Synchronous Buck-Boost Converter for Energy Harvesting Application

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Abstract **- In this paper, the synchronous buck-boost converters are developed. Unlike the traditional buck-boost converters, the synchronous converter has fast transient response, similar to the behaviour of the buck converter with synchronous rectification. In addition, it has a non-pulsating output current. The synchronous buck-boost converter operates in the Continuous Current Mode (CCM) which, not only reduces the stress on the output capacitor, but also reduces the ripple of the output voltage. Simulation results are provided to demonstrate the effectiveness of the proposed control system.**

Keywords—Synchronous buck-boost, CCM, Stress, Ripple voltage

I. INTRODUCTION

In many applications such as portable devices, electronic devices in cars, etc., where the output voltage range of the battery is considerably large, buck-boost converters are required. There are numerous types of buck-boost converters such as the non-isolated Cuk converter [1], the SEPIC converter [2], the Zeta converter [3] and the Sheppard-Taylor topologies [4-5]. However, corresponding feedback regulators that would ensure fast close-loop transients as well as high stability are difficult to design. In addition to that, each of these topologies requires two inductors instead of one, increasing thus the cost and bulkiness of the system. Also, their small signal model is a fourth order one, making the control design more difficult and complex. In comparison with these converters mentioned previously, the proposed 2D KY converter [6] has an ultra-fast transient response, similar to the behaviour of the buck converter. Moreover, this converter operates in continuous current mode (CCM) which reduces the stress on the output capacitor and decreases the output ripple. In this paper, the detailed operation of the synchronous buck-boost converter is first illustrated, and then a mathematical representation of the converter is developed both in the state-space and the frequency domain. Based on the proposed model, a linear feedback voltage regulator is designed to ensure high transient performance. Simulation results are finally presented to demonstrate the effectiveness of the control system.

II. SYNCHRONOUS BUCK-BOOST CONVERTER

Fig. 1 shows the synchronous buck-boost converter's topology. It consists of four power MOSFETs Q1, Q2, Q3 and Q4 with anti parallel diodes. It consists also of a diode D, an output inductor L, an output capacitor C0 and an energy transfer capacitor C which is large enough to maintain a constant voltage across itself, which is equal to the input voltage. Here the output of the converter is controlled by the PIPWM because of this controller is high accuracy and more reliable compare with other converter.

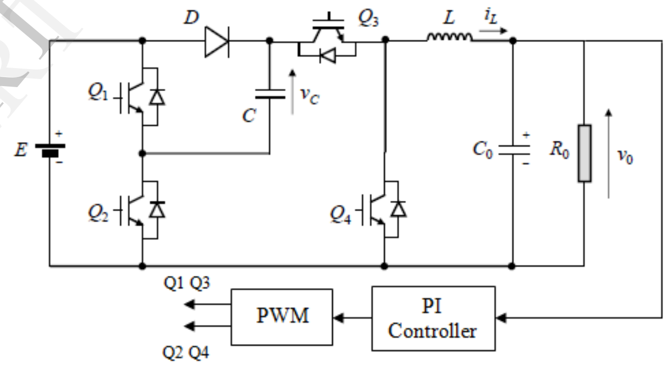


Fig. 1. Synchronous buck-boost topology with its voltage control circuit

III. PRINCIPLE OF OPERATION

The converter generates an output voltage v0 across the load represented by R0 from an assumed ideal voltage source E. The current flowing through the inductor L is designated by i_L. The pair of switches (Q_1, Q_3) has a same control signal characterized by a duty cycle d and a switching period T. Similarly, the pair (Q_2, Q_4) is controlled synchronously. In the Continuous Current Mode (CCM) operation, the converter has the two successive configurations:

State 1: $0 < t < dT$. The pair (Q_1, Q_3) are turned ON and (Q_2, Q_4) are turned OFF, as illustrated in Fig. 2. The diode D is not conducting and the intermediate capacitor C is discharging. In this case, the voltage across L is equal to the input voltage ($E+v_c-v_0$), which causes the magnetization of the inductor.

The state equations for this configuration are as follows:

$$L\frac{di_L}{dt} = E + v_c - v_o \qquad (1.a)$$

$$C_0\frac{dv_0}{dt} = i_L - \frac{v_0}{R_0} \qquad (1.b)$$

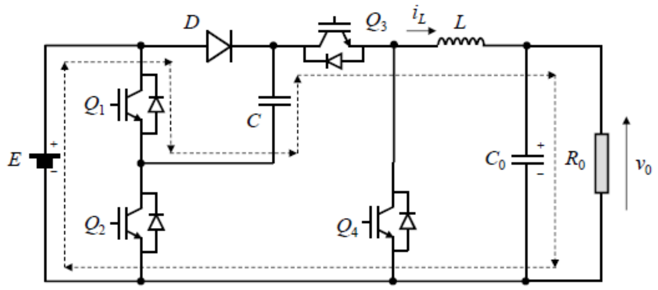


Fig.2. Current path in state 1 configuration

State 2: dT < t < T. The pair (Q_2, Q_4) are now turned ON and (Q_1, Q_3) are turned OFF, as depicted in Fig. 3. The diode D conducts the source current and allows, thus, the instantaneous charging of capacitor C. During this interval, the voltage across C is constantly equal to E, and the voltage across inductor L is equal to ($-v_0$) causing the demagnetization of L. The resulting state equations are as follows:

$$L\frac{di_L}{dt} = -v_0 \qquad (2.a)$$

$$C_0\frac{dv_0}{dt} = i_L - \frac{v_0}{R_0} \qquad (2.b)$$

$$v_C = E \qquad (2.c)$$

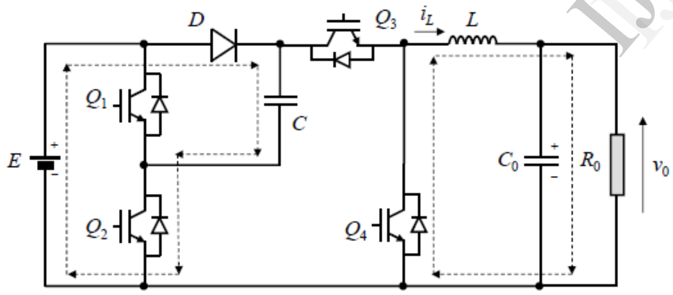


Fig.3. Current path in state 2 configurations

By neglecting the switching ripple in the inductor current and the capacitors voltages, we get in the steady state:

$$v_C \cong V_C = E \qquad (3.a)$$

$$\frac{V_0}{E} = 2D \qquad (3.b)$$

$$I_L = \frac{V_0}{R_0} \qquad (3.c)$$

Where D, I_L, V_0 and V_C are respectively the static values of the duty cycle, the inductor current, the output voltage and the voltage across the intermediate capacitor. In addition, since the duty cycle range is between 0 and 1, the output voltage can increase from 0 to twice the input voltage, which makes this topology pertain to the buck-boost family of converters.

$$C_0\frac{dv_C}{dt} = -i_L \qquad (1.c)$$

IV. DESIGN CONSIDERATIONS

The value of the inductor L should be high enough to limit at an acceptable value the switching frequency ripple in the current i_L. It yields:

$$L > L_{min} = \frac{V_0(1-D)}{f_s \Delta i_{L,max}} \qquad (4)$$

Where $f_S = 1/T$ is the switching frequency, and $\Delta i_{L,max}$ denotes the admissible value of the current ripple.

Similarly, the output capacitor C_0 should limit the voltage ripple across it and, therefore, should be chosen as follows:

$$C_0 > C_{0,min} = \frac{V_0(1-D)}{4Lf_s^2 \Delta v_{0,max}} \qquad (5)$$

Where $\Delta v_{0,max}$ represents the admissible value of the output voltage ripple.

V. AVERAGED MODEL OF THE SYNCHRONOUS CONVERTER

The converter must be associated to an adequately designed control circuit to maintain a constant output voltage. Any change in the input voltage and/or the load current can cause in open-loop an output voltage different from that desired. For that, a feedback control law is necessary to compensate the voltage gap and bring quickly the output voltage to the desired level. Modeling plays a key role in revealing the dynamic behavior of the converter and provides a basis in designing the control system.

The adopted modeling approach is known as the state-space average modeling technique. It is based on: 1) the formulation of state-space equations for each configuration in a switching cycle, 2) averaging these equations in order to obtain a single state space model, and 3) if they obtained model is nonlinear, the application of a small-signal linearization around a static point, that yields the computation of the transfer functions on the basis of which the linear voltage regulator would be finally designed.

Referring to section III, the converter presents in the CCM operation two configurations in a switching cycle *T*. The elementary state models corresponding to each configuration are given respectively by equations (1) and (2). Combining these two elementary mathematical representations of the converter within a whole switching period leads to the following averaged state-model:

$$\begin{cases} L\dfrac{di_L}{dt} = 2dE - v_0 \\ C_0\dfrac{dv_0}{dt} = i_L - \dfrac{v_0}{R_0} \end{cases} \tag{6}$$

Model (6) is purely linear and the required transfer functions can be obtained naturally without the necessity of applying the small-signal linearization process as it is the case for most of the DC-DC converters. The performance of the linear regulator that will be developed later would thus be unaffected by a variation of the setup point within the whole range of operation.

Applying the Laplace transform to model (6) yields the following duty-cycle-to-input current and duty-cycle-to-output voltage transfer functions:

$$G_{i_L,d}(s) = \frac{2E}{L} * \frac{s + \frac{1}{R_0C_0}}{s^2 + \frac{s}{R_0C_0} + \frac{1}{LC_0}} \tag{7.a}$$

$$G_{v_0,d}(s) = \frac{2E}{LC_0} * \frac{1}{s^2 + \frac{s}{R_0C_0} + \frac{1}{LC_0}} \tag{7.b}$$

VI. PIPWM CONTROL DESIGN

The aim of the feedback control circuit is to regulate the output voltage v_0. This voltage is compared with the reference value V_0, and the resulting error is feed to PI controller output of the PI signal compared to a sawtooth signal using a comparator, as illustrated in Fig. 4.

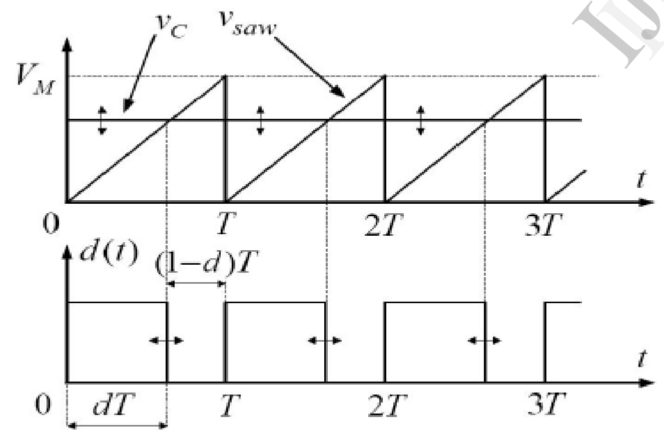


Fig. 4. Generation of the switches gate signals

This energy harvesting application note describes a simple implementation of a discrete Proportional Integral (PI) controller. When working with applications where control of the system output due to changes in the reference value or state is needed, implementation of a control algorithm may be necessary. Examples of such applications are motor control, control .of temperature, pressure, flow rate, speed, force or other variables. The PI controller can be used to control any measurable variable, as long as this variable can be affected by manipulating some other process variables. Many control solutions have been used over the time, but the PI controller has become the 'industry standard' due to its simplicity and good performance.

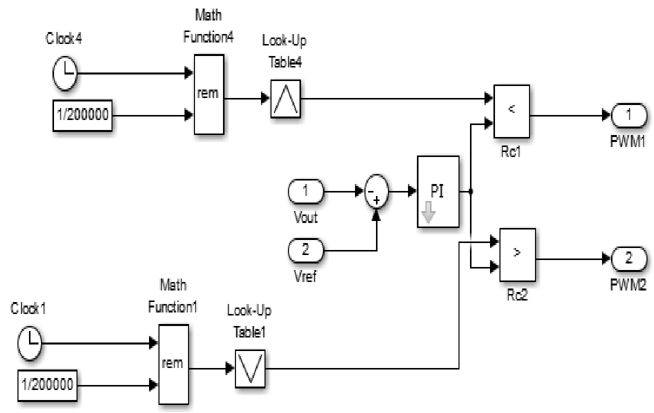


Fig. 5. Simulink PI-PWM gate signals for switches

In Fig. 5 a schematic of a system with a PI controller is shown. The PI controller compares the measured process value v with a reference set point value, V_0. The difference or error, e, is then processed to calculate a new process input, u. This input will try to adjust the measured process value back to the desired setpoint.

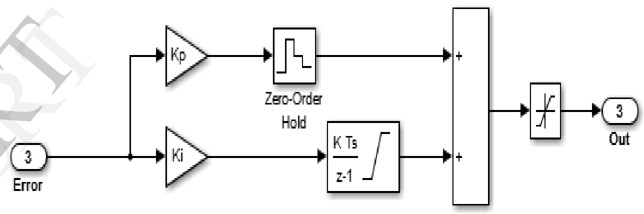


Fig. 6. Discrete PI Controller

The alternative to a closed loop control scheme such as the PI controller is an open loop controller. Open loop control (no feedback) is in many cases not satisfactory, and is often impossible due to the system properties. By adding feedback from the system output, performance can be improved. Fig. 6 shows the discrete PI controller.

VII. SIMULATION RESULTS

The converter of Fig. 1 and its control circuit were implemented numerically using the SimPower Blockset of the Matlab/Simulink tool. The adopted parameters and operating conditions are the following:

- The rated input voltage E is 9 to 16V.
- The rated output voltage V_0 is set to 12V.
- The rated output current is equal to 4A, which corresponds to R0 = 3Ω
- The switching frequency f_S is 200kHz.
- The rise time tm is set to 0.1s.
- L = 14µH, C = 470µF and C_0 = 470µF

The waveforms of the current in the input inductor and the voltage across the output capacitor at rated operating conditions are represented in Fig. 5. In addition, in order to test the dynamics of the control system, an input voltage disturbance, a load variation and set point offset have been applied successively. The system's responses are given respectively in Figs. 7to 11. In Fig. 11 (a) shows zoomed portion of the inductor current, when system start bucking condition in this waveform duty cycle D less the 1-D and Fig. 11 (b) shows zoomed portion of the inductor current, when system start boosting condition in this waveform duty cycle D is grater then 1-D .

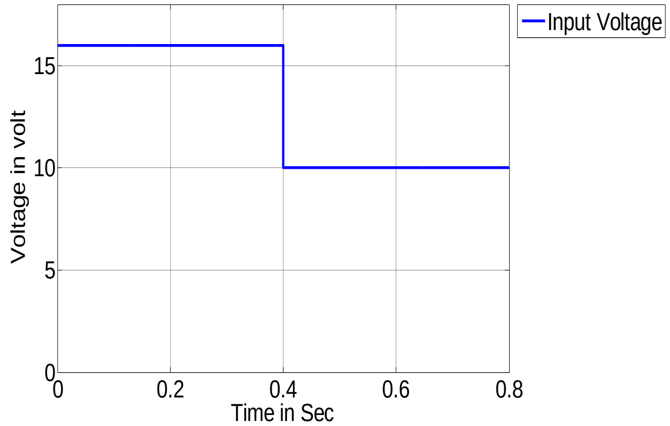


Fig. 7. Waveform of the input voltage decreased from 16V to 10V

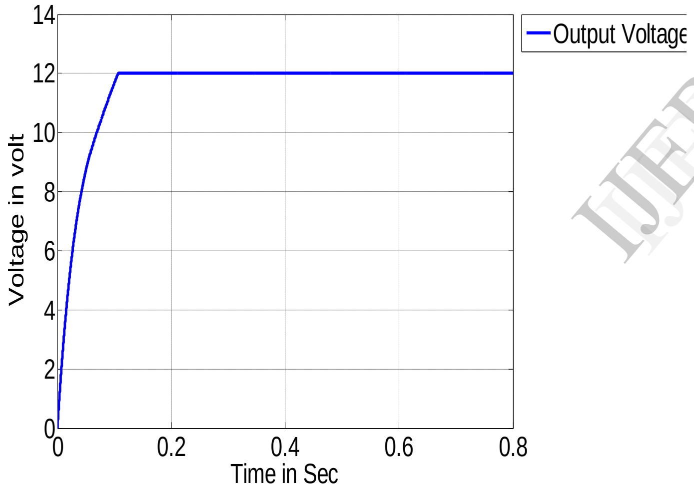


Fig. 8. Waveform of the output voltage ie 12vots

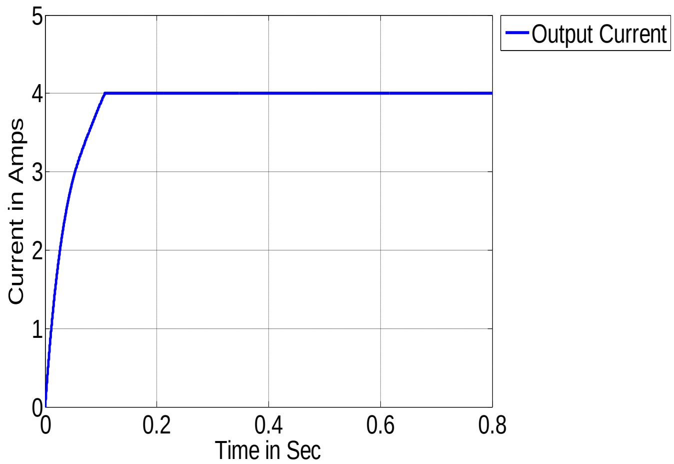


Fig. 9. System response of output current

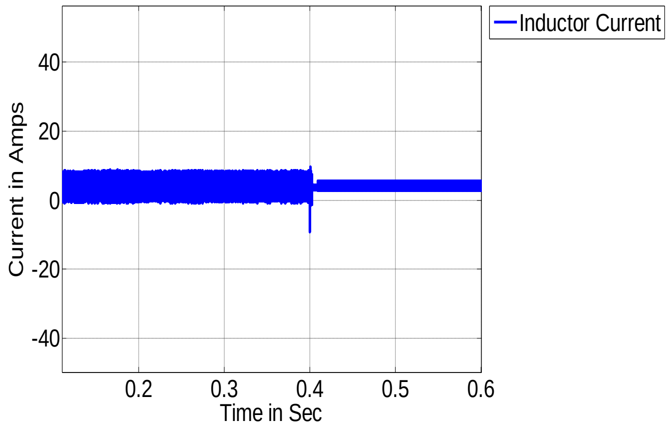
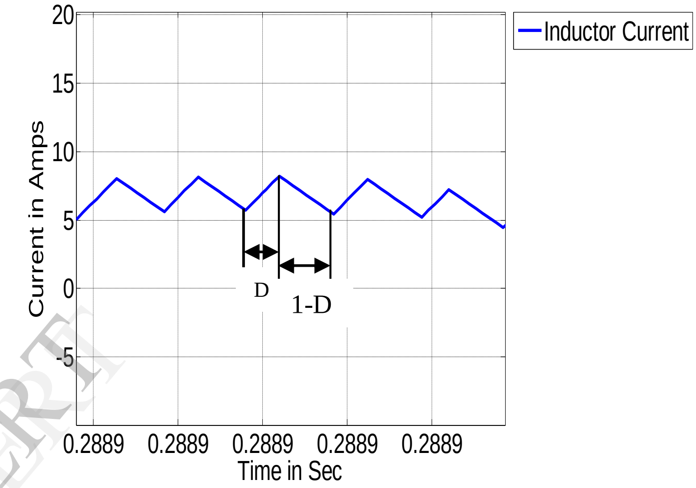
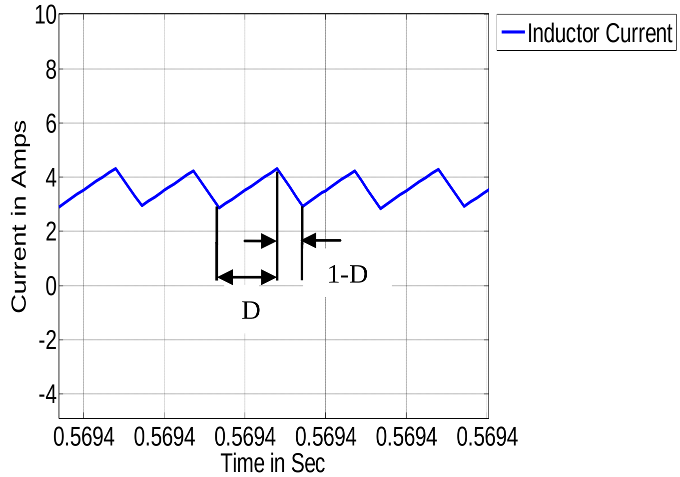


Fig. 10. System response of inductor current



(a)



(b)

Fig. 11. Zoom waveforms of for inductor current

VIII.CONCLUSION

This paper presents the PI-PWM controller and synchronous buck-boost DC-DC converter operating in a continuous current mode. The proposed control system was digitally implemented and tested using the Matlab/Simulink/SimPower simulation tool. It was shown through the obtained results that the converter with its control circuit exhibits high dynamic performance during start-up or following a set point offset. Moreover, the switching between buck and boost modes in this proposed control scheme is nearly smooth. Finally, a 9–16 V input, 12 V output is simulated.

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