Sliding-Mode Control of the DC-DC Ćuk Converter in Discontinuous Conduction Mode

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ABSTRACT

In this paper, a novel approach for two-loop control of the DC-DC Ćuk converter in discontinuous conduction mode is presented using a sliding mode controller. The proposed controller can regulate the output of the converter in a wide range of input voltage and load resistance. Controller parameters are selected using PSO algorithms. In order to verify the accuracy and efficiency of the developed sliding mode controller, the proposed method is simulated in MATLAB/Simulink. It is shown that the developed controller has, the faster dynamic response compared with standard integrated circuit (MIC38C42-5) based regulators.

KEYWORDS: sliding mode controller, steady-state error, a double-loop controller, inductor current sampling, Particle Swarm Optimization (PSO) and discontinuous conduction mode

1. INTRODUCTION

Since its introduction, the Ćuk converter has caught the interest of many power supply designers with its well-regarded advantages, which include magnetic components inerrability, smooth input and output currents, wide conversion ratio and full transformer utilization etc. [1]. To be more specific, an important feature of this converter is that through careful adjustment of its integrated magnetic circuits, it is practically possible to achieve a ripple-free input current or output current. This unique feature makes it an excellent candidate for future electronic applications that involve fuel cell or light-emitting diode (LED), both which require ripple-free current to maintain a long lifespan.

Considering the presence of different switches in the circuit topology of the DC-DC power supplies, model of the power electronic converters is nonlinear. For this reason, application of conventional linear controllers or fixed compensators is not a suitable choice in a wide range of operation [2]. Considering the nonlinear behavior of the DC-DC converters, different nonlinear controllers such as Lyapunov based adaptive controller [3], passivity based controller [4], feedback linearization [5] and sliding mode controller [6] are reported. Among these nonlinear methods, sliding mode approach is more popular due to its simplicity and robustness. In addition, this nonlinear controller can be implemented using simple analog circuits. Hence, it is possible to increase switching frequency of the DC-DC converter considerably. For example, a 200 kHz DC-DC buck converter is reported in [7] using sliding mode controller. Obviously, if a DC-DC converter is used in high switching frequencies, it is possible to choose smaller inductor and capacitor which improves dynamic response of the closed-loop system.

The idea of using a sliding mode controller in nonlinear systems was proposed by Utkin [8]. Middlebrook suggested application of this approach in DC-DC converters [9]. The advantages of using a sliding mode method in DC-DC converters are presented in [10]. Also in [11], sliding mode control of the DC-DC buck/boost converter has been reviewed comprehensively. Different problems and their possible solutions associated with practical implementation of the sliding mode controller with DC-DC buck converters are reported in [7]. Although a simple and practical method for implementing of the sliding mode controller is presented in [7], however frequency of the proposed controller is not constant. In this condition, filter design will be completely difficult. To stabilize the switching frequency, a sliding mode controller based on adaptive feed-forward and feedback control schemes is reported in [12]. However, implementation of this method is completely complicated. In addition, the method described in [12] is not always results in constant switching frequency and in the transient moments-related to the load and input voltage changes-switching frequency is not constant.

Using equivalent control based sliding-mode [13], switching frequency of the converter can be kept constant and application of this approach in CCM operation of DC-DC boost converter is reported in [14] and [15]. In these papers, presented results clearly demonstrate that the converter switching frequency is completely constant. Also in [16], application of the equivalent control theory in the output voltage regulation of basic DC-DC converters (buck, boost and buck/boost) has been reported in Discontinuous Conduction Mode (DCM). Moreover, in [17, 18], sliding-mode control of the Ćuk converter in CCM operation is presented.

Application of the sliding mode method in Continuous Conduction Mode (CCM) in basic DC-DC converters have been studied completely. On the other hand, closed-loop nonlinear control of the DC-DC converters in DCM operation are more complicated. In fact, the output voltage of these converters can't be regulated properly using a fixed compensating network, [16]. Hence, most of the presented papers on DC-DC converters control are limited to CCM operation.

In addition to difficulties related to nonlinear controller design in DCM, there is another problem in designing a proper controller for DC-DC converters: load resistance and input voltage value may be changed widely. In [16], these variations are resulted in considerable steady-state error. Also, from controller design viewpoint, DC-DC Ćuk converter is non-minimum phase and direct control of the output

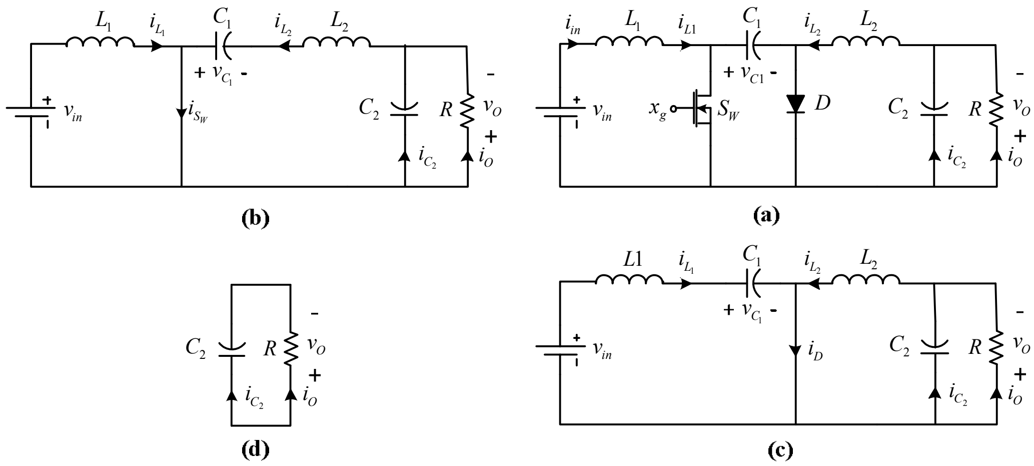

Fig.1. a) Power circuit of the DC-DC Ćuk converter, equivalent circuit of the DC-DC Ćuk converter in b) $0 \leq t < DT$, c) $DT \leq t < (D + D')T$ d) $(D + D')T \leq t < T$

voltage may result in instability of the closed-loop system [19].

On the other hand, the DCM operation of the DC-DC Ćuk converter is completely advantageous. For example, diode current at turn off instant in DCM operation is zero, which results in lower switching loss compared to CCM operation. In addition, due to the small value of the transformer magnetic inductance, dynamic response of the converter is faster in DCM. In addition, the average value of the inductor current in DCM operation is lower which reduces the possibility of the core magnetic saturation.

According to our little search and try, recently no further reported paper were found which apply the sliding mode controller for DC-DC Ćuk converter in DCM operation.

In this paper, a novel approach for output voltage control of the DC-DC Ćuk converter in DCM operation is presented. Due to wide changes of load resistance, input voltage and output voltage reference, and also considering the non-minimum phase nature of this converter, a double-loop control method [20] is used. In this method, the simultaneous applications of output voltage and inductor current feedbacks improve the dynamic response of the closed-loop system considerably. The developed sliding-mode controller is designed based on equivalent control method and hence, switching frequency of the converter is constant. In addition, an integral term is considered in the generation of converter reference current which eliminates steady-state error of the output voltage under different conditions. Also, Controller parameters are selected using PSO algorithms. Finally, in order to investigate the accuracy and effectiveness of the proposed controller compared with standard MIC38C42-5 based regulators [19]; some simulation results are obtained using MATLAB/Simulink software.

2. PRINCIPLES OF DC-DC ĆUK CONVERTER DCM OPERATION

Circuit structure of DC to DC Ćuk converter in DCM at various times is shown in Figure 1. Considering that, i_{L_1}And i_{L_2} are getting (being) simultaneously zero, so the converter turned into DCM when the i_{L_1} is being zero. When the i_{L_1} is discontinuous

(Figure 2), three different operation area can be considered by switch (S_W) work condition and inductor current. For modeling the converter, switching function (u) which indicates the operating status of the switch (can be 0 or 1), is defined by:

$$\mathrm{u} = \begin{cases} 1 & (\text{When} \quad \text{S is ON}) \\ 0 & (\text{When} \quad \text{S is OFF}) \end{cases} \tag{1}$$

Within the zero range of u ($DT_S \leq t \leq T_S$), the switching function (u_B) for determining the zero inductor current can be also defined as follows:

$$\mathrm{u_B} = \begin{cases} 1 & \text{When} \quad \mathrm{i_{Lm}} > 0 \quad \textit{and} \quad u = 0 \\ 0 & \text{When} \quad \mathrm{i_{Lm}} = 0 \quad \text{and} \quad \mathrm{u} = 0 \end{cases} \tag{2}$$

It is clear that in CCM, $i_{L_1} > 0$ and $u_B = 1$.

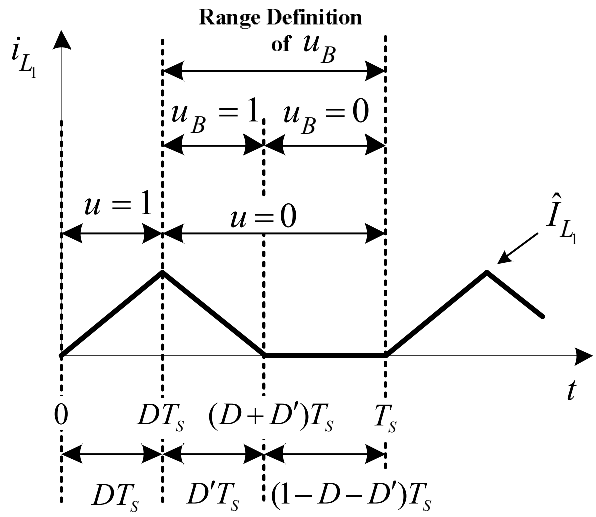

Fig.2. Inductor current of the DC-DC converter in DCM operations

In steady state, considering the voltage changes in L1 and L2 inductors (figures (3 - a) and (3 - b)) and also because the integral of them in a period is zero, an equation between input and output voltage in DCM, can be simply written as follows:

$$\mathrm{L_1}: \quad \mathrm{V_{in} DT_S} + (\mathrm{V_{in}} - \mathrm{V_{C_1}})\mathrm{D'T_S} = 0 \ \Rightarrow \ \mathrm{V_{C_1}} = \frac{\mathrm{D+D'}}{\mathrm{D'}} \mathrm{V_{in}} \tag{3}$$

$$\mathrm{L_2}: \quad (\mathrm{V_{C_1}} - \mathrm{V_O})\mathrm{DT_S} + (-\mathrm{V_O})\mathrm{D'T_S} = 0 \ \Rightarrow \ \mathrm{V_{C_1}} = \frac{\mathrm{D+D'}}{\mathrm{D}} \mathrm{V_O} \tag{4}$$

Equaling the above equations:

$$\mathrm{V_O} = \frac{\mathrm{D}}{\mathrm{D'}} \mathrm{V_{in}} \tag{5}$$

3. AVERAGED STATE SPACE MODEL OF DC TO DC ĆUK CONVERTER IN DCM

The converter state variables, by considering the number of energy storage elements, can be simply defined as follows:

$$\mathbf{X}^{\mathrm{T}} = (\mathrm{x_1}, \mathrm{x_2}, \mathrm{x_3}, \mathrm{x_4}) = (\mathrm{i_{L_1}}, \mathrm{i_{L_2}}, \mathrm{v_{C_1}}, \mathrm{v_o}) \tag{6}$$

When the switch is on, the following equation can be written by considering figure (1-b) $(0 \leq t < DT_S)$

$$\dot{\mathbf{X}} = \mathbf{A}_1\mathbf{X} + \mathbf{B}_1\mathrm{U} = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 0 & 0 & \frac{1}{\mathrm{L_2}} & -\frac{1}{\mathrm{L_2}} \\ 0 & -\frac{1}{\mathrm{C_1}} & 0 & 0 \\ 0 & \frac{1}{\mathrm{C_2}} & 0 & -\frac{1}{\mathrm{RC_2}} \end{bmatrix} \begin{bmatrix} \mathrm{i_{L_1}} \\ \mathrm{i_{L_2}} \\ \mathrm{v_{C_1}} \\ \mathrm{v_o} \end{bmatrix}$$

$$+ \begin{bmatrix} \frac{1}{\mathrm{L_1}} \\ 0 \\ 0 \\ 0 \end{bmatrix} \mathrm{v_{in}} \quad 0 \leq \mathrm{t} \leq \mathrm{DT_S} \tag{7}$$

When the switch is off and $i_{L_1} > 0$, the following state equation can be written by considering figure (1-c) $(DT_S \leq t <$ $D{+}D'TS$

$$\dot{\mathbf{X}} = \mathbf{A}_2\mathbf{X} + \mathbf{B}_2\mathrm{U} = \begin{bmatrix} 0 & 0 & -\frac{1}{\mathrm{L_1}} & 0 \\ 0 & 0 & 0 & -\frac{1}{\mathrm{L_2}} \\ \frac{1}{\mathrm{C_1}} & 0 & 0 & 0 \\ 0 & \frac{1}{\mathrm{C_2}} & 0 & -\frac{1}{\mathrm{RC_2}} \end{bmatrix} \begin{bmatrix} \mathrm{i_{L_1}} \\ \mathrm{i_{L_2}} \\ \mathrm{v_{C_1}} \\ \mathrm{v_o} \end{bmatrix}$$

$$+ \begin{bmatrix} \frac{1}{\mathrm{L_1}} \\ 0 \\ 0 \\ 0 \end{bmatrix} \mathrm{v_{in}} \qquad \mathrm{DT_S} \leq \mathrm{t} < (\mathrm{D} + \mathrm{D'})\mathrm{T_S} \tag{8}$$

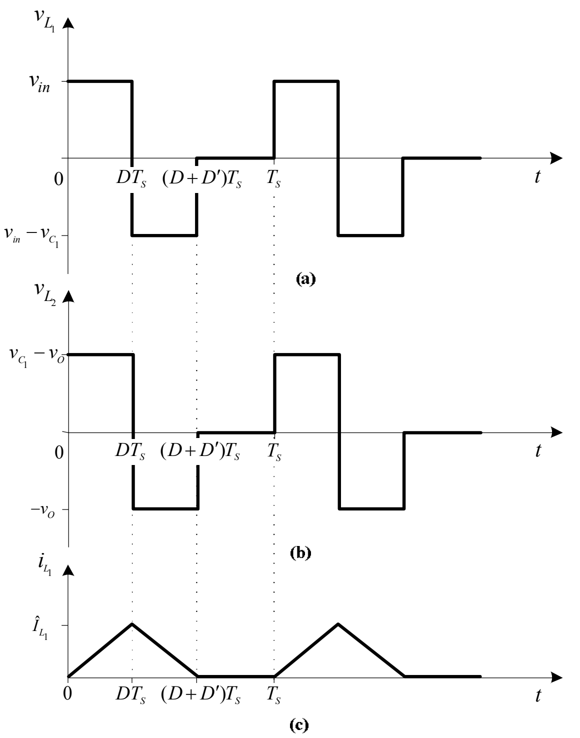

Fig.3. a) L_1 voltage b) L_2 voltage c) i_{L_1} in wave DCM

And finally when the switch S_W is off and $i_{L_1}=0$, the following state equation can be written by considering figure (1-d) $\left((D+D')T_S \le t < T_S\right)$

$$\dot{\mathbf{X}} = \mathbf{A}_3\mathbf{X} + \mathbf{B}_3\mathrm{U} = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & -\frac{1}{\mathrm{RC_2}} \end{bmatrix} \begin{bmatrix} \mathrm{i_{L_1}} \\ \mathrm{i_{L_2}} \\ \mathrm{v_{C_1}} \\ \mathrm{v_o} \end{bmatrix} + \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} \mathrm{v_{in}} \qquad \left(\mathrm{D}+\mathrm{D}'\right)\mathrm{T_S} \le \mathrm{t} < \mathrm{T_S} \tag{9}$$

Due to the averaging method, the general model of Ćuk converter can be derived as follows:

$$\dot{\mathbf{X}} = \mathbf{A}_{\mathrm{ave}}\mathbf{X} + \mathbf{B}_{\mathrm{av}}\mathrm{U} = \begin{bmatrix} 0 & 0 & -\frac{\mathrm{u_B}}{\mathrm{L_1}} & 0 \\ 0 & 0 & \frac{\mathrm{u}}{\mathrm{L_2}} & -\frac{\mathrm{u}+\mathrm{u_B}}{\mathrm{L_2}} \\ \frac{\mathrm{u_B}}{\mathrm{C_1}} & -\frac{\mathrm{u}}{\mathrm{C_1}} & 0 & 0 \\ 0 & \frac{\mathrm{u}+\mathrm{u_B}}{\mathrm{C_2}} & 0 & -\frac{1}{\mathrm{RC_2}} \end{bmatrix} \begin{bmatrix} \mathrm{i_{L_1}} \\ \mathrm{i_{L_2}} \\ \mathrm{v_{C_1}} \\ \mathrm{v_o} \end{bmatrix} + \begin{bmatrix} \frac{\mathrm{u}+\mathrm{u_B}}{\mathrm{L_1}} \\ 0 \\ 0 \\ 0 \end{bmatrix} \mathrm{v_{in}} \qquad \left(\mathrm{D}+\mathrm{D}'\right)\mathrm{T_S} \le \mathrm{t} < \mathrm{T_S} \tag{10}$$

4. FIXED FREQUENCY SLIDING MODE CONTROL OF THE DC-DC FLYBACK CONVERTER IN DCM OPERATION

In this section, sliding mode control of the converter in DCM operation is presented in detail. First sliding mode control theory is reviewed.

4.1. Theory of the applied sliding mode controller

Suppose that, the nonlinear system model on R^n is assumed as below. Suppose the origin of the coordinate as an operating point of the system in steady-state.

$$\dot{\mathbf{z}} = \mathbf{Az} + \mathbf{uBz} \tag{11}$$

where A and B (square $n \times n$) matrices are fixed with real components. The scalar control function u takes values 0 and 1. In the sliding mode control method, u is considered as [13]:

$$\mathrm{u} = \frac{1}{2}\left(1 + \mathrm{sgnS(z)}\right) \tag{12}$$

where in this equation, sgn is the sign function symbol and $S(z)$ is called sliding surface. Equation (12) states that:

$$\text{if } \mathrm{S(z)} > 0 \;\Rightarrow \mathrm{u} = 1 \;\text{ and if }\; \mathrm{S(z)} < 0 \;\Rightarrow \mathrm{u} = 0 \tag{13}$$

Necessary and sufficient conditions for the existence of sliding motion on the sliding surface can be written as [13]:

$$\lim \mathrm{S}.\dot{\mathrm{S}} < 0 \tag{14}$$

The smooth control function for nonlinear system which its model is written in (11) and adopts sliding surface as a local integral manifold, is known as equivalent control and is shown by u_{eq}. Equivalent control can be calculated by equating derivative of the sliding surface to zero [13]:

$$\frac{dS}{dt} = 0 \Rightarrow \left[\frac{\partial S}{\partial z}\right]^T \dot{z} = 0 \tag{15}$$

Considering that the value of $\dot{z}$ is given in (12), the controller can be obtained according to (15):

$$u_{eq} = -\frac{\left[\frac{\partial S}{\partial z}\right]^T Az}{\left[\frac{\partial S}{\partial z}\right]^T Bz} \tag{16}$$

The sliding motion exists locally on the sliding surface, if and only if, u_{eq} satisfies the following condition [13]:

$$0 < u_{eq} < 1 \tag{17}$$

4.2. Calculation of u_B

It is clear that according to equation 10, it would be difficult to implement the controller directly because of the presence of u_B in final control equation. That's why we try to eliminate the u_B parameter from control equation (law) by considering the Cúk converter equations in DCM .

According to Figure 2, the average value of the L_1 current can be obtained from the following equation:

$$I_{L_1} = \frac{1}{T_S}\int_0^{T_S} i_{L_1}(t)dt = \frac{D+D'}{2}\hat{I}_{L_1} \tag{18}$$

Also the amount of $\hat{I}_{L_1}$ can be simply calculated as follows:

$$i_{L_1} = \frac{1}{L_1}\int v_{L_1}dt \quad \rightarrow \quad \hat{I}_{L_1} - 0 = \frac{1}{L_1}\int_0^{DT_S} V_{in}dt \quad \rightarrow \quad \hat{I}_{L_1} = \frac{1}{L_1}V_{in}DT_S \tag{19}$$

Substituting the amount of $\hat{I}_{L_1}$ in equation (18):

$$I_{L_1} = \frac{V_{in}}{2L_1}DT_S(D + D') \tag{20}$$

Ignoring the switching losses in the converter:

$$P_{out} = P_{in} \quad \rightarrow \quad V_o I_o = V_{in} I_{in} \tag{21}$$

Where in the above equation, $I_{L_1} = I_{in}$, and:

V_o I_o = V_in I_(L_1) → V_o I_o = V_in (V_in/(2L_1) DT_S (D + D^′)) (22)

Replacing the amount of D' from equation (6):

$$\frac{{V_o}^2}{R} = \frac{{V_{in}}^2}{2L_1}DT_S\left(D + \frac{V_{in}}{V_o}D\right) \tag{23}$$

or

$$D = D_{DCM} = \sqrt{\frac{2L_1}{RT_S}}\frac{V_o}{V_{in}}\sqrt{\frac{V_{in}}{V_{in}+V_o}} \tag{24}$$

and

$$D' = \sqrt{\frac{2L_1}{RT_S}}\sqrt{\frac{V_{in}}{V_{in}+V_o}} \tag{25}$$

Also, the duty cycle of converters in CCM is calculated as follow:

$$D_{CCM} = \frac{V_o}{V_{in}+V_o} \tag{26}$$

It is obvious that the maximum distance in which u_B is equal to one is related to the case which the converter works in CCM. So:

$$D'_{max}T_S = (1 - D_{CCM})T_S \tag{27}$$

And we know that in DCM, $u_B = 1$ is true for $D'T_S$. That's why u_{Beq} can be defined as follows. It should be noted that $0 < u_{Beq} < 1$ and is called as the equivalent virtual switching function.

$$u_{Beq} = \frac{D'T_S}{D'_{max}T_S} \tag{28}$$

u_{Beq} can be calculated as follows using equations (26) and (28):

$$u_{Beq} = \sqrt{\frac{2L_1}{RT_S}}\sqrt{1 + \frac{V_o}{V_{in}}} \tag{29}$$

4.3. Sliding surface selection

The switching manifold for [8] the sliding mode current control is designed as Fig. 4.

$$S = x_{1(ref)} - x_1 = i_{L_1(ref)} - i_{L_1} \quad (30)$$

In recent relationship, $i_{L_{1(ref)}}$ will be as follows:

$$i_{L_1(ref)} = K_P [v_{O(ref)} - v_O] + K_I \int [v_{O(ref)} - v_O] \, dt \quad (31)$$

4.4. Designing of equivalent controller

Considering the performance of Cúk converter in DCM, the dynamic model of the system by regarding the basic rules of the electrical circuit and the equations (30) and (31) can be derived and u_{eq} will be calculated as follows by placing $\dot{S} = 0$ in equation 16:

$$u_{eq} = \frac{1}{\left[K_{i_{L_2}} i_{L_2} + K_{v_{in}} v_{in}\right]} \left\{K_{v_o} v_o + K_I\left(v_O - v_{O(ref)}\right) + K i L2' i L2 + K v C1' v C1 + K vin' vin1 + VoVin \quad (32)$$

where

$$\begin{cases} K_{i_{L_2}} = -\frac{K_P}{C_2} \\ K_{v_{in}} = \frac{1}{L_1} \\ K_{v_o} = -\frac{K_P}{RC_2} \\ K'_{i_{L_2}} = \frac{K_P}{C_2}\sqrt{\frac{2L_1}{RT_S}} \\ K'_{v_{C_1}} = \sqrt{\frac{2}{L_1 RT_S}} \\ K'_{v_{in}} = -\sqrt{\frac{2}{L_1 RT_S}} \end{cases} \quad (33)$$

5. PARTICLE SWARM OPTIMIZATION

Particle swarm optimization is a group algorithm in which the group of particles search in possible space in order to find the optimal solution of an objective function. Each particle with adjustable speed move in the search space and hold the best position ever earned in its memory. The best position achieved by the entire members of the group is transmitted to the others. In PSO, the members doesn't change to the new one, but their behavior, including their movement and speed, is directed for the best answer, and modified through subsequent iterations. The first value is the best answer so far has been found by any of the members, this value is called $pbest$. It is assumed that the search space consists of n dimensions, then the $i-th$ particle can be defined with two n-dimensional vector, positions (x_i) and speed (v_i). The equations (34) and (35) show these two vectors [21-23]:

$$X_i = [x_{i1} \quad x_{i2} \quad x_{i3} \quad \cdots \quad x_{in}]^T \quad (34)$$

$$V_i = [v_{i1} \quad v_{i2} \quad v_{i3} \quad \cdots \quad v_{in}]^T \quad (35)$$

Where i = 1,2,3, ..., N and N is the number of members and Superscript T is the transpose operator. In this algorithm, the i-th particle saves (stores) the best position ever achieved under the vector of $P_i = [p_{i1} \quad p_{i2} \quad p_{i3} \quad \cdots \quad p_{in}]^T$ in its memory. The vector $G = [g_1 \quad g_2 \quad g_3 \quad \cdots \quad g_n]^T$ is the best position ever attained by the whole group. The position of i-th particle in (t+1)$_{th}$ iteration can be defined by equations (36) and (37).

$$v_i(t+1) = \omega(t) \times V_i(t) + C_1(t) \times r_1 \times (P(t)_i - X_i(t)) + C_2(t) \times r_2 \times (G(t) - X_i(t)) \quad (36)$$

$$X_i(t+1) = X_i(t) + \chi \times V_i(t+1) \quad (37)$$

In the above equations, ω is the inertia coefficient which indicates the impact of the previous velocity vector in the current iteration. χ is the contraction factor which is used to limit the effect of the velocity vector. C1 and C2, respectively, are cognitive parameters and social. r_1 and r_2 are two real numbers which will be selected randomly and according to a uniform distribution function between zero and one.

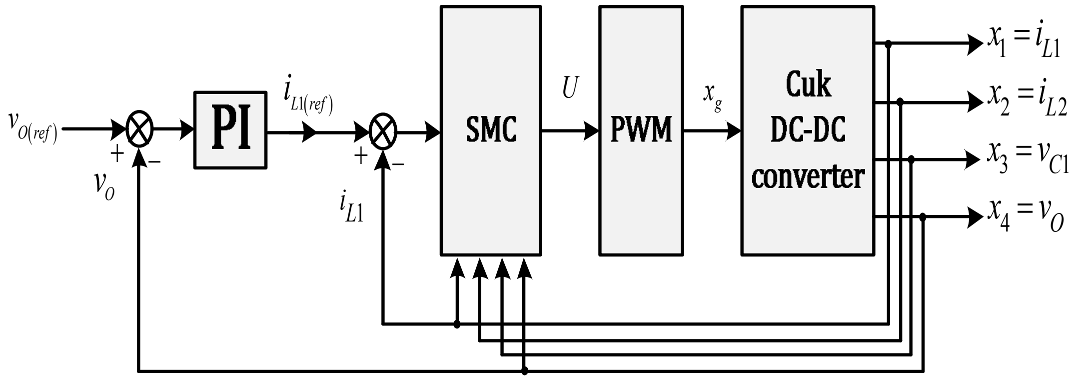

Fig.4. Indirect output voltage control in fly back converter based on the inductor current regulation

If the $c_1 \times r_1$ is greater, the i-th particle will move more rapidly towards the best position ever earned.

Particle velocity in the best position direction which has obtained by the whole group is also influenced by $c_2 \times r_2$. If the inertia coefficient is larger, the group will search in a larger range of space. While the smaller inertia coefficient, will increase the group accuracy in local searches. Based on the experiences gained, it is suggested that, at the beginning of searching process large amounts of ω is dedicated to prioritize the global search over the local search, then in order to achieve the best possible answer, its value are decreasing gradually to a small value like zero.

6. SIMULATION

In this section, DC-DC Ćuk converter has been simulated in DCM operation based on the developed controller (equation (32)) using MATLAB/Simulink. In this case, in order to implement the sliding mode controller, output voltage and power switch current are sampled. Due to switching, measured values may have a significant ripple. Large voltage and current ripple deteriorate response of the closed-loop system. Therefor application of the low-pass filters and calculation of the state variables average values in power electronic converter controller design are completely accepted. Converter and controller parameters are listed in TABLE I. The maximum step size for all simulations is taken $100ns$.

In order to evaluate the overall response of the proposed controller to the variation of load and input voltage, different tests are considered in detail. Also response of the developed sliding mode controller is compared with standard IC based regulators.

Table 1.nominal specifications of the DC-DC Ĉuk converter

$\boldsymbol{Parameter}$	$\boldsymbol{Value}$
$Input\ voltage\ (v_{in})$:	$25\ V$
$Converter\ inductor\ (L_1)$:	$1.9\ mH$
$Converter\ inductor\ (L_2)$:	$0.9\ mH$
$Converter\ capacitor\ (C_1)$:	$850\ \mu F$
$Output\ capacitor\ (C_2)$:	$47\ \mu F$
$Load\ resistance\ (R)$:	$100\ \Omega$
$Reference\ output\ voltage$:	$20\ V$
$Switching\ frequency\ (f_s)$:	$20\ kHz$

Table 2. Obtained values by PSO

Parameter	*Value*
K_P	0.2425
K_I	90.3350
Maximum Overshoot	0.0113
Settling Time	0.1611
Steady State Error	0.0002
Rise Time	0.0869

6.1. Optimization

K_P and K_I parameters setting is led to change in the closed-loop system performance and thus creates different responses. In this section, PSO is used to optimize the response of proposed controller by designing the controller parameters. The fitness function is considered according to equation (38) with the aim of minimizing the sitting time(t_s), rise time (t_r), overshoot (M_p) and steady-state error (E_{ss}).

$$\text{Fitness function} = (1 - e^{-1}) \times (M_p + E_{ss}) + (e^{-1}) \times (t_s - t_r) \qquad (38)$$

K_P and K_I parameters, sitting time(t_s), rise time (t_r), overshoot (M_p) and steady-state error (E_{ss}) are shown in table 2. PSO algorithm parameters and the nominal values of DC-DC Ćuk converter are also shown in table 1.

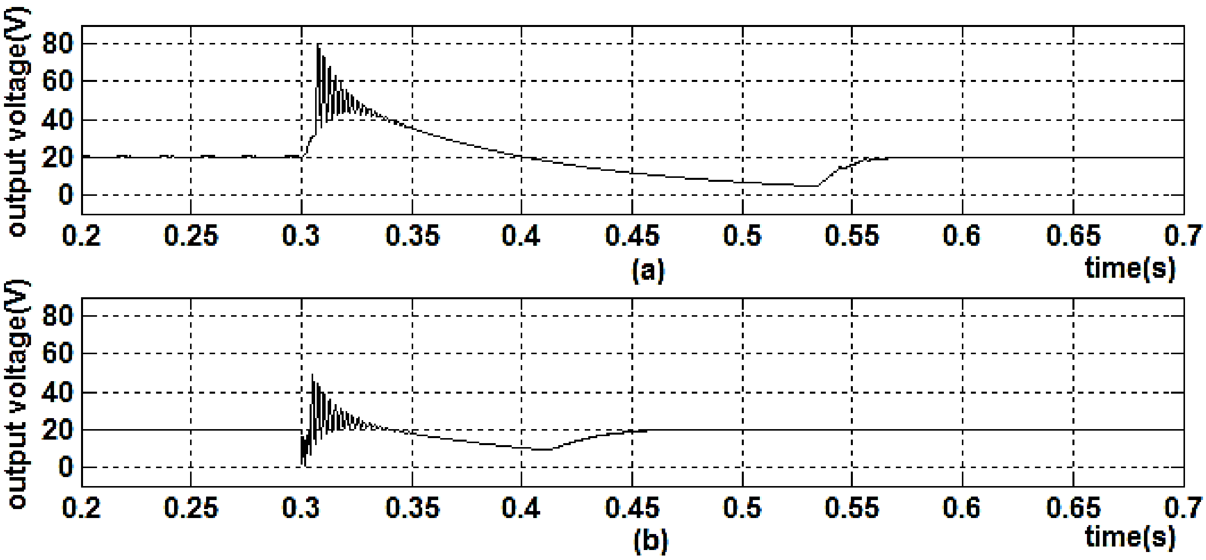

Fig.5. Response of the standard IC based (a) and proposed sliding mode controllers (b). In both plots, input voltage of the converter is stepped from 10V to 30V. It is clear that proposed has better dynamic response.

6.2. Response of the Proposed Controller toInput Voltage Changes.

Usually uncontrolled diode rectifiers are used to implement input voltage source of the DC-DC converters. For this reason, the controller response to input voltage variations is important. The response of the proposed controller to step changes in input voltage is illustrated in sliding mode Fig.6-b. At $t = 0.3s$, the converter input voltage is increased from 10V to 30V. This plot clearly shows the stability of the proposed controller. Also, response of the standard IC (MIC38C42) based regulators is plotted in Fig.6-a. It is obvious that the developed fixed frequency sliding mode controller has better dynamic response compared with standard regulators.

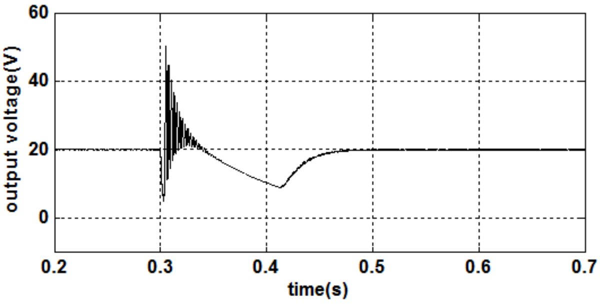

Fig.6. Response of the developed sliding mode controller to simultaneous changes of input voltage, load resistance and output capacitor

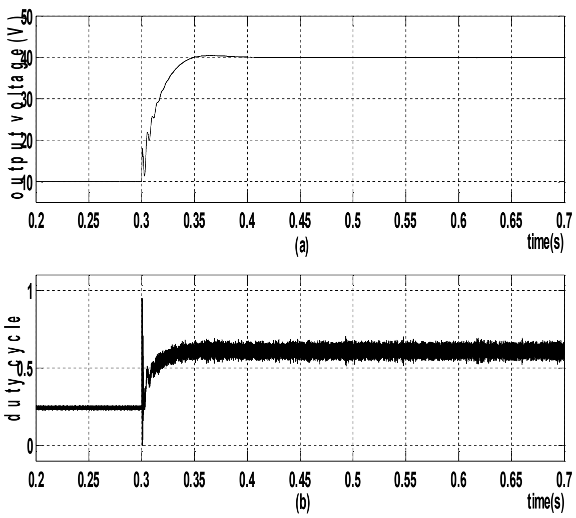

Fig.7. Response of the proposed controller to reference voltage changes

6.3. Simultaneous changes in input voltage, load resistance and output capacitor

In order to evaluate the overall performance of the developed sliding mode controller, input voltage, load resistance and output capacitor of the system are changed simultaneously and in this case, response of the controller is shown in Fig.7. At $t = 0.3s$, input voltage is stepped from 10V to 30V, load resistance is changed from 200Ω to 100Ω and finally output capacitor is stepped from $94\mu F$ to $47\mu F$ simultaneously. Obviously, the converter operated in DCM and it is completely stable.

6.4. Response of the Controller to Reference Voltage Changes

In Fig.8, the response of the designed sliding mode controller to reference voltage changes is illustrated. Considering the parameters given in Table I, the reference voltage is stepped from 10V to 40V at $t = 0.3s$. It is clear that, in spite of large changes in the reference, the controller is able to follow the desired voltages and corresponding steady-state error is zero.

CONCLUSION

In this paper, a sliding mode controller is proposed to regulate the output voltage of the DC-DC fly back converter in DCM operation with constant switching frequency. Considering non-minimum phase nature of the converter, indirect control of the output voltage is used based on two-loop control method. Selection of the sliding surface and calculation of the reference current is developed so that, integral of the output voltage error is present in the final control law which results in zero steady state error. In addition, a novel and simple method is proposed for inductor current average value measurement. Designed controller is simulated based on MATLAB/ Simulink software. Simulation results clearly show that the proposed method has faster dynamic response compared with standard controller MIC38C42-5. In spite of large changes in input voltage, load resistance and reference voltage, the proposed sliding mode controller is completely stable, also; the steady state error of the closed loop system is zero.

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