Discrete-Time Sliding Mode Controlled Synchronous DC-DC Buck Converter

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***Abstract*—This paper proposed a new design of digital sliding mode control for synchronous DC–DC buck converter. In presented control law, disturbance item is considered in the dynamic equation of the synchronous buck converter to achieve fast transient response and high robustness under load and line variations. Also, stability analysis of proposed digital sliding mode control is presented. In fact, we design a control law that closed-loop system has strong robustness to uncertainties as it reduced chattering, steady state error, overshoot and undershoot of output voltage. The simulation results is obtained using the MATLAB/SIMULINK software. Simulation results show that the proposed control law exhibits considerable reduction in terms of the chattering, and overshoot and undershoot of output voltage response during immediate load and line voltage variations. The proposed method reduces the chattering to about 1.1%. Also, the overshoot and undershoot values are 1.3% and 1.8%, respectively. Also, when the input voltage is changed to 26% of its value, there is almost no overshoot and undershoot. The selected synchronous buck converter in this paper has an input voltage of 38V and an output voltage of 5V and a 10W output power.**

***Index Terms*—DC–DC buck converter, sliding mode control, digital control, variable structure system.**

I. INTRODUCTION

Modern electronic systems require power supplies with features such as high reliability, high efficiency, small size. Because of these features as well as flexible design, DC/DC switching-power converters are widely used in industrial systems. DC/DC power converters have many industrial applications, including power supplies for personal computers, photovoltaic (PV) systems [1], electric vehicles [2], robot systems [3]. Switched Mode Power suppliers are variable-structure nonlinear systems. Therefore, a high-performance nonlinear controller is required for these systems to have the ability of disturbance eliminating, and to obtain a small steady state error, fast dynamic response, low overshoot and low undershoot [4].

Linear control methods fail in high signal performance conditions and these methods cannot provide regulation over a wide range of performance. To overcome these problems, so far, various nonlinear control methods, such as passivity control [9], back stepping control [5], linear matrix inequality (LMI) control [6], adaptive control [7], robust control [8], sliding-mode control (SMC) [9 to 11], predictive control [12] have been proposed for DC-DC power converters.

Among the nonlinear control methods offered to control DC-DC converters that require high level computing and logic (this results in the complexity of the control architecture), the sliding mode control (SMC) method with simple circuit architecture can be used and is a suitable choice [13]. SMC has advantages such as guaranteed stability, robustness against parameter variations, fast dynamic response and simplicity in implementation [14].

However, in practice, SMC implementation in power electronic systems is impossible, since it ideally requires infinite switching frequency to achieve steady-state performance and desired dynamic response. On the other hand, the infinite switching frequency causes excessive switching losses and emission of electromagnetic interference (EMI) noise [13]. Limiting the infinite switching frequency in a practical range is possible by applying boundary layers around the sliding surface.

Due to the significant development of digital devices over the past decade, the desire to use digital controllers in DC-DC switching power converters has been increased. Some of the advantages of digital devices include a high noise immunity, absence of aging effects of the device, ability to implement complex control algorithms and flexibility in changing controller parameters, and ability to communicate with other digital systems [15]. Therefore, digital controllers can be a suitable and practical alternative to control switching converters. The implementation of SMC in digital platforms, itself allows the fixed frequency switching operation [13]. Due to the existence of a QSM band (due to sampling interval) around the switching surface, such controllers are inherently working like a boundary layered SMC and can be used to reduce chattering [13]. In fact, replacing the analog blocks with their digital counterpart makes switching converters more robust to component variations[16].

In this paper, the synchronous buck converter has been investigated in which freewheeling diode is replaced with a MOSFET. With this substitution, the conduction losses is reduced and the bidirectional current flow is allowed [17]. The practical application of the buck converter is in low and average power systems and is applicable in cases such as portable devices, photovoltaic systems, and DC distribution systems [18].

Fig. 1 shows a basic structure of a digitally controlled synchronous buck converter. The digital pulse width modulator (DPWM) block generates a duty cycle control command. When DPWM is used in a circuit, a prerequisite is that the resolution of the DPWM must be better than the ADC feedback [16]. If this condition is not met, then the phenomenon known as the limit cycle oscillation occurs [19]. The DPWM also causes delay and phase lag, which is not acceptable in the feedback control system [16]. SMC can be used to resolve the need for DPWM.

This paper is organized as follows. The design process of proposed DSMC for controlling the synchronous buck converter is fully described in Section 2. In Section 3, the simulation results are given to confirm proposed method for a 5-V, 2-A synchronous buck converter. Finally, the conclusions are addressed in Section 4.

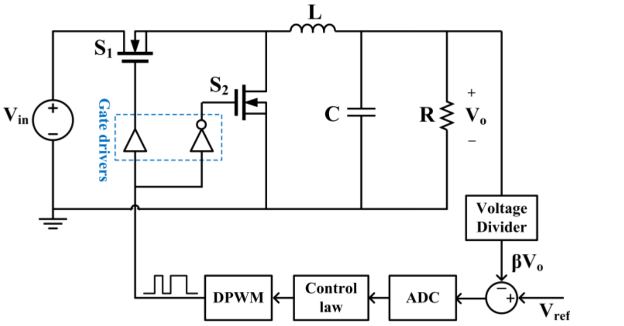


Fig. 1. Schematic of a digital controller for a synchronous buck converter.

II. PROPOSED DIGITAL SLIDING MODE CONTROL FOR SYNCHRONOUS BUCK CONVERTER

A. Dynamical model of the synchronous buck converter

In Fig. 1, a synchronous dc-dc buck converter was shown, in which variables are defined as follows:
V_{in}: DC input voltage (line voltage); S_1 and S_2: controlled switches; L: inductor; C: capacitor; R: load resistance; i_L: inductor current; V_o: output voltage.

When S_1 is on and S_2 is off, equations describing the operation of the converter are as follows:

$$\frac{di_L}{dt} = \frac{1}{L}V_{in} - \frac{1}{L}V_o \tag{1}$$
$$\frac{dV_o}{dt} = \frac{1}{C}i_L - \frac{1}{RC}V_o \tag{2}$$

Now if S_1 is off and S_2 is on, the equations will be as follows:

$$\frac{di_L}{dt} = -\frac{1}{L}V_o \tag{3}$$
$$\frac{dV_o}{dt} = \frac{1}{C}i_L - \frac{1}{RC}V_o \tag{4}$$

We consider u as the control input and is 1 and 0 for ON and OFF modes, respectively. The above relations can be summarized in the following two relations (in fact, u is the discontinuous control input):

$$\frac{di_L}{dt} = \frac{1}{L}uV_{in} - \frac{1}{L}V_o \tag{5}$$
$$\frac{dV_o}{dt} = \frac{1}{C}i_L - \frac{1}{RC}V_o \tag{6}$$

B. Continuous- time definition of variables

The sliding mode controller provides an averaged control law that maintains the state trajectory on the sliding surface [13]. State variables can be defined as follows:

$$x_1 = e \quad , \quad x_2 = \frac{dx_1}{dt} = \dot{e} \tag{7}$$

Where e is the error voltage equal to $e = V_{ref} - V_o$. Therefore, state variables can be written as:

$$x_1 = V_{ref} - V_o \quad , \quad x_2 = -\frac{dV_o}{dt} \tag{8}$$

According to (6):

$$x_2 = -\frac{1}{C}i_L + \frac{1}{RC}V_o \tag{9}$$

Dynamic equations of the system can be written as follows:

$$\frac{dx_1}{dt} = -\frac{dV_o}{dt} = x_2 \tag{10}$$
$$\frac{dx_2}{dt} = -\frac{1}{C}\frac{di_L}{dt} + \frac{1}{RC}\frac{dV_o}{dt} \tag{11}$$

By using (5) and (10) in (11), the system dynamic equations are obtained as follows:

$$\frac{dx_1}{dt} = x_2$$
$$\frac{dx_2}{dt} = -\frac{1}{LC}x_1 - \frac{1}{RC}x_2 - \frac{1}{LC}V_{in}u + \frac{1}{LC}V_{ref} \tag{12}$$

The dynamic (motion) equation of the converter in the state-space model can be expressed in general form as:

$$\frac{d\boldsymbol{X}}{dt} = \boldsymbol{A_c X} + \boldsymbol{B_c}u + \boldsymbol{D_c} \tag{13}$$

according to (12), the following equation is obtained:

$$\begin{bmatrix}\dot{x}_1\\ \dot{x}_2\end{bmatrix} = \begin{bmatrix}0 & 1\\ -\frac{1}{LC} & -\frac{1}{RC}\end{bmatrix}\begin{bmatrix}x_1\\ x_2\end{bmatrix} + \begin{bmatrix}0\\ -\frac{1}{LC}V_{in}\end{bmatrix}u + \begin{bmatrix}0\\ \frac{1}{LC}V_{ref}\end{bmatrix} \tag{14}$$

C. Variables definition in digital form

It should be noted that, theoretically, discrete variable structure control can not be obtained by means of simple equivalence from their continuous counterpart [20]. In fact, the principles of sliding mode in discrete-time systems differs from the continuous-time SM systems in power converters, which continuous-time SM have been more extensively investigated and used [21]. Contrary to ideal continuous-time systems which the system trajectory moves along the sliding surface, the trajectory may move around the surface in the discrete-time sliding mode. therefor it is usually called the quasi-sliding mode (QSM). As previously mentioned, according to (13), the continuous- time state space model of buck converter is as follows:

$$\dot{\boldsymbol{X}} = \boldsymbol{A_c X} + \boldsymbol{B_c}u + \boldsymbol{D_c} \tag{15}$$

By considering the discretization of above-mentioned continuous-time system, the discrete-time model can be written as:

$$\boldsymbol{X}(k+1) = \boldsymbol{AX}(k) + \boldsymbol{B}u(k) + \boldsymbol{D} \tag{16}$$

Where X(k) is the matrix of state variables at the moment k (k-th instant) and is defined as X(k) = X(kT), where T is the sampling interval. X(k) consists of two variables $x_1(k)$ and $x_2(k)$, accordance to (7), are the output error voltage (between reference voltage and output voltage) and its derivative, respectively. As the following relations:

$$\mathbf{X}(\mathrm{k})=\begin{bmatrix}\mathrm{x}_1(\mathrm{k})\\ \mathrm{x}_2(\mathrm{k})\end{bmatrix}, \quad x_1(k)=e(k)$$
$$x_2(k)=e(k)-e(k-1) \tag{17}$$

Where:

$$e(k)=V_{ref}-V_o(k)$$

The relationship between the continuous-time and discrete-time coefficients of state equations, namely, the coefficients of (15) and (16), are as follows [13]:

$$\boldsymbol{A}=e^{\boldsymbol{A_c}T}, \qquad \boldsymbol{B}=\int_0^T e^{\boldsymbol{A_c}\lambda}\,d\lambda\,.\boldsymbol{B_c}$$
$$\boldsymbol{D}=\int_0^T e^{\boldsymbol{A_c}\lambda}\,d\lambda\,.\boldsymbol{D_c} \tag{18}$$

Where according to (14), A_c and B_c and D_c are as follows:

$$\boldsymbol{A_c}=\begin{bmatrix}0 & 1\\ -\frac{1}{LC} & -\frac{1}{RC}\end{bmatrix},$$
$$\boldsymbol{B_c}=\begin{bmatrix}0\\ -\frac{1}{LC}V_{in}\end{bmatrix}, \qquad \boldsymbol{D_c}=\begin{bmatrix}0\\ \frac{1}{LC}V_{ref}\end{bmatrix} \tag{19}$$

D. Explanation of sliding surface

In this paper, we propose a discrete-time sliding surface in the form of a linear combination of error, error derivative, and the one-step delayed value of error integral, according to the following equation:

$$s(k)=c_1x_1(k)+c_2x_2(k)+c_3E(k-1) \tag{20}$$

Where c_1, c_2 and c_3 are sliding coefficients. E(k-1), which is an error integral, is actually the sum of the error until the moment (k-1), and is defined as:

$$E(k-1)=E(k)-e(k)=\sum_{i=0}^{k}e(i)-e(k) \tag{21}$$

E. Design of the control law

After defining the sliding surface, the next step is to define a control law. The purpose of control law is to direct the state variables towards the sliding surface and hold on it, so that, will reach a desired operating point. In practice, the control law maintain the motion of state variables within a small vicinity around the switching surface.

In fact, the control law consists of two parts, one section directs the state variables towards the sliding surface (u_{sw}), and the another part maintains state variables around the sliding surface, which is equivalent control law (u_{eq}). Therefore, the sliding mode control law, is written as follows:

$$u_{sm}(k)=u_{eq}(k)+u_{sw}(k)$$
$$=u_{eq}(k)+k_s sign\big(s(k)\big) \tag{22}$$

Assuming that there is disturbance, we consider the discrete-time model of the buck converter as follows (in fact, we write the relation (16) as follows):

$$\boldsymbol{X}(k+1)=\boldsymbol{AX}(k)+\Delta\boldsymbol{AX}(k)+\boldsymbol{B}u(k)+\Delta\boldsymbol{B}u(k)+\boldsymbol{D}$$
$$+\Delta\boldsymbol{D}+\boldsymbol{W}(k) \tag{23}$$

Where ΔA, ΔB and ΔD are parameter perturbations and W(k) is external disturbances. In fact, the purpose of control is that, in spite of the uncertainties, directes X(k) towards the desired value of zero. We model these uncertainties with N(k):

$$\boldsymbol{N}(k)=\Delta\boldsymbol{AX}(k)+\Delta\boldsymbol{B}u(k)+\Delta\boldsymbol{D}+\boldsymbol{W}(k)$$

So the relation (40) is written as follows:

$$\boldsymbol{X}(k+1)=\boldsymbol{A}X(k)+\boldsymbol{B}u(k)+\boldsymbol{D}+\boldsymbol{N}(k) \tag{24}$$

Because disturbance N(k) has not been known in the current time step, therefor we consider one-step delayed estimation of it [22, 23]:

$$\widehat{\boldsymbol{N}}(k)=\boldsymbol{N}(k-1)$$

Accordimg to (24) we have the following relation:

$$\boldsymbol{N}(k-1)=\boldsymbol{X}(k)-\boldsymbol{AX}(k-1)-\boldsymbol{B}u(k-1)$$
$$-\boldsymbol{D} \tag{25}$$

The equivalent control law, $u_{eq}(k)$, is obtained from the solution of the following equation [24]:

$$\Delta s=s(k+1)-s(k)=0 \tag{26}$$

In fact, as mentioned, $u_{eq}(k)$ is used to maintain the state variables around the switching surface (on the switching surface in the ideal case), and since on the switching surface we have $s(k)=0$, therefore, the derivative will be zero on it. From Solving drivative of the sliding surface equal to zero, the equivalent control law is obtained. Substituting the sliding surface (20) in (26) gives:

$$\boldsymbol{CX}(k+1)+c_3E(k)-s(k)=0$$

Assuming $\boldsymbol{C}=\begin{bmatrix}c_1\\ c_2\end{bmatrix}$ and $\boldsymbol{X}(k+1)=\begin{bmatrix}x_1(k+1)\\ x_2(k+1)\end{bmatrix}$:

Taking into account (24) and (25):

$$\boldsymbol{CAX}(k)+\boldsymbol{CB}u(k)+\boldsymbol{CD}+\boldsymbol{CN}(k-1)+c_3E(k)$$
$$-s(k)=0$$
$$u_{eq}(k)=(\boldsymbol{CB})^{-1}\big(-\boldsymbol{CAX}(k)-\boldsymbol{CD}-\boldsymbol{CN}(k-1)$$
$$-c_3E(k)+s(k)\big) \tag{27}$$

By substituting s(k) from (20), another relation for the equivalent control law is obtained as follows:

$$u_{eq}(k) = (\boldsymbol{CB})^{-1}\big((\boldsymbol{C}-\boldsymbol{CA})\boldsymbol{X}(k) - c_3 x_1(k) - \boldsymbol{CD} - \boldsymbol{CN}(k-1)\big) \quad (28)$$

Now, by replacing $N(k-1)$ from (25), the relation $u_{eq}(k)$ will be:

$$u_{eq}(k) = (\boldsymbol{CB})^{-1}\big(-\boldsymbol{CAX}(k) + \boldsymbol{CAX}(k-1) - c_3 x_1(k) + \boldsymbol{CB}u(k-1)\big) \quad (29)$$

Regarding (22), the control law is obtained as follows:

$$u_{sm}(k) = (\boldsymbol{CB})^{-1}\big((\boldsymbol{C}-\boldsymbol{CA})\boldsymbol{X}(k) - c_3 x_1(k) - \boldsymbol{CD} - \boldsymbol{CN}(k-1)\big) - (\boldsymbol{CB})^{-1} k_s sign\big(s(k)\big) \quad (30)$$

In this equation, we used (44) as $u_{eq}(k)$, but each of the relations (27) and (29) can also be used. We can replace the signum function with saturation function in order to reduce the chattering. In Fig. 2, the proposed DSM-controlled synchronous buck converter is presented. In this figure, we assume $CA = [\gamma_1 \;\; \gamma_2]$ and we use the equivalent control law according to (29).

F. Stability analysis

For verifying the stability of the proposed system, we want to prove that if the proposed controller (30) is applied to the described system in (24) and with the sliding surface presented in (20), then the quasi-sliding mode occurs in a finite number of steps.

Proof: Inserting the sliding mode control law ((30) with the equivalent control law acording to (27)) in relation s(k+1) gives:

$$\begin{aligned} s(k+1) &= \boldsymbol{CAX}(k) + \boldsymbol{CB}u(k) + \boldsymbol{CD} + c_3 E(k) \\ &= \boldsymbol{CAX}(k) \\ &\quad + \boldsymbol{CB}\Big((\boldsymbol{CB})^{-1}\big(-\boldsymbol{CAX}(k) - \boldsymbol{CD} \\ &\quad - \boldsymbol{CN}(k-1) - c_3 E(k) + s(k)\big) \\ &\quad - (\boldsymbol{CB})^{-1} k_s sign\big(s(k)\big)\Big) + \boldsymbol{CD} \\ &\quad + c_3 E(k) \\ s(k+1) &= s(k) + \boldsymbol{C}\big(\boldsymbol{N}(k) - \boldsymbol{N}(k-1)\big) \\ &\quad - k_s sign\big(s(k)\big) \end{aligned} \quad (31)$$

In reference [23], the following definitions have been provided in order to examine the existence conditions:

1. If $|s(k)| \le \varepsilon$, we say that QSM occurs in the ε vicinity of the sliding surface s(k) (ε is the quasi-sliding mode band (QSMB)).

2. The system (24) satisfies the quasi sliding mode reaching conditions in the ε vicinity of the switching surface s(k)= 0, if for each k> 0, the following conditions are met:

$$\begin{aligned} &if \quad s(k) > \varepsilon \quad \rightarrow \quad -\varepsilon \le s(k+1) < s(k) \\ &if \quad s(k) < -\varepsilon \quad \rightarrow \quad s(k) < s(k+1) \le \varepsilon \\ &if \quad |s(k)| \le \varepsilon \quad \rightarrow \quad |s(k+1)| \le \varepsilon \;, \end{aligned} \quad (32)$$

3. When QSMB is zero, QSM is converted to an ideal QSM.

Now, in view of the definitions 1, 2, and 3, we want to obtain the restrictions imposed on k_s and N(k) so that QSM is realized in a finite number of steps:

Assumption: The change rate of disturbances or $\dot{N}(t)$ is bounded, in other words, $\dot{N}(t)$ is less than or equal to a positive constant value of δ:

$$\begin{aligned} &\dot{N}(t) = \frac{N(k) - N(k-1)}{T} \\ &\boldsymbol{N}(k) - \boldsymbol{N}(k-1) = \dot{\boldsymbol{N}}(t)T \\ &|\boldsymbol{N}(k) - \boldsymbol{N}(k-1)| \le \delta T \\ &|\boldsymbol{C}(\boldsymbol{N}(k) - \boldsymbol{N}(k-1))| \le \boldsymbol{C}\delta T \end{aligned}$$

in order to satisfy (32) (The second definition), in consideration of (31), the limitation on k_s can be expressed as:

$$C\delta T < k_s < \varepsilon + C\delta T \quad (33)$$

It means that by choosing k_s in this range, The second definition will be always true for (31).

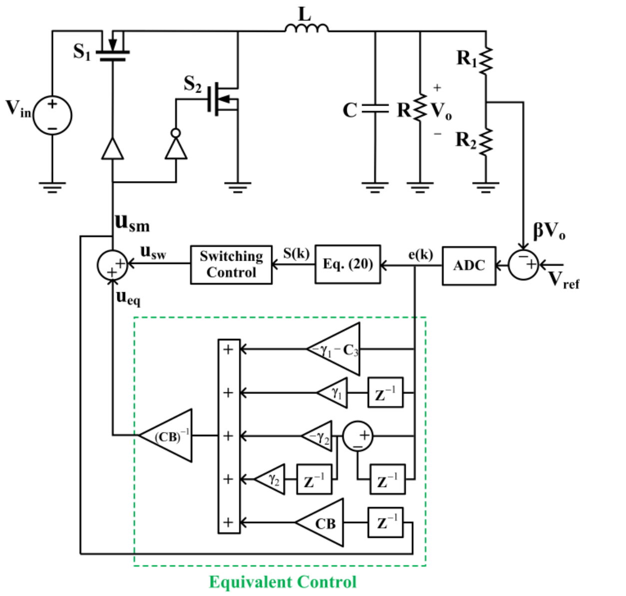


Fig. 2. Diagram of the proposed DSM-controlled Synchronous buck converter.

III. SIMULATION RESULTS

In order to demonstrate the performance of the proposed DSM-controlled synchronous buck converter, simulation results are provided to validate the theoretical design. Simulations are carried out using the Simulink of Matlab. The specifications of the synchronous buck converter are listed in Table I. To satisfy the existence conditions, the ratio of the c_1 (proportional coefficient) to c_2 (differential coefficient) must be greater than twice the frequency of the LC filter [25], as follows:

$$\left(\frac{c_1}{c_2}\right)_{min} = \frac{2}{2\pi\sqrt{LC}}$$

To attain this, the values of $c_1 = 0.1$ and $c_2 = 3.33e\text{-}5$ were selected and the integral coefficient (c_3) was set to value of 0.01. In the following, simulation results are given for the proposed control law.

Fig. 3 shows the simulated start up response of the synchronous buck converter with the proposed control method, that the input and output voltage are 38V and 5V, respectively, and the load current is 2A. As shown in Fig. 3, the ripple value is 0.055 V, which is 1.1%. Step-up line changes, step-down line changes, step-up load transient, and step-down load transient have been shown in Fig. 4. step-up and step-down line transients performance of proposed method have been shown in Fig. 5 (the input voltage is changed to 26% of its value). As can be seen, the ripple value increases by almost 1.54% in transition conditions. Fig. 6 present load transient response. It is observed that the maximum overshoot is 0.065 V (1.3%) and the maximum undershoot is 0.09 V (1.8%).

TABLE I
COMPONENT VALUES OF THE SYNCHRONOUS BUCK CONVERTER

Description	Paremeter	Nominal Value
Input voltage	Vin	38 V
Output voltage	Vout	5 V
Inductor value	L	115 µH
Capacitor value	C	100 µF
Load resistance	R	2.5 Ω
Reference voltage	Vref	2.5 V
Switching frequency	fs	100 kHz
Voltage divider ratio	β	0.5
Equivalent series resistanc	ESR	0.3 Ω

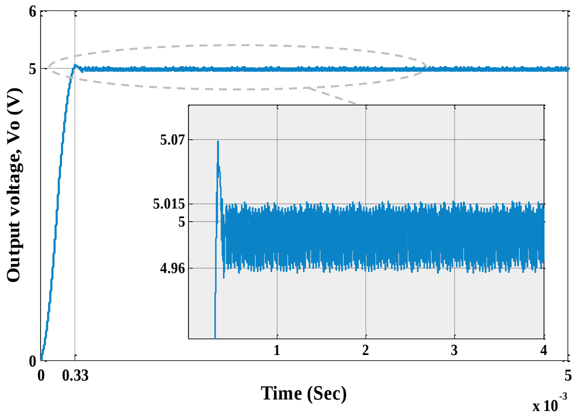
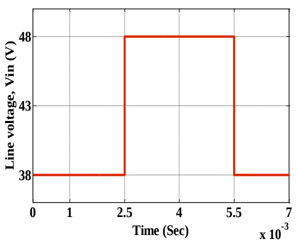
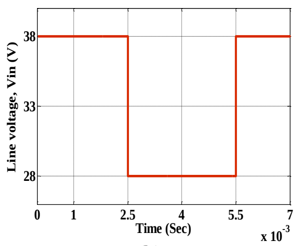


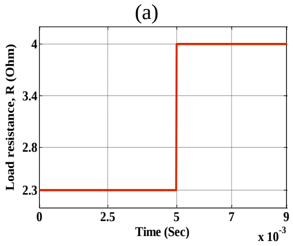
Fig. 3. start up and steady-state response of output voltage that Vin=38 V, Vout=5 V, R=2.5 Ω .



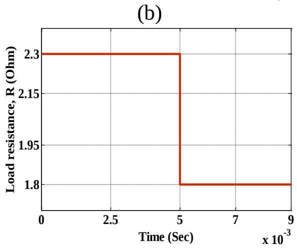
(a)



(b)

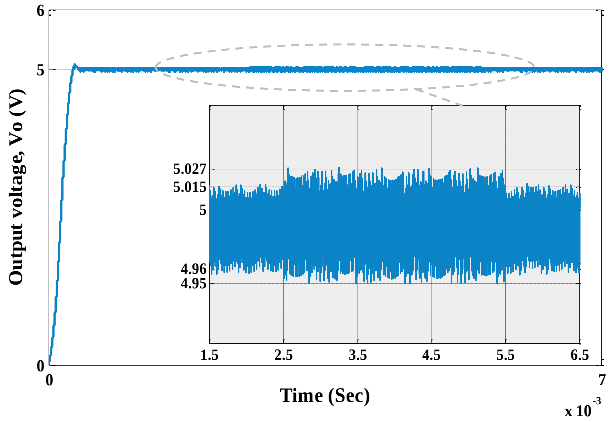


(c)

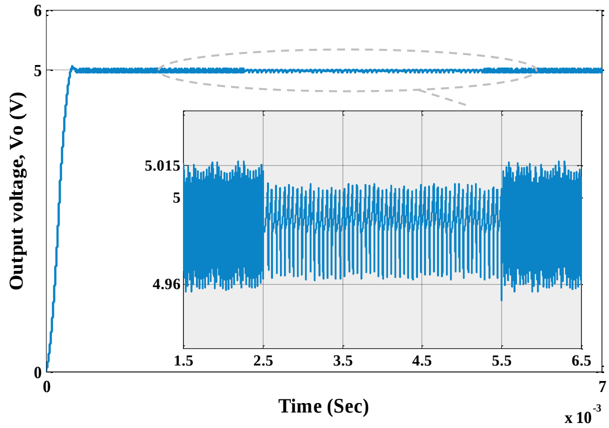


(d)

Fig. 4. (a) Step-up line changes (34% of its value) (b) step-down line changes (26% of its value) (c) step-up load transient (d) step-down load transient.

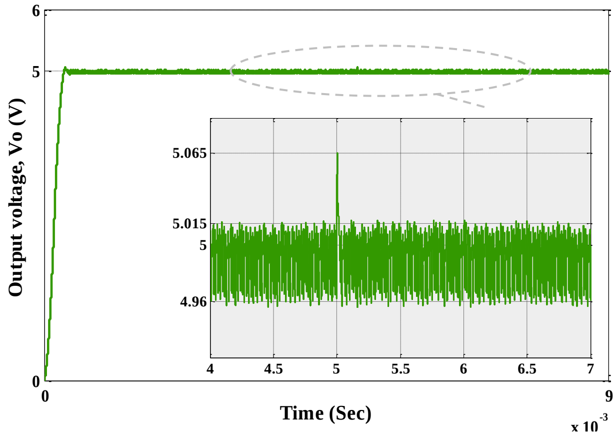


(a)

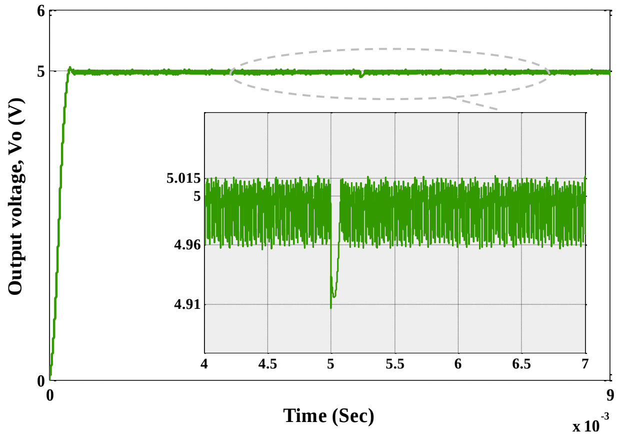


(b)

Fig. 5. Output voltage when line voltage has step-up transient (Fig. 4a), (b) output voltage when line voltage has step-down change (Fig. 4b).



(a)



(b)

Fig. 6. Transient response of output voltage for (a) step-up load change (Fig. 4c), (b) step-down load transient (Fig. 4d).

In Table II, the performance summary of the proposed method has been compared to two of the articles. it shows that in the proposed DSM-controlled synchronous buck converter, chattering, overshoot and undershoot have been reduced acceptably.

TABLE II
PERFORMANCE COMPARISON BETWEEN THIS WORK AND OTHER WORKS

Parameter	[16], 2015	[26], 2016	[27], 2019	This work
Vin (V)	2.8-3.8	5-10	24	38
Vout (V)	1-2.5	1.25	10	5
Vout ripple	23mV (1.9%)	32mV (2.6%)	N/A	0.055V (1.1%)
Switching frequency	1MHz	100KHz	100KHz	100 KHz
Settling time	5μs	N/A	1.6 ms	0.33ms
Overshoot	N/A	100mV (8%)*	0.25V (2.5%)	0.06V (1.3%)
Undershoot	60 mV (5%)	50mV (4.1%)*	0.35V (3.5%)	0.09V (1.8%)
Settling time line reg.	20μs (for 20% line jump)	N/A	1.5ms (for 16% line jump)	~0 (for 26% line jump)
Overshoot for line reg.	3%	N/A	0.25V (2.5%)	0.03%

* Not reported,extracted from figures.

IV. CONCLUSION

In this paper Digital sliding-mode control approach is applied to synchronous buck dc–dc converters. Because the SMC analog implementation requires several amplifiers, and suffer from temperature and voltage variations, digital implementation of SMC as well as stability analysis presented in this paper. In fact, in the proposed DSMC, a stable output voltage has been achieved, and it minimizes chattering, the steady state error, the overshoot and undershoot of the output voltage.

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