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# Sliding Mode Control of the DC-DC Fly back Converter with Zero Steady-State Error

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## ABSTRACT

In this paper a novel approach for controlling the output voltage of the isolated flyback converter is presented using sliding mode controller. Due to non-minimum-phase nature of the converter and presence of a right-half-plan-zeroin voltage transfer function, indirect regulation of the output voltage is applied. In the designed controller, simultaneous use of output voltage and transformer current feedbacks improves dynamic response of the controller. The sliding mode controller is developed so that in the final control law, integral of the output voltage error is present and consequently the output voltage error will be zero in the steady-state. Also, in order to measure transformer magnetizing current, simple and efficient method is proposed. Finally, in order to evaluate response of the controller, flyback converter is a simulated using MATLAB / Simulink.
**KEYWORDS**: Sliding mode controller, steady-state voltage error, two-loop control, magnetizing inductance current measuring, non-minimum phase nature

## 1 – INTRODUCTION

In recent years, the use of linear compensators in power electronic converters have been widely replaced by nonlinear controllers and successful applications of feedback linearization[1], adaptive backstepping [2], passive based control[3] and sliding mode [4] have been reported in closed-loop voltage regulation of DC-DC converters. Considering the simplicity of implementation, sliding mode controllers have been used more and recently some efforts have been done to employ this controller in industrial applications. Generally, from practical point of view, it is possible to implement this controller based on simple analog circuits and is not necessary to use digital processors. Since calculation of the control law in sliding mode method is not time consuming, it possible to increase the switching frequency of the converter to relatively large values. It is worth noting that, increment of switching frequency reduces the output filter size, improves the dynamic response of the converter and... . DC-DC converters are implemented up to 200kHz (switching frequencies) based on sliding mode controllers [5]. In addition, stability of the controller and its robustness against changes in line voltage and load resistance are the outstanding features of the sliding mode approach. However, using this controller in DC-DC converters is associated with two major problems:

1– Variation of the switching frequency

2- Steady-state error

Variation of the switching frequency leads to lower-order harmonic components. Filtering of these components will be complicated. It is proven that the life-time of the power switches is relatively low in this condition [6]. To solve this problem,at least three main approachesare proposed:

1- Adaptive changes in hysteresis bandwidth [7]

2–Application of a fixed- frequency clock functions in the control circuit [8]

3- Indirect implementation of thesliding mode based on equivalent control [9]

These three methods are compared comprehensively in [10].Due to fast dynamic response and ease of implementation, equivalent control is used in this paper.

In addition to steady-state error, non-minimum phase nature of the flybackDC-DC converter is another major challenge in controlling the output voltage of the converter.To overcome this problem, indirect regulation of the output voltage based on the inductor current loop in non-minimum phase converters has been used [4].The general block-diagram of this method is given in Fig.1.

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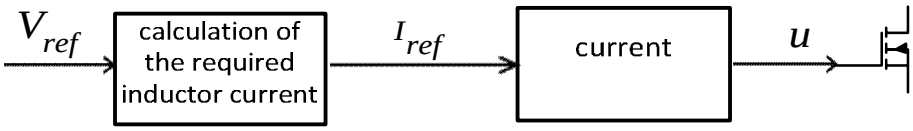


*Fig.1: indirect voltage control in non-minimum phase DC-DC converters*

In this method, considering the reference value of the output voltage($V_{ref}$), the required inductor current ($I_{ref}$) is calculated at first. Then using a suitable current controller, the power switch of the converter is controlled so that the current of the inductor be equal to $I_{ref}$.Although it is possible to calculate the relation between inductor current and output voltage in indirect control based on steady-state analysis, but it should be considered that:

1- In order to calculate the required inductor current, the exact values of the input voltage and load resistance must be known. Since these values are uncertain, indirect controller which is illustrated in Fig.1 cannot respond satisfactorily.

2- Usually during modeling of the converter, effects of the parasitic elements are not considered. For example, voltage drop across the power switch and diode, equivalent series resistance of the output capacitor, losses associated with the transformer and inductorare not considered in the equations governing the dynamic behavior and steady-state response of the converter. Although the impact of these elements may be negligiblein low-power applications, but when the load resistance becomes considerably small, the effects of these parasitic elements cannot be ignored. More importantly, these elements are strictly depended on the temperature and operating point of the converter and basically cannot be assumed constant.

Subsequently it is clear that the relation between $I_{ref}$ and $V_{ref}$ cannot be calculated based on steady-state behavior of the converter. In the structure shown in Fig. 1, the absence of direct feedback from the output voltage can also lead to the steady-state error (in addition to steady-state error of the sliding mode controller). To solve this problem, in [8] and [10], a direct feedback of the output voltage is added during calculation of the reference current. Although the steady-state error associated with non-minimum phase nature of the converter has beensolved in these papers, but the presented results clearly show that the inherent error of the sliding mode controller (which is used to implement the current controller) is not zero.

Based on our little search and try, recently no reported paper where found which applies the sliding mode controller to flyback converterwith zero steady-state error. Considering the popularity of this converter in industrial applications and its ability to provide multiple isolate outputs, in this paper a novel sliding mode controller is presented to eliminate steady-state error of the output voltage in continuous conduction mode (CCM).Assuming this scope, generation of the required inductor current and also development of the sliding mode controller is considered so that in the final control law, the integral of the output voltage error will be present and as a result, steady-state error of the system will be eliminated.Also, due to practical difficulties in measuring the magnetizing inductor current, a simple method is proposed using the power switch current. It is worth noting that the simultaneous use ofthe output voltage and the inductor current feedbacks can improve the dynamic response of the converter considerably[12].

The organization of this paper is as following: firstly,the state-space averaged model of the converter is studied. Then transfer functions of the flybackpower supply are calculated and non-minimum phase nature of the converter is proved. In section III, sliding mode theory based on the equivalent controlis described briefly and then a novelindirect sliding mode controller is developed to regulate the output voltage of the converter. Finally, in order to evaluate accuracy of the proposed method, flyback converter is simulated using MATLAB / Simulink software.

**2– Operation principal of the flyback DC-DC converter and its modeling:**

**2-1–ideal condition:**Flyback converter power circuit is shown in Fig.2. In this figure, $L_m$represents equivalent magnetizing inductance of the transformer which is referred to the primary side. The role of this inductor in the operation of the flybackconverter is completely important and its relative size determines the operating mode (continuous or discontinuous) of the converter. For example, if operation of the converter in CCM is required, magnetizing inductance must have acceptable value and this should be considered during transformer design.

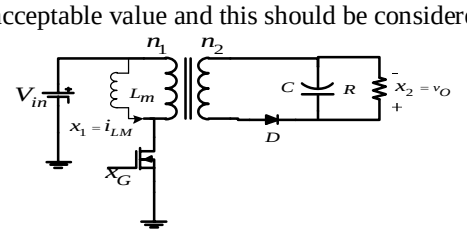


*Fig.2: Power circuit of the flyback converter ($L_m$ represents equivalent magnetizing inductance of the transformer)*

In CCMregion,twodifferentmodesshould beconsidered according to switch positions: If the power switchis turned on,considering the transformersecondaryvoltage polarity, diode will bereversedbias. Inthis case,the output voltagewill besuppliedwiththeoutput capacitorC. Also,according to positivevoltagedrop across the$L_m$,inductor current($i_{Lm}$)willincreaselinearly. When power switchis turned off,$i_{Lm}$flowsinto theprimarywindingandthediodewill beforwardbias. Inthis case,the energywhich is stored in theinductorwill betransferred to theload resistorandoutput capacitor.

**2-2–Steady-stateanalysis of the converter:**During on state of the power switch, equivalentcircuit ofthe converterisshown in Fig.3-a.

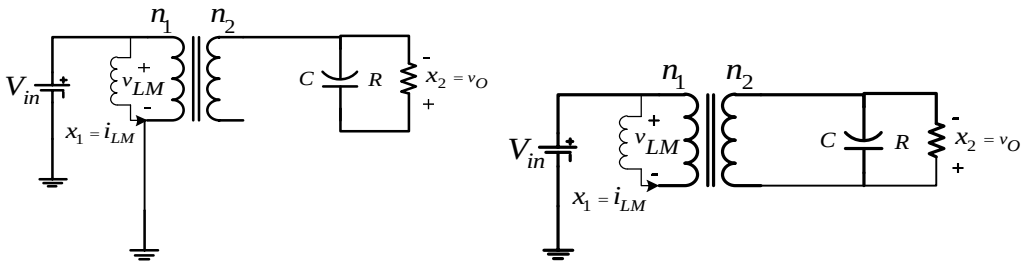


*Fig.3: equivalent circuits of the flyback converter*
*(a) power switch is turned on and diode is reverse biased*
*(b) power switch is turned off and the diode is forward bias*

Obviously in this case secondary winding of the transformer is open circuit. Considering inductor voltage, $i_{Lm}$ could be written as:

$$V_{Lm} = +V_{in} \Rightarrow i_{Lm} = \frac{V_{in}}{L_m}t + I_1 \qquad 0 \le t \le t_{on} \qquad (1)$$

In this equation, $I_1$represents the inductor whenthe power switch is turned on. At $t = t_{on}$ (when the power switch is turned off), inductor current can be calculated as below:

$$I_2 = \frac{V_{in}}{L_m}t_{on} + I_1 \qquad (2)$$

When the power switch is turned off, subsequently the diode D will be on. In this case,$i_{Lm}$ can be written as:

$$V_{Lm} = -\frac{n_1}{n_2}V_o \Rightarrow i_{Lm} = \left(-\frac{n_1}{n_2}\right)\frac{V_o}{L_m}(t - t_{on}) + I_2 t_{on} \le t \le T_S \qquad (3)$$

where$T_S$ is switching period of the converter. In the steady-state conditions,at $t = T_S$, inductor current($i_{Lm}$) will be equal to $I_1$.Considering$D = \frac{t_{on}}{T_S}$, the relation between input and output voltages can be written as:

$$V_o = \frac{n_2}{n_1}\frac{D}{1-D}V_{in} \qquad (4)$$

Magnetizing inductor voltage and current waveforms are shown in Fig.4. In this case, if integral of the inductor voltage in a period is equated to zero,equation (4) can be resulted.

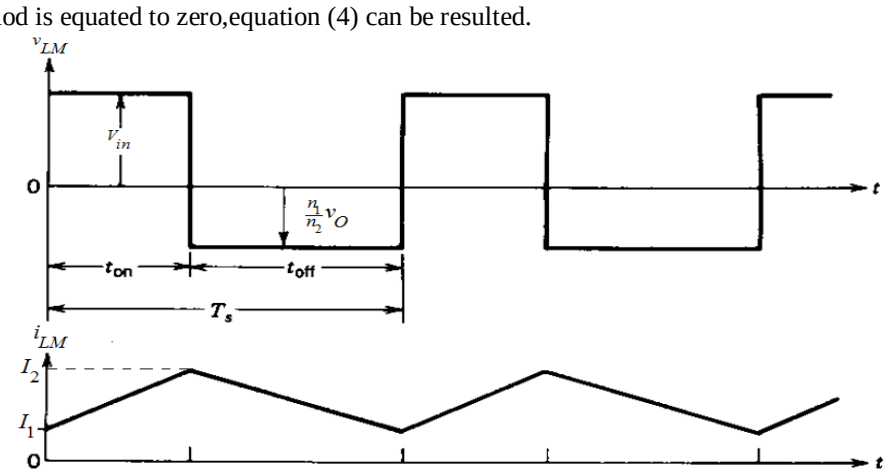


*Fig.4: steady-state voltage and current of the magnetizing inductor in CCM*

**3-2- Proposed method for measuring average value of the inductor current:**As mentioned before, in order to design sliding mode controller, state variables of the system including inductor current and capacitor voltage should be measured.Obviously, magnetizing inductor current cannot be measured directly and in this section, a simple approach is proposed for measuring the inductor current based on the power switch current ($i_s$).Considering Fig.4, average value of the $i_{Lm}$ can be calculated simply considering the area under the related curve in a period:

$$I_{Lm} = \frac{1}{2}(I_1 + I_2) \qquad (5)$$

Switch current waveform is given in Fig.(5).Similarly, power switch average current can be written as following:

$$I_S = \frac{1}{2}(I_1 + I_2)D \qquad (6)$$

Considering (5) and (6), the following equation is concluded:

$$I_{Lm} = \frac{I_S}{D} \qquad (7)$$

Also according to (4), duty cycle (D) of the converter can be calculated as below:

$$D = \frac{V_o}{V_o + \frac{n_2}{n_1}V_{in}} \qquad (8)$$

Obviously, in the steady-state, the output voltage of the converter will be equal to reference value ($v_o = V_{ref}$). Considering equations (7) and (8), the average value of the magnetizing inductor currentcan be calculated:

$$I_{Lm} = \frac{V_{ref} + \frac{n_2}{n_1}V_{in}}{V_{ref}} I_S \qquad (9)$$

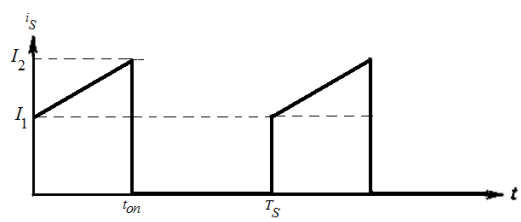


*Fig.5: power switch current in CCM*

**2-4 - -State-space averaged modeling of the flyback converter:** steady-state behavior of the converter is presented in previous sections. State-space modeling is simple and powerful method for studying the dynamic behavior of the power electronic converters. Inductor current and output voltage of the converter are considered as state variables:

$$\boldsymbol{X}^T = (x_1, x_2) = (i_{Lm}, v_o) \qquad (10)$$

During the on state of the power switch, considering equivalent circuit which is shown in Fig.3-a,the following equations can be written:

$$\dot{\boldsymbol{X}} = \boldsymbol{A_{on}}\boldsymbol{X} + \boldsymbol{B_{on}} \qquad 0 \le t \le t_{on} \qquad (11)$$

$$\boldsymbol{A_{on}} = \begin{pmatrix} 0 & 0 \\ 0 & -\frac{1}{RC} \end{pmatrix}$$

$$\boldsymbol{B_{on}} = \begin{pmatrix} \frac{V_{in}}{L_m} \\ 0 \end{pmatrix}$$

If power switch is turned off, the following state equations can be calculated according to Fig.3-b:

$$\dot{\boldsymbol{X}} = \boldsymbol{A_{off}}\boldsymbol{X} + \boldsymbol{B_{off}} t_{on} \le t \le T_S \qquad (12)$$

$$\boldsymbol{A_{off}} = \begin{pmatrix} 0 & -\frac{n_1}{n_2}\frac{1}{L_m} \\ \frac{n_1}{n_2}\frac{1}{C} & -\frac{1}{RC} \end{pmatrix}$$

$$\boldsymbol{B_{off}} = \begin{pmatrix} 0 \\ 0 \end{pmatrix}$$

Considering averaging theory in power electronic converters modeling [11], state-space model of the flyback converter can be written as:

$$\dot{\boldsymbol{X}} = \boldsymbol{A_{avg.}}\boldsymbol{X} + \boldsymbol{B_{avg.}} \qquad 0 \le t \le T_S \qquad (13)$$

$$\boldsymbol{A_{avg.}} = D\boldsymbol{A_{on}} + (1-D)\boldsymbol{A_{off}} = \begin{pmatrix} 0 & -\frac{n_1}{n_2}\frac{1}{L_m}(1-D) \\ \frac{n_1}{n_2}\frac{1}{C}(1-D) & -\frac{1}{RC} \end{pmatrix}$$

$$\boldsymbol{B_{avg.}} = D\boldsymbol{B_{on}} + (1-D)\boldsymbol{B_{off}} = \begin{pmatrix} \frac{V_{in}}{L_m}D \\ 0 \end{pmatrix}$$

**2-5- Transfer functions of the flyback converter:** Generally, in DC-DC converters, state-variables and control input (duty cycle) are composed of the dc steady-state quantities and small ac perturbations. Therefore:

$$i_{LM} = I_{LM} + i_{lm}$$

*(14)*

$$v_O = V_O + v_o$$

$$D = \Delta + d$$

Replacing (14)in the averaged state-space model of the converter (equation 13) results in:

$$\begin{pmatrix} \dot{I}_{LM} \\ \dot{V}_O \end{pmatrix} + \begin{pmatrix} \dot{i}_{lm} \\ \dot{v}_o \end{pmatrix} = \begin{pmatrix} 0 & -\frac{n_1}{n_2}\frac{1}{L_m}(1-\Delta) \\ \frac{n_1}{n_2}\frac{1}{C}(1-\Delta) & -\frac{1}{RC} \end{pmatrix}\begin{pmatrix} I_{LM} \\ V_O \end{pmatrix} + \begin{pmatrix} \frac{V_{in}}{L_m}\Delta \\ 0 \end{pmatrix} + \begin{pmatrix} 0 & -\frac{n_1}{n_2}\frac{1}{L_m}(1-\Delta) \\ \frac{n_1}{n_2}\frac{1}{C}(1-\Delta) & -\frac{1}{RC} \end{pmatrix}\begin{pmatrix} i_{lm} \\ v_o \end{pmatrix} + \begin{pmatrix} 0 & -\frac{n_1}{n_2}\frac{1}{L_m}(-d) \\ \frac{n_1}{n_2}\frac{1}{C}(-d) & 0 \end{pmatrix}\begin{pmatrix} I_{LM} \\ V_O \end{pmatrix} + \begin{pmatrix} 0 & -\frac{n_1}{n_2}\frac{1}{L_m}(-d) \\ \frac{n_1}{n_2}\frac{1}{C}(-d) & 0 \end{pmatrix}\begin{pmatrix} i_{lm} \\ v_o \end{pmatrix} + \begin{pmatrix} \frac{V_{in}}{L_m}d \\ 0 \end{pmatrix} \quad (15)$$

If only the DC components of the equation (15) are consider:

$$\begin{pmatrix} \dot{I}_{LM} \\ \dot{V}_O \end{pmatrix} = \begin{pmatrix} 0 & -\frac{n_1}{n_2}\frac{1}{L_m}(1-\Delta) \\ \frac{n_1}{n_2}\frac{1}{C}(1-\Delta) & -\frac{1}{RC} \end{pmatrix}\begin{pmatrix} I_{LM} \\ V_O \end{pmatrix} + \begin{pmatrix} \frac{V_{in}}{L_m}\Delta \\ 0 \end{pmatrix} = \begin{pmatrix} 0 \\ 0 \end{pmatrix}$$

*(16)*

Neglecting higher order term where two small ac components are multiplied, linearized state-space model can be written as following:

$$\begin{pmatrix} \dot{i}_{lm} \\ \dot{v}_o \end{pmatrix} = \begin{pmatrix} 0 & -\frac{n_1}{n_2}\frac{1}{L_m}(1-\Delta) \\ \frac{n_1}{n_2}\frac{1}{C}(1-\Delta) & -\frac{1}{RC} \end{pmatrix}\begin{pmatrix} i_{lm} \\ v_o \end{pmatrix} + \begin{pmatrix} \frac{V_{in}}{L_m} + \frac{n_1}{n_2}\frac{1}{L_m}v_o \\ 0 \end{pmatrix} d \quad (17)$$

Applying Laplace transforms to equation. (17) with zero initial conditions, the following equations can be obtained:

$$\begin{pmatrix} i_{lm} \\ v_o \end{pmatrix} = \left[ SI - \begin{pmatrix} 0 & -\frac{n_1}{n_2}\frac{1}{L_m}(1-\Delta) \\ \frac{n_1}{n_2}\frac{1}{C}(1-\Delta) & -\frac{1}{RC} \end{pmatrix} \right]^{-1} \begin{pmatrix} \frac{V_{in}}{L_m} + \frac{n_1}{n_2}\frac{1}{L_m}v_o \\ 0 \end{pmatrix} d \quad (18)$$

From Control view-point, each of the system state variables can be considered as an output. Note that, the inductor current ($i_{lm}$) and output voltage ($v_o$) of the converter are controlled by the duty cycle adjustment and $d$can be defined as a control input. Flyback converters transfer functions are defined as:

$$H_1(s) = \frac{v_o}{d} \qquad H_2(s) = \frac{i_{lm}}{d} \quad (19)$$

Considering (18), transfer functions can be obtained as:

$$H_1(s) = \frac{V_{in}\left(R - sL_m\frac{n_1}{n_2}\Delta/(1-\Delta)^2\right)}{s^2L_mRC(\frac{n_1}{n_2})^2 + sL_m\frac{n_1}{n_2} + R(1-\Delta)^2}$$

*(21)*

$$H_2(s) = \frac{V_{in}\left(1+\Delta+sRC\frac{n_1}{n_2}/(1-\Delta)\right)}{s^2L_mRC(\frac{n_1}{n_2})^2 + sL_m\frac{n_1}{n_2} + R(1-\Delta)^2}$$

*(22)*

Equation (21) obviously shows that assuming $v_o$ as an output and controlling it based on the duty cycle adjustment results in a non-minimum phase system. Therefor direct controlling of the output voltage due to the presence of a right-half-plan-zero will result in unstable behavior of the converter. On the other hand, such problem is not seen in the second transfer function and inductor current can be controlled simply by choosing a suitable compensating network.

**2-6- Indirect control of the flyback converters output voltage:** The overall structure of the controller which is used in this paper is shown in Fig.6.

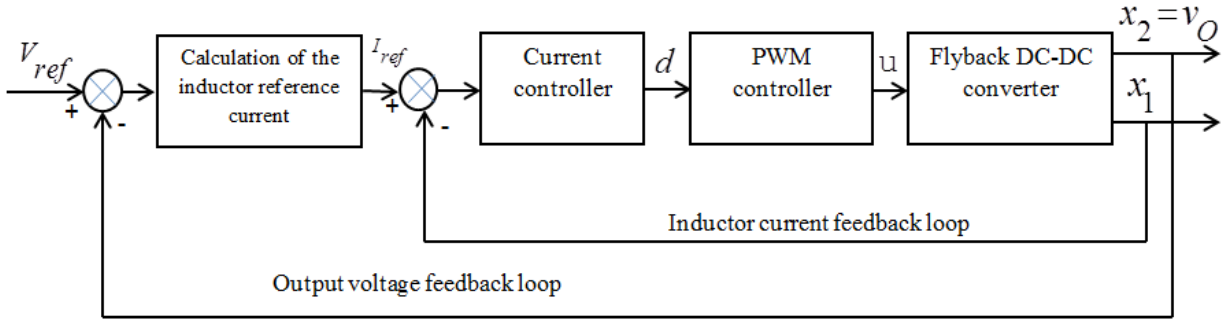


*Fig.6: indirect output voltage control in flyback converter based on the inductor current regulation*

As mentioned in the previous section, due to the presence of a right-half-plan- zero in the output voltage transfer function,in flyback DC-DC converters the inductor current feedback loop must be used. First, considering the output voltage error, desired value of inductor current is calculatedin an outer loop. Then using a current controller, converter is switched suitably to obtain $i_{LM} = I_{ref}$ in an inner loop.This structure is called two-loop control and its design process based on sliding mode current controller is presented in the next section.It is worth noting that the internal current loop can improve the dynamic responseof the system.

**3 - Controller design for flyback converters using sliding mode:**

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

**3-1 - Theory of the applied sliding mode controller:** Suppose that, the nonlinear system model on$R^n$is assumedas below. Suppose the origin of the coordinate as an operating point of the system in steady-state.

$$\dot{\boldsymbol{z}} = \boldsymbol{Az} + u\boldsymbol{Bz} \quad (23)$$

where$A$ and $B$ (square n×n) matricesare fixed with real components. The scalar control function $u$takes values 0 and 1. In the sliding mode control method, $u$is considered as[12]:

$$u = \frac{1}{2}(1 + sgnS(\boldsymbol{z})) \quad (24)$$

where in this equation, $sgn$is the sign function symbol and $S(\boldsymbol{z})$ is called sliding surface. Equation (24) states that:

$$if\ \ S(\boldsymbol{z}) > 0\ \ \Rightarrow u = 1 \text{ and} if\ \ S(\boldsymbol{z}) < 0\ \ \Rightarrow u = 0$$

Necessary and sufficient conditions for the existence of sliding motion on the sliding surface can be written as [12]:

$$\lim S \times \dot{S} < 0 \quad (25)$$

The smooth control function for which nonlinear system which its model is written in (23), adopts sliding surface as a local integral manifold is known as *equivalent control* and isshown by $u_{eq}$. Equivalent control can be calculated by equating derivative of the sliding surface to zero [12]:

$$\frac{dS}{dt} = 0 \Rightarrow \left[\frac{\partial S}{\partial z}\right]^T \dot{\boldsymbol{z}} = 0 \quad (26)$$

Considering that the value of $\dot{\boldsymbol{z}}$is given in (23), the controller can be obtained according to (26):

$$u_{eq} = -\frac{\left[\frac{\partial S}{\partial z}\right]^T \boldsymbol{Az}}{\left[\frac{\partial S}{\partial z}\right]^T \boldsymbol{Bz}} \quad (27)$$

The sliding motion exist locally on the sliding surface, if and only if,$u_{eq}$ satisfies the following condition [12]:

$$0 < u_{eq} < 1 \quad (28)$$

**3-2- Inductor current reference generation:**Usually in two-loop control method, the reference currentis considered as a coefficient of the voltage error:

$$i_{ref} = K[V_{ref} - v_o] \quad (29)$$

where$V_{ref}$ and$v_o$indicate converter reference and output voltage values.Also $K$is the voltage gain error. Choosing a large value for $K$reduces steady-state error. On the other hand, an increase in $K$may lead to system instability. The main difficulty in using equation (29) for generating a reference current is presence of the steady-state error. This error value will change with respect to load, input voltage and ambient temperature. So in this article, the reference is considered as:

$$i_{ref} = K_P\left[V_{ref} - v_o\right] + K_I \int (V_{ref} - v_o)dt \quad (30)$$

Such an idea eliminates steady-state error of the system completely and on the other hand, doesn't increase complexity of the controller so much.

**3-3 –Sliding surface selection:** usually sliding surfacecan be considered as a linear combination of the system state variables:

$$S(\boldsymbol{z}) = \alpha_1 z_1 + \alpha_2 z_2 + \alpha_3 z_3 \quad (31)$$

where $\alpha_1$,$\alpha_2$and $\alpha_3$are fixed design gains.During development of the sliding mode controller, the selected state variables shouldhave zero steady-state values. For this reason, the output voltage and inductor current errors are considered as state variables $z_1$and$z_2$. In fact, both of the output voltage and inductor current errors will be used in the current controller design. Use of the voltage error in the inner loop increases accuracy of the proposed controller. In fixed frequency sliding mode, the control is not perfect and in this case, it is commonly known that, an additional controlled state variable ($z_3$) should be introduced to reduce the steady-state error[13]:

$$z_1 = i_{ref} - i_{LM} \quad (32)$$

$$z_2 = V_{ref} - v_o$$

$$z_3 = \int (z_1 + z_2)dt$$

It must be noted that selected state-variable in (32) will not eliminate steady-state error directly. For instance, although in references [8] and [10] the same state-variables are used in DC-DC (boost and buck) converters control, but presented results clearly show steady-state error in the output voltage.

**3-4- Equivalent controller design with zero steady-state error:**Considering equations (32) and behavior of the converter in CCM, dynamic model of the flyback converter can be written as:

$$\dot{z}_1 = \frac{d}{dt}\left(i_{ref} - i_{LM}\right) = -\frac{K_P}{C} i_C + K_I\left(V_{ref} - v_o\right) - \frac{uV_{in} - (1-u)\frac{n_1}{n_2}v_o}{L_m} \quad (33)$$

$$\dot{z}_2 = \frac{d}{dt}\left(V_{ref} - v_o\right) = -\frac{1}{C} i_C$$

$$\dot{z}_3 = z_1 + z_2 = (K_P + 1)\left(V_{ref} - v_o\right) - i_{LM} + K_I \int \left(V_{ref} - v_o\right)dt$$

In these equations, $u$is thelogic state of the power switch($= 0$is related to off state of the switch and$u = 1$is related to on state). Also, $i_C$ represents output capacitor current. Considering equation (33), equating time-derivative of sliding surface to zero, sliding mode controller can be calculated as:

$$u_{eq} = \frac{1}{V_{in} + K_2 v_o}\left[K_1\left(V_{ref} - v_o\right) + K_2 v_o + K_3 i_C + K_4 i_{LM} + K_5 \int \left(V_{ref} - v_o\right)dt\right] \quad (34)$$

where$K_1$ to $K_5$ are fixed controller gains:

$$K_1 = L_m\left[K_I + \frac{\alpha_3}{\alpha_1}(K_P + 1)\right] \quad (35)$$

$$K_2 = \frac{n_1}{n_2}$$

$$K_3 = -L_m \frac{\alpha_3}{\alpha_1}$$

$$K_4 = -\frac{L_m}{C}$$

$$K_5 = L_m \frac{\alpha_3}{\alpha_1} K_I$$

As it is clear, thedeveloped control law includes integral of the voltage error. Therefor proposed control method - which regulates the output voltage of the flyback converter –can eliminate steady-state error.

## 4 – SIMULATION RESULTS AND DISCUSSION

In this section, flyback DC-DC converter has been simulated based on the developed controller (equation 30) using MATLAB/Simulink. The block diagram of the proposed sliding mode controlleris shown in Fig.7. In this case, in order to implement the sliding mode controller, output voltage and inductor current are sampled. Due to switching, measured values may have a significant ripple.Large voltage and current ripple deteriorate operation of

the controller. Therefor application of the low-pass filters and calculation of the state variables average values in power electronic converters controller design are completely accepted.Also, considering the difficultiesin direct measurement of the inductor current, the proposed method in section 2-3 is used.Converter and controller parameters are listed in Table I. The maximum step size for all simulations is taken 100ns.

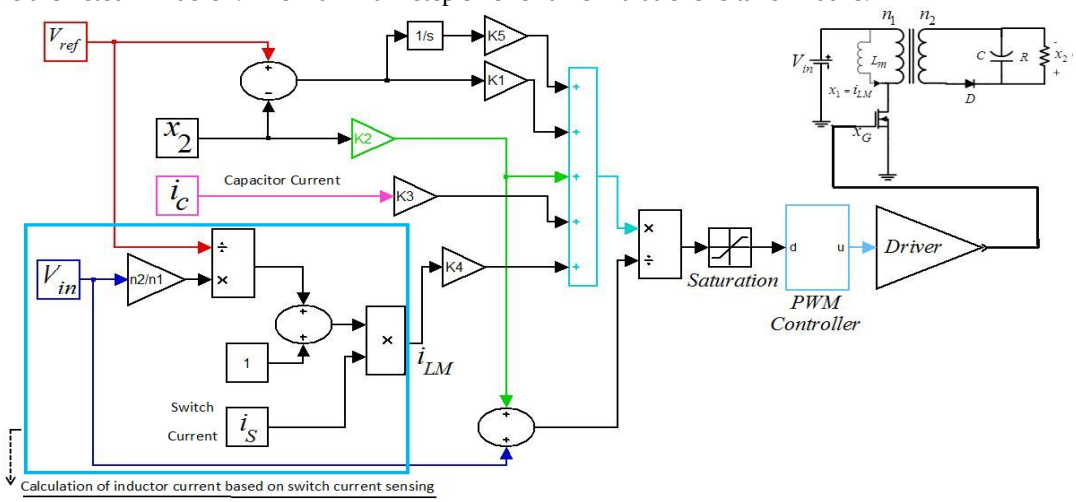


*Fig.7: block diagram of the proposed controller*

*Table I: nominal specifications of the simulated converter and controllers gains*

| | | |
|---|---|---|
| 1- | Input voltage ($V_{in}$): | 10 V |
| 2- | Magnetizing inductor ($L_m$): | 100μH |
| 3- | Output capacitor (C): | 470μF |
| 4- | Load resistance (R) | 10 Ω |
| 5- | Switching frequency ($f_s$): | 40kHz |
| 6- | Transformer turn ration($N_1/N_2$): | 0.5 |
| 7- | Output voltage reference ($V_{ref}$): | 20 V |
| 8- | Controller gain $K_1$: | 0.1 |
| 9- | Controller gain $K_2$: | 0.5 |
| 10- | Controller gain $K_3$: | 0.3 |
| 11- | Controller gain $K_4$: | 2.5 |
| 12- | Controller gain $K_5$: | 300 |

In order to evaluatethe overallresponseof the proposed controller to changes inloadandinput voltage, differenttestsareconsideredindetail.

**4-1- Changes in the reference voltage:**InFig.8, the response of the designedslidingmodecontrollerforreference voltagechanges isillustrated. Considering theparameters given inTable I, the referencevoltageis stepped at$t = 0.1s$from 10V to 30V. It is clear that in spite of large changes in the reference, the controller is able tofollowthe desiredvaluesandthe correspondingsteady-state erroris zero. Also,waveforms of the input voltage, output currentandduty cycleareplotted.

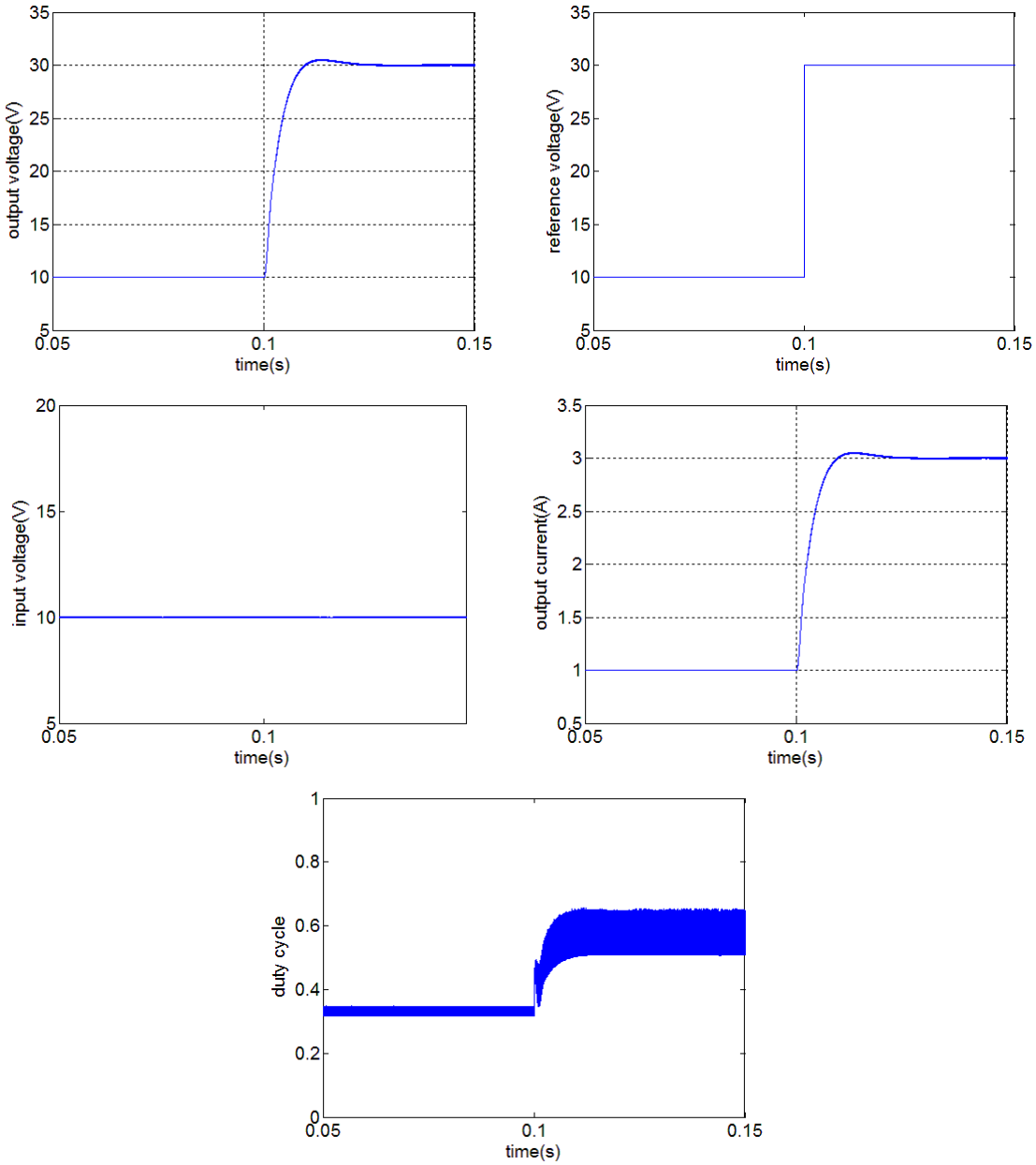


*Fig.8: response of the proposed controller to reference voltage changes*

**4-2- Response of the proposed controller to load changes:**Response of the proposed controller to step changes in load resistance is presented in Fig.8. In this test, the load resistance is stepped from $10\Omega$to$3.3\Omega$ at$t = 0.1s$. In spite of load changes in wide range (300 percent increment in load current), the proposed controller has good dynamic and steady state response. During this test, the reference voltage of the converter is 20V.

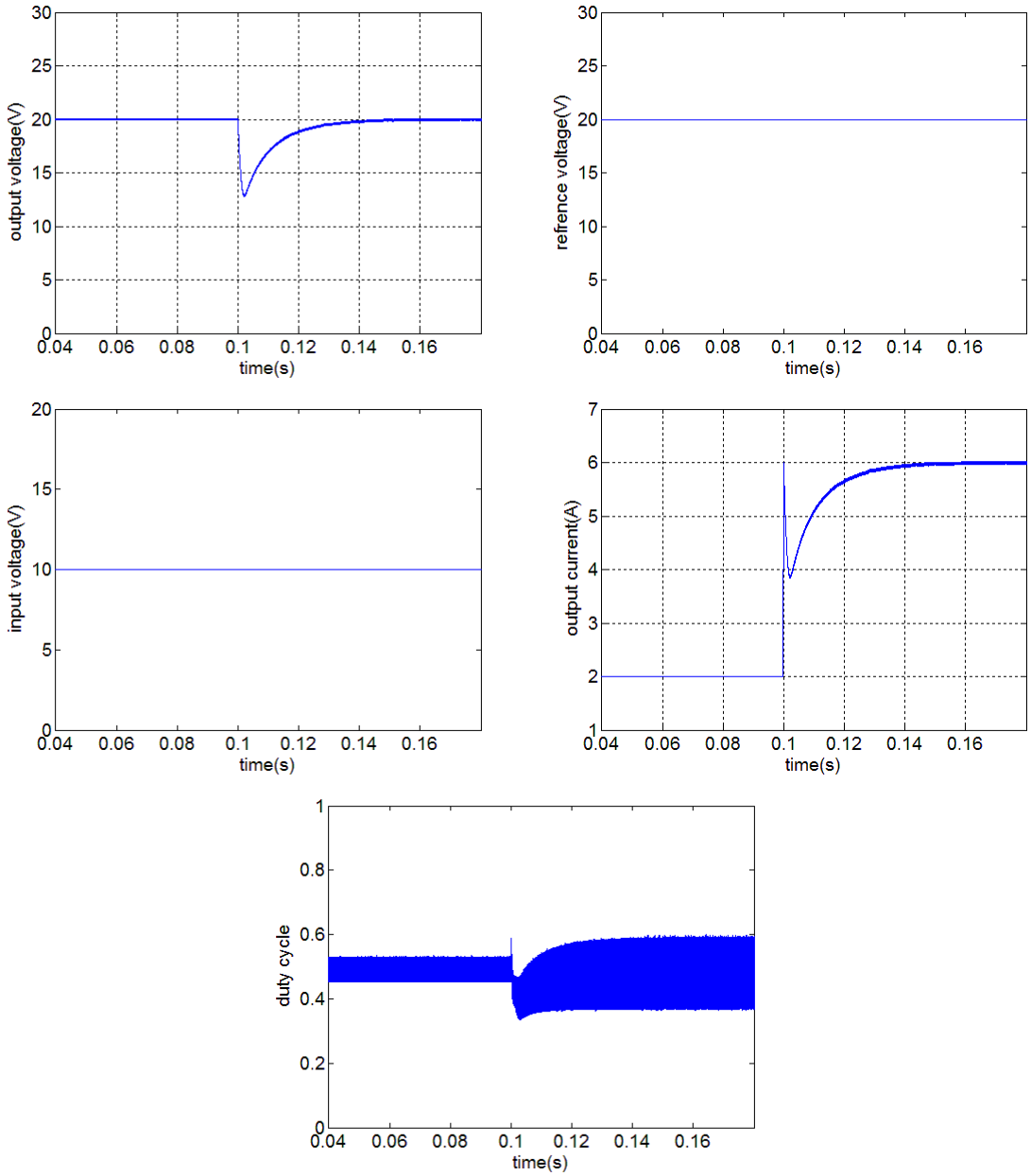


*Fig.9:response of the proposed sliding mode controller to load resistance changes*

**4-3–Response of the proposed controller to input voltage changes:** Usually uncontrolled diode rectifiers are used to implement input voltage source of the DC-DC converters. For this reason,the controllerresponsetoinput voltagevariationsis important. The response of the proposed controller to step changes in input voltage is illustrated in Fig.10.At$t = 0.1s$,the converterinput voltageis increased from 10V to15V. Fig.10clearlyshowsthestability ofthe proposedcontroller to variation of the input voltage.

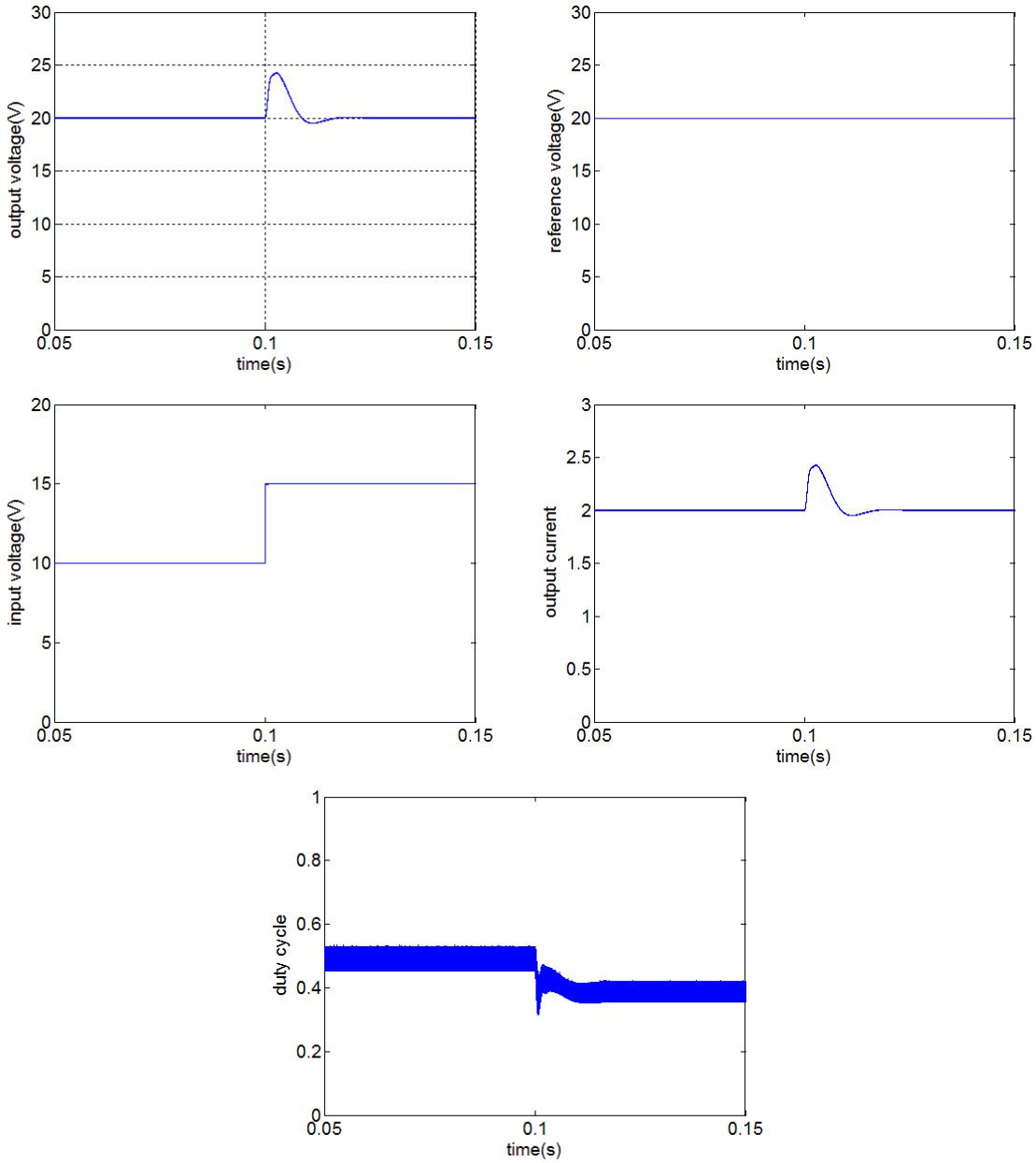


*Fig.10: response of the proposed controller to input voltage changes*

**4-4-Simultaneous changes in input voltage, load resistance and reference voltage:** In order toevaluate theoverallperformance ofthe developed sliding mode controller, input voltage, load resistance and reference voltage of the system are changed simultaneously and the response of the controller is shown in Fig.11.At$t = 0.1s$,input` voltageis stepped from 10V to15V, reference voltage from 10V to 30Vandfinally load resistanceis changed from$10\Omega$to$3.3\Omega$simultaneously.

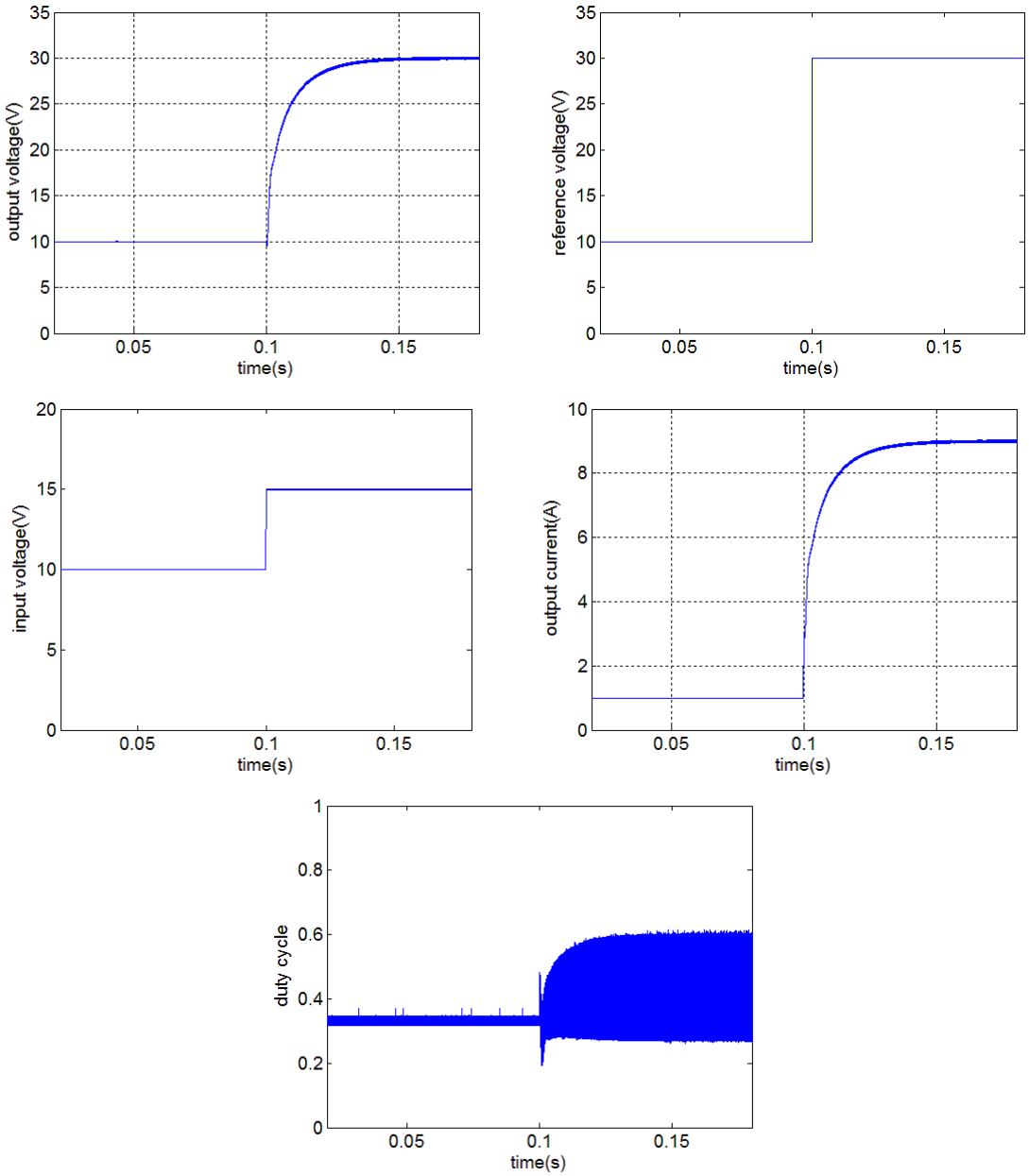


*Fig.11: response of the controller to simultaneous changes in input voltage, load resistance and reference voltage*

**Conclusions:** In this paper, a novel sliding mode controller is presentedbased on equivalent control method to adjust the output voltage of the DC to DC flyback converters with constant frequency.Due to the non-minimum phase nature of the system, indirect control of output voltage is applied according to two-loop control. Reference current generation in the outer loop and selection of the sliding surface in the inner loop are considered carefully. Presence of the output voltage error integral in the final control law is resulted in zero steady state error. Also asimple method is used to measure average value of the inductor current. The developedcontrol method is simulated based on MATLAB/ Simulink software. Simulation results clearly show that, in spite of large changes in input voltage, load resistance and output voltage reference, proposed method can control output voltage of the converter stably with zero steady-state error.

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