n v_o / L_m & t \in [0, \alpha T_s] \\ \frac{v_c}{(L_m + L_r)} & t \in [\alpha T_s, d T_s] \\ -n v_o / L_m & t \in [d T_s, d T_s] \end{cases}
\]

4) Coupling capacitor voltage
3. DESIGNING OF SLIDING MODE CONTROLLER

The sliding mode controller (SMC) is a variable structure control technique compatible with the nonlinear behavior of a boost converter. Determining a suitable switching or sliding surface is the first step to design a sliding mode controller. Here error is denoted as

\[ e = i_{\text{Ref}} - i_L \]  \hfill (10)

for this converter, the sliding surface \( S \) are defined as

\[ S = Ke = K(i_{\text{Ref}} - i_L) \]  \hfill (11)

where \( K \) is the sliding coefficient and \( i_{\text{Ref}} \) is the desired output current produced by a PI loop in Fig. 5. Now a switching strategy should be designed to make the system reach sliding surface in finite time. After reaching the surface, the system achieves desired system dynamics and becomes globally asymptotic stable [16]. A positive definite Lyapunov function \( P \) may be defined [6] as

\[ P = \frac{1}{2} S^2 \]  \hfill (12)

Ensuring stability for the system in sliding mode requires the derivative of \( P \) be negative definite i.e. \( \dot{P} < 0 \) and hence the following inequality should be fulfilled:

\[ P = SS < 0 \]  \hfill (13)

So, both reaching mode behavior and sliding mode stability are ensured by the following switching law:

\[ d = \begin{cases} 0 & \text{S < 0} \\ 1 & \text{S > 0} \end{cases} \]  \hfill (14)

As in [8] and [18], the sliding mode control law \( d \) is

\[ d = \frac{1}{2}(1 + \text{sign}(S)) \]  \hfill (15)

Design procedure and necessary equations for SMC were presented in [8]. Output curves for this sliding controller were also presented against input and reference voltages. SMC suffers from a major drawback like chattering phenomenon originated from switching at infinite frequency between the two structures [2], [16]. An FSMC scheme is proposed in the next section to check the drawbacks of the sliding mode control and attain more accuracy.

IV. DESIGNING OF FUZZY SLIDING MODE CONTROLLER

Fuzzy logic when combined with sliding mode control contributes significantly to the improvement of performance of nonlinear systems. The fuzzy sliding mode controller

The above model differs from existing reduced-order models in that the effects of the fast state variables are captured in the model. On the other hand, unlike the existing full-order models [22], this new reduced-order model captures the effects of the fast state variables without actually including them in the final model, hence is much simpler and easier to use.
proposed in this paper goes in line with the inequality (13), which implies the multiplication of sliding surface (11) and its derivative be negative definite [6], [15].

\[ d = \frac{1}{2} \left[ 1 + \text{sign}(S) \right] \]  

Here \( S_f \) determines the duration of \( d \) to attain desired output. \( K \) in (9) is determined to be 1. For fuzzy logic, Mamdani fuzzy inference system is used.

Figure 4 and figure 5, show two triangle membership functions which are designed for fuzzy block inputs. Also \( E \) is applied to fuzzy block according to figure 3. The control signal \( S_f \) is produced by using fuzzy rules are given in TABLE 1. Also triangle membership function of output is shown in figures 6.

\[
\begin{align*}
\text{TABLE 1. Fuzzy rules} \\
\hline
\text{e}_2 & \text{e}_1 & \text{NB} & \text{NM} & \text{NS} & \text{ZE} & \text{PS} & \text{PM} & \text{PB} \\
\text{NB} & \text{NB} & \text{NB} & \text{NB} & \text{NM} & \text{NM} & \text{NS} & \text{ZE} \\
\text{NM} & \text{NB} & \text{NB} & \text{NM} & \text{NS} & \text{NS} & \text{ZE} & \text{PS} \\
\text{NS} & \text{NB} & \text{NM} & \text{NS} & \text{NS} & \text{ZE} & \text{PS} & \text{PM} \\
\text{ZE} & \text{NM} & \text{NS} & \text{NS} & \text{ZE} & \text{PS} & \text{PS} & \text{PM} \\
\text{PS} & \text{NM} & \text{NS} & \text{ZE} & \text{PS} & \text{PS} & \text{PM} & \text{PB} \\
\text{PM} & \text{NS} & \text{ZE} & \text{PS} & \text{PS} & \text{PM} & \text{PB} & \text{PB} \\
\text{PB} & \text{ZE} & \text{PS} & \text{PM} & \text{PM} & \text{PB} & \text{PB} & \text{PB} \\
\hline
\end{align*}
\]

**V. SIMULATION RESULTS**

A block diagram of the proposed system is shown in Fig.7.
TABLE 2. PARAMETERS OF THE EXPERIMENTAL SEPIC CONVERTER

<table>
<thead>
<tr>
<th>PARAMETER</th>
<th>VALUE</th>
<th>PARAMETER</th>
<th>VALUE</th>
</tr>
</thead>
<tbody>
<tr>
<td>$v_{in}$</td>
<td>115v</td>
<td>$v_o$</td>
<td>15v</td>
</tr>
<tr>
<td>$P_o$</td>
<td>62.5W</td>
<td>$R$</td>
<td>10Ω</td>
</tr>
<tr>
<td>$L$</td>
<td>1.14mH</td>
<td>$L_m$</td>
<td>247μH</td>
</tr>
<tr>
<td>$L_r$</td>
<td>20μH</td>
<td>$C_s$</td>
<td>0.47μF</td>
</tr>
<tr>
<td>$C_C$</td>
<td>0.47μF</td>
<td>$C_o$</td>
<td>1200μF</td>
</tr>
<tr>
<td>$N$</td>
<td>4</td>
<td>$f_s$</td>
<td>100kHz</td>
</tr>
</tbody>
</table>

Robustness is checked by testing the system response to (1) step change in reference voltage $V_{ref}$ from 15 to 25 V (2) step change in input voltage $V_{in}$ from 115 to 200 volts at 0.3 sec.

In Figures 9 and 10, voltage and inductor current are shown for both cases respectively and these results verify robustness of proposed controller against perturbations.

Figure 9. The output voltage under load and input voltage variations in 0.3S.

Figure 10. The inductor current $i_L$ under load and input voltage variations in 0.3S.

It is necessary to design control system to offset the variance of proportional gain $k_p$. Designing fuzzy membership functions is a trial-and-error process and hence, different techniques can be applied to achieve maximum satisfactory performance.

VI. CONCLUSIONS

In this paper a single variable fuzzy sliding mode control scheme with the sliding surface as input in order to improve the robustness and performance of a boost DCDC soft switching sepic converter is proposed. The simulation results show that the proposed controller operates under variable frequency and the PI loop does not suit all operating conditions. But more accuracy is achieved by modifying fuzzy rule base and membership functions.

REFERENCES


