A New Variable Frequency Zero Voltage Switching Control Method for Boost Converter Operating in Boundary Conduction Mode

S. Norouzi, H. Ghoreishy*, A. Ale Ahmad, F. Tahami

*Department of Electrical and Computer Engineering, Babol Noshirvani University of Technology, Babol, Iran
Department of Electrical and Computer Engineering, Sharif University of Technology, Tehran, Iran

A B S T R A C T

This paper proposes a new variable frequency zero voltage switching (ZVS) control method for boost converter operating in boundary conduction mode (BCM). The intended method keeps the converter in BCM despite of the load and input voltage variations. This is done by changing switching frequency in a certain specified range. The proposed method can guarantee circuit performance in BCM via zero-crossing detection of the inductor current and changing the switching frequency. In addition, with a slight modification in control structure, it is possible to achieve a fully ZVS in all cases. This converter control is carried out in analog form without using microprocessors which, compared with the digital one, has less noise, cost and processing challenges in high frequency applications. Simulation results obtained from applying the proposed method on a GaN-based synchronous boost converter in two different switching frequency ranges (100KHz and 1MHz) are indicative of the proposed strategy advantages.

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1. INTRODUCTION

DC-DC boost converters are particularly significant due to their high voltage gain, input current continuity, and power factor correction. As it is known, different Pulse Width Modulation (PWM) methods are the well-known control strategies for these converters. Conventionally, in these methods, the voltage on the switch or its current is unexpectedly switched off and on, known as hard switching. Switching losses, high stresses, limitations in higher frequencies, noise amplification, and electromagnetic interferences are among the important shortcomings of hard switching. In general, to minimize the size and volume of the power electronic converters, switching frequency should be increased which, in turn, increases switching losses and Electromagnetic Interferences (EMI). To solve these issues, soft switching methods are widely employed [1, 2].

*Corresponding Author Institutional Email: ghoreishy@nit.ac.ir
(H. Ghoreishy)
Soft switching techniques include Zero Current Switching (ZCS) and Zero Voltage Switching (ZVS). Using soft switching causes switching losses reduction, high efficiency, better temperature control of switches in power electronic converters [3–7].

Resonant converters are the first category of the soft switching converters. Resonance circuits are used for creating soft switching conditions in these converters. One drawback of the resonant converters is the change in structure of the converter topology where the application of passive elements in the system leads to volume and cost increment [8]. One desirable method for the realization of soft switching in DC-DC converters is converter operation in Boundary Conduction Mode (BCM); so that, unlike resonant converters, this is conducted without the slightest change in the structure of the power electronic converter. Designing in BCM mode causes reduction of the input inductor size, loss reduction of switch turn-on time and reduction of some issues regarding reverse recovery compared with design in Continuous Conduction Mode (CCM). In addition, by slightly modifying the controller, the ZVS or near ZVS (also called valley switching) can be easily achieved. Boost Power Factor Correction (PFC) converters are widely used since the switching loss of the power switch can be minimized by ZVS or valley switching [9–11].

Over the past decades, due to space limitations and issues related to weight and cost reduction, the evolution of power electronic systems tended toward systems with high energy density, leading to an efficiency increase [12–15]. High frequency Silicon switch circuits have unacceptable switching loss. Therefore, in order to obtain the appropriate voltage and current ripples, large passive elements are required. The capacitors and inductors mainly comprise a large amount of volume and area. Due to the usage of large capacitors, problems such as reduced converter efficiency and inappropriate ripple have arisen. Nevertheless, designing the converter in BCM generates more current stress, which increases switch conductive losses and size of the input filter. Wide-bandgap (WBG) switches such as SiC (Silicon carbide) and GaN (Gallium nitride) are used to solve this problem due to the low gate charge and the input and output capacitors that operate at a higher switching frequency [16–19].

Regarding properties of GaN transistors, BCM with valley switching in a boost converter was exposed as an ideal application of newly emerged GaN products. This is because the parasitic elements of the devices can be best accommodated at high frequencies i.e. losses of the devices can be minimized [20].

One of the highly significant considerations in designing circuit operation in BCM is the zero-crossing detection of the inductor current. Various methods are proposed for zero-crossing detection of the inductor current and converter operation in BCM. An inductor was used for current monitoring in [21]. Inverse current detection by a freewheeling switch was investigated and implemented in [22]. An RC filter for zero current detection was used in [23] by a series resistance in predictive online digital control method. Some sensors are employed for measuring inductor current, which are not appropriate in high frequency applications due to limitations in bandwidth, price and inaccurate current measurement. Therefore, current mirror technique, which is known as SenseFet in power electronic applications, are used for current monitoring. This technique is a simple and low-cost method with high efficiency [24, 25]. In this method, a GaN switch is used as SenseFet and the voltage drop of the sensor resistor is proportionate with switch current. Bandwidth is not limited in this method and measures current with high accuracy [26].

The BCM performance in DC-DC converters can be guaranteed by controlling the switching frequency. Regarding this matter, Variable on-time (VOT) control, which is a variable frequency technique, has been described in [27]. In this method, input voltage [28] or peak switch current [29] sampling is used to define the slope of the ramp signal, which is compared with the error amplifier output. Another method is the hysteresis control strategy in which the inductor current is limited between upper and lower bands. If the current crosses either bands, the switch control signal will be changed and the converter will remain in BCM [30].

A new method to implement ZVS in BCM with 5 MHz switching frequency is proposed in [31] which increases efficiency up to \( \eta = 98\% \). As presented in [32, 33], having a MHz switching frequency greatly reduces size of the inductor and converter input filter that significantly affects power density of the system. Nonetheless, conditions for a fully ZVS requires input and output voltages to be within a certain range. A ZVS extension method for BCM synchronous converter is presented in [34], such that synchronous rectifier ON-time is increased in bidirectional converter; therefore, a negative inductor current, capable of discharging capacitor of the main switch, is generated. A control method for achieving ZVS is implemented in [35] by eliminating dependency to input and output ratio in a MHz BCM boost based power factor correction circuit.

To obtain the negative inductor current needed for achieving ZVS, sampling of the converter input and output voltages is required [36].

All the above methods were performed assuming that the circuit remains in BCM despite the changes in input and output parameters. However, any research that covers both BCM guarantee and circuit performance in fully ZVS has not been conducted yet.

A new high frequency method for controlling boost converter in BCM is proposed in this paper. The proposed method keeps boost converter in BCM despite of load and input voltage variations by changing...
switching frequency in a certain specified range. After determining the desired frequency at which the converter returns to BCM, the controller keeps the switching frequency in a projected range. The applied converter is a synchronous boost converter using WBG devices. This method can guarantee circuit performance in BCM by zero-crossing detection of the inductor current and by changing switching frequency. In addition, with a slight change in control structure, it’s possible to achieve a fully ZVS in all cases. This converter control is carried out in analog form without using microprocessors which, compared with digital one, has less noise, cost and processing challenges in high frequency applications.

The basic performance of the proposed control method is discussed in Section 2; presents the BCM operation both under valley and extended ZVS. The equations that define the converter operation under these modes are presented in Section 3. The proposed self-regulating control method for ZVS extension operation is proposed in Section 4. Simulation results are examined in Section 5 to ensure the accuracy of the proposed control algorithm method.

2. PROPOSED BOUNDARY CONDUCTION MODE CONTROL IN A BOOST CONVERTER MODEL

The proposed BCM control method is studied in this section. In order to show the desirable performance of this technique, it is applied to a synchronous boost converter in which the diode is replaced by a power electronic switch leading to conduction loss reduction and system efficiency increase [37].

It should be noted that the proposed control method is also applicable to the conventional boost converter. On the contrary, the realization of fully ZVS occurs only when the converter diode is replaced with a synchronous switch. In order to have an integrated control strategy, both BCM control and fully ZVS realization have been applied on the synchronous boost converter.

Synchronous boost converter along with the proposed BCM control technique is shown in Figure 1.

Firstly, the boost converter is designed in BCM. Then, applying the boundary condition, inductor size is obtained via Equation (1):

$$ L_{BCM} = \frac{0.5D(1-D)R}{f_{sw}} $$

(1)

where D is the main switch duty cycle, R is the load and f_{sw} is the switching frequency. In this control method, the switching frequency varies in a pre-defined range. This range, by which the circuit can continue to operate in BCM, must first be calculated. To this end, the relationship between the input voltage, output power and output voltage variations are shown in Equation (2).

$$ L_{BCM} = \frac{2f_{ult}(V_{in} - V_{out})}{V_{out}^2} = \frac{2f_{ult}}{f_{sw}} $$

(2)

According to Equation (2), if load current increases at the presence of the constant input and output voltages, the converter performance mode will change from BCM to CCM. The frequency variation range in which the circuit can be maintained in BCM is dependent on the input voltage, output voltage and power variations.

Figure 2 shows the frequency variations versus the input voltage and output power according to Equation (2). As can be seen from the figure, by a proper control of the switching frequency, converter can always be kept in BCM despite of the load and input voltage variations. As the output current increases, the converter performance changes from BCM to CCM.

As shown in Figure 3, the inductor current has the same mean value in both BCM and CCM operation modes. If the converter is shifted to CCM, the control strategy can return it to BCM by reducing the frequency.

Figure 1. The boost converter and the proposed control block diagram

Figure 2. Frequency variation range versus input voltage and output power
In the proposed method, after inductor current sampling, an offset with a value close to zero is compared with this current. Comparator output is shown as $V_{\text{comp}}$ in Equation (3). According to this equation, if inductor current is lower than offset value, the comparator output would be $+5$V.

$$V_{\text{comp}} = \begin{cases} +5 & I_L < \text{offset} \\ 0 & I_L > \text{offset} \end{cases}$$

where $V_{\text{comp}}$ is a square wave. By applying this waveform to control on average block, $V_{\text{comp}}$ average signal ($V_{\text{control}}$) is computed according to Equation (4). Then, according to Equation (5), $V_{\text{control}}$ is applied to regulate $I_{\text{control}}$. This current charges $C_f$ and plays a crucial role in determining switching frequency of the converter.

$$V_{\text{control}} = V_c - \frac{R_F V_{\text{comp}} \text{dt}}{\tau}$$

$$I_{\text{control}} = \frac{V_c V_{\text{comp}}}{R_{\text{control}}}$$

If $I_{\text{control}}$ is zero, there would be no charging source for $C_f$, and production of the switching frequency required for circuit performance would be disrupted. In this case, an auxiliary current source should be used to generate $I_{\text{min}}$ for $C_f$ charging and minimum required frequency. In general, required current for $C_f$ charging is calculated according to Equation (6).

$$I_f = I_{\text{control}} + I_{\text{min}}$$

$$I_{\text{min}} = \frac{V_c V_{\text{comp}}}{R_{\text{min}}}$$

To discharge the capacitor, its voltage should be compared with $V_c$ reference value. When the capacitor voltage reaches to $V_c$, the comparator output connected to the control switch gate is set to logic state 1. Therefore, the intended switch is turned on and the capacitor will be short-circuited. Hence, the capacitor begins to discharge. By continuous charging and discharging of the capacitor, a pulse is produced at the comparator output, the frequency of which is according to Equation (8).

$$f_{\text{sw}} = \frac{f_f}{V_c}$$

Therefore, by applying the proposed control method, switching frequency could be regulated for circuit performance in BCM. Peak current mode method is used to stabilize the output voltage in this converter. The output pulse generated by the capacitor charging and discharging (which determines switching frequency of the converter), is inserted into a RS flip-flop set port. On the other hand, reset pulse of the intended flip-flop is applied by PI controller responsible for setting the output voltage error to zero. The flip-flop output and its complement, generate commands for $S_1$ and $S_2$ switches, respectively. In this paper, by a slight change in control structure, the condition is provided for circuit performance in fully ZVS, as discussed in the following.

### 3. OPERATION ANALYSIS OF BOOST CONVERTER

Inductor current in a switching interval of synchronous boost converter working in BCM is shown in Figure 4 [38].

It should be noted that given the input and output ranges and converter control method, each of these three states in Figure 4 could occur in this converter. If complete control is not applied to the converter operating in BCM, depending on the input and output ratio, modes (a) and (b) can be possible; such that in the case of $V_{\text{in}} < V_{\text{out}} / 2$, (a) and for $V_{\text{in}} > V_{\text{out}} / 2$, (b) is possible. Likewise, by appropriate control, fully ZVS can be achieved according to (c). Each switching interval includes two major times ($T_{\text{ON}}$ and $T_{\text{OFF}}$) and two resonant times in all three above-mentioned states. However, all three states are similar in the first two subintervals and different in third and fourth subintervals. In the following, the first two subintervals are examined first, then three states are explained, separately.

Primary current of the inductor ($I_L$) and primary voltage of the capacitor of the main switch ($S_1$) at the beginning of each circuit switching interval is assumed as Equation (9):

$$I_L(t_0) = 0, \quad V_{DS1}(t_0) = 0$$

Each switching interval of this converter is divided into four subintervals, each of which is discussed below.

#### Subinterval 1 [$t_0 < t < t_1$]: $S_1$ is turned on during this interval and inductor current increases linearly with the slope of $\frac{di}{dt} = \frac{V_{\text{in}}}{L}$. This subinterval is called turn-on time ($T_{\text{ON}}$), the length of which is determined by control loop of the output voltage regulation. $S_1$ is turned off at $t_1$.
The ultimate value of $I_L$ and $V_{DS1}$ and length of this subinterval are as below:

$$T_{ON}=t_1-t_0, \quad I_L(t_1)=I_p, \quad V_{DS1}(t_1)=0$$  \hspace{1cm} (10)

**Subinterval 2 $[t_1<t_2]$**: During this subinterval, both $S_1$ and $S_2$ are off, and a resonance occurs between the non-linear capacitors of $S_1$ and $S_2$ ($C_{OSSI}, C_{OSS2}$) and the inductor; therefore, $C_{OSSI}, C_{OSS2}$ will be discharged. The length of this subinterval is short, and high inductor current is with the amount of $I_p$. Hence, this subinterval can be neglected for the following calculations.

The ultimate value of $I_L$ and $V_{DS1}$ and length of this subinterval are as below:

$$T_{R1}=t_2-t_1, \quad I_L(t_2)=I_p, \quad V_{DS1}(t_2)=V_{out}$$  \hspace{1cm} (11)

**Subinterval 3 $[t_2<t_3]$**: When $C_{OSS2}$ is fully discharged, diode of $S_2$ switch begins to flow inductor current. In fact, after a short delay ($T_{R1}$), $S_2$ switch is turned on at zero voltage. At this interval, voltage applied to the inductor is negative, thus $I_L$ is reduced. Due to the fact that $S_1$ is off at this interval, this is the off-time ($T_{off}$).

The ultimate value of $I_L$ and $V_{DS1}$ and length of this subinterval are as below:

$$T_{off}=t_3-t_2, \quad I_L(t_3)=0, \quad V_{DS1}(t_2)=V_{out}$$  \hspace{1cm} (12)

**Subinterval 4 $[t_3<t_4]$**: When $I_L$ reaches to zero, switch $S_1$ is turned off and the equivalent circuit is like Figure 6; therefore, a resonance occurs between the capacitor and inductor and an oscillation with the frequency of $\omega_0 = \frac{1}{\sqrt{L(C_{OSSI}+C_{OSS2})}}$ begins, the behavior of which depends on the ratio of input and output voltage, i.e. $\frac{V_{in}}{V_{out}}$.

3. 1. **Natural Switching Operation** ($V_{in} < \frac{V_{out}}{2}$)

When input voltage is less than $V_{out}/2$, according to Figure 5(a), $V_{DS1}$ will be zero, and $S_1$ is turned on at zero voltage and negative current.

3. 2. **Valley Switching Operation** ($V_{in} > \frac{V_{out}}{2}$)

When input voltage is greater than $V_{out}/2$, according to Figure 5(b), $V_{DS1}$ will be zero at resonance interval of $T_{R2}$, because inductor does not have enough energy to discharge capacitor and $S_2$ turns on at a small voltage of $V_{valley}$. In this case, switching loss is reduced, but fully ZVS does not occur.

The ultimate value of $I_L$ and $V_{DS1}$ and length of this subinterval are as below:

$$T_{off}=t_4-t_3, \quad I_L(t_4)=0, \quad V_{DS1}(t_4)=V_{valley}$$  \hspace{1cm} (13)

To solve this issue and achieve ZVS, $S_2$ should be stayed on for a longer period of time. As it can be observed in Figure 5(c), during $T_{R2}$, inductor current should reach to an appropriate negative current so that $C_{OSSI}$ is fully discharged and $S_1$ turned on both at zero voltage and current.

Figure 5. Current and voltage waveforms: (a) BCM natural switching operating waveforms when $V_{in} \leq 0.5V_{out}$, (b) BCM valley switching operating waveforms when $V_{in} > 0.5V_{out}$, (c) Operation waveforms based on the proposed extended ZVS control.
4. PROPOSED ANALOG ZVS CONTROL STRATEGY

In this section, the required current (Ie) to achieve fully ZVS is calculated. Since this current depends on Coss switch capacitor and this capacitor is non-linear, according to the datasheet [39], the equation between Coss and VDS in [0-Vout] is estimated as Equation (14):

\[
c_{\text{oss}} = aV_D^3 + bV_D^2 + cV_D + d
\]

(14)

The coefficient of a, b, c, and d are calculated using cftool software.

To achieve fully ZVS, Ie should be reduced to Ie. After reaching to Ie according to Figure 5(c), Si is turned off and Coss1 begins to charge and Coss2 begins to discharge. In this case, according to Figure 4, Coss1 and Coss2 are in parallel (\(C = C_{\text{oss1}}|C_{\text{oss2}}\)), and the equivalent circuit is according to Figure 6. Capacitor C is assumed linear and its average value is estimated as Equation (15).

\[
\begin{align*}
Q_R &= \frac{1}{v_{\text{out}}}C_{\text{ave}}d\psi_\text{DS} \\
C &= \frac{Q_r}{v_{\text{out}}}
\end{align*}
\]

(15)

According to Figure 7, inductor current will be as follows:

\[
\begin{align*}
I_L(t) &= A\cos\omega_0 t + B\sin\omega_0 t + CV_{in} \\
A &= I_e - CV_{in} \\
B &= \frac{1}{L}(V_{in} - V_{out}) \\
\omega_0 &= \frac{1}{\sqrt{L}}
\end{align*}
\]

(16)

![Figure 6. Equivalent resonant circuit](image)

![Figure 7. Detailed inductor current](image)

To calculate Ie current, Equation (17) is used in diagram of Figure 7:

\[
Q_R = \int_{t_m}^{t_1} I_e dt = \int_{t_2}^{t_3} I_e dt
\]

(17)

In \([t_s - t_m]\), QR is shown as Equation (18):

\[
Q_R = \int_{t_m}^{t_1} \frac{\Delta}{\omega_0} \sin\omega_0 t_m - \frac{\pi}{\omega_0} \cos\omega_0 t_m + CV_m t_m + \frac{\pi}{\omega_0}
\]

(18)

And in \([t_m - t_f]\):

\[
Q_R = \int_{t_m}^{0} \frac{\Delta}{\omega_0} (\sin\omega_0 t_f - \sin\omega_0 t_m)
\]

\[
- \frac{\pi}{\omega_0} (\cos\omega_0 t_f - \cos\omega_0 t_m) + CV_m (t_f - t_m)
\]

(19)

\(T_m\) and \(t_f\) values are calculated according to Equations (20) and (21):

\[
t_m = \frac{L}{V_{\text{in}}} \omega_0 V_{out}
\]

(20)

\[
t_f = \frac{1}{V_{\text{in}}} \int_{t_1}^{t_2} V_{\text{out}} dt - \frac{1}{V_{\text{in}}} V_{out}
\]

(21)

Using Equations (18)-(21) and Taylor extension, \(Q_\text{w}\) will be close to zero, \(I_e\) and \(I_{\text{in}}\) (\(I_{\text{em}}\) is the negative maximum of \(I_e\)) will be according to Equations (22) and (23):

\[
I_e = \frac{k_1}{L} (1 - Lk_1/2V_{out})^2 - 4(1 - k_1(V_{out}V_{in}))(k_2(V_{out} + V_{in}))
\]

\[
k_1 = \frac{Q}{L} (V_{in}V_{out})
\]

\[
k_2 = \frac{1}{2V_{in}}
\]

\[
I_{\text{em}} = \frac{k_1 + k_2^2}{I_e}
\]

(22)

(23)

Since \(T_{sw} = T_{\text{on}} + T_{\text{off}} + T_e + T_R\), each subinterval is obtained according to Equation (24):

\[
T_{\text{on}} = \frac{I_{\text{in}}}{V_{\text{in}}} V_{out}
\]

\[
T_{\text{off}} = \frac{I_{\text{in}}}{V_{\text{in}}} V_{out}
\]

\[
T_e = \frac{1}{V_{\text{in}}} \int_{t_1}^{t_2} V_{\text{out}} dt - \frac{1}{V_{\text{in}}} V_{out}
\]

\[
T_R = \frac{1}{V_{\text{in}}} (I_{\text{in}} - I_{\text{out}})
\]

(24)

As stated in Section 2, the proposed closed-loop control system guarantees converter performance in BCM. Given the attained equations, fully ZVS can be obtained by a slight change in this control system. To this end, pulse commands going to S1 and S2 should be changed according to Figure 8. In addition, offset value computed by Ie should be replaced by Equation (22). Therefore, control loop of inductor current, in this section, determines Ie instead of zero-crossing detection. When Ie becomes equal with Ie, controlled pulse of the switching frequency goes to the R port of flip-flop2 and commands S2 to turn off. Then, this pulse is delayed as long as \(T_R\) time span, calculated in Equation (24). \(C_{\text{oss1}}\) is fully discharged during this time, then \(S_2\) is commanded to be turned on. It should be noted that, as before, \(S_1\) turn-off...
command is issued by control loop of voltage stabilizer. To reduce the loss more than ever, turn-on command of $S_2$ is issued after reaching $S_1$ current to zero. Given the above, it is clear that the proposed control method guarantees both ZVS and ZCS.

5. SIMULATION RESULTS

In order to validate the accuracy of the circuit performance, the proposed method is applied to a synchronous boost converter and implemented in the LTspice software. Also, the values of the circuit parameters are displayed in Table 1.

Boost converter simulation is conducted by the proposed control technique in two switching frequency ranges. At first, the intended converter is simulated at 100KHz to prove the accurate performance in BCM despite load changes; then, at 1MHz switching frequency to prove that control system can achieve fully ZVS.

The steady state inductor current in BCM is shown in Figure 9(a). As can be seen from the figure, under nominal frequency and rated load, the proposed control algorithm maintains the inductor current in BCM. Figure 9(b) shows the output voltage. As is depicted in the figure, by using the peak current mode control method, the output voltage of the circuit conforms well to its reference value. This voltage reaches to its ultimate value after 1.5ms. In this case, the load resistance and the switching frequency values are equal to 25Ω and 100 kHz respectively.

Figure 10 shows the converter output voltage at the presence of increasing the load current up to 20%. As it can be seen in Figure 10, the output voltage reaches to its reference value after 0.5ms. Moreover, due to the increase in the load current, the converter mode has changed from BCM to CCM. However, the controller performs in such a manner that the inductor current returns to BCM shortly after the load changes. The load resistance and the switching frequency values are equal to 20Ω and 80KHz respectively. It means that the controller has decreased the switching frequency in order to yield the BCM condition.

![Figure 8](image_url)  
**Figure 8.** The simplified control diagram of the proposed system control strategy

![Figure 9](image_url)  
**Figure 9.** Simulation results: (a) Inductor current, (b) Output voltage

![Figure 10](image_url)  
**Figure 10.** Output voltage at CCM and BCM regions

### Table 1. Simulation parameters

| Parameters/Devices | Symbol | Value/Number |
|--------------------|--------|--------------|
| Inductor           | $L$    | $17\mu H(f_{sw}=100\text{KHz})$ |
|                    |        | $1.7\mu H(f_{sw}=1\text{MHz})$ |
| Output Capacitor   | $C$    | $17.2\mu F(f_{sw}=100\text{KHz})$ |
|                    |        | $17.2\mu F(f_{sw}=1\text{MHz})$ |
| Switch             | $S$    | GS66508T     |
| Input Voltage      | $V_{in}$ | 17V          |
| Output Voltage     | $V_{out}$ | 30V          |
| Load               | $R_{Load}$ | 25Ω          |
| Output Power       | $P_{out}$ | 36W          |
Figure 11 illustrates the output pulse from the comparator. According to this figure, inductor current returns from CCM to BCM after 0.5ms. As noted before, this pulse will then be entered to the averaged block.

Figure 12 shows inductor current during load current variations. As is depicted in the figure, inductor current is primarily in CCM mode. In this case, controller begins to operate and reduces the switching frequency.

Also, in Figure 13, by 20% reduction of load current, first inductor current flows from BCM to DCM; then in a short period of time (about 0.4ms), the controller...
increases switching frequency with an accurate detection of the converter performance, and converter is returned to BCM. The output voltage, in these variations, has reached to reference value in less than 0.6ms, shown in Figure 14.

In the following, simulation results of the intended converter will be examined using the proposed control method to achieve fully ZVS in 1MHz switching frequency.

Extended closed-loop control with \( V_{\text{IN}}, V_{\text{OUT}}, V_{\text{DS1}}, \) and \( I_t \) in \( V_{\text{out}}=30V, V_{\text{in}}=17V, f_s=1\text{MHz} \) to achieve fully ZVS is shown in Figure 15.

Pulse commands applied to \( S_1 \) and \( S_2 \) are shown in Figure 16. Turn-on command of \( S_1 \) is issued in ZVS and ZCS. \( S_1 \) state is “ON” during \( T_{\text{on}} \). \( S_2 \) turn-on command will be issued when \( S_1 \) current reaches to zero. When \( I_t \) is equal to \( I_r \) (calculated in Section 4), \( S_2 \) will be turned off. Also seen in the related figure, \( S_1 \) turn-on command is delayed as much as \( T_{\text{R}} \) time. \( V_{\text{DS1}} \) and \( V_{\text{DS2}} \) (Figure 17) proves the accurate performance of the proposed control method.

6. CONCLUSION

A new variable frequency ZVS control method for boost converter operating in BCM has been proposed in this paper. This technique is carried out in analog form without using microprocessor as analog controllers have less noise, less cost and processing problems in high frequency applications. The intended method keeps the converter in BCM in spite of load and input voltage variations by changing switching frequency in a certain specified range. In addition, with a slight change in control structure, it’s possible to achieve a fully ZVS in both \( V_{\text{in}} < V_{\text{out}}/2 \) and \( V_{\text{in}} > V_{\text{out}}/2 \) cases. Simulation results obtained from applying the proposed method on a GaN-based synchronous boost converter in two different switching frequency ranges (100KHz and 1MHz) verify the proposed strategy advantages.

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چکیده
در این مقاله یک روش کنترل فرکانس متغیر از کلیدزنی در وانر صفر برای مبدل بوستی که در مد هدایت مرزی کار می‌کند پیشنهاد گردید. روشهای مذکور قادر است انرژی تعیین نامبره در کاهش مصرف برق کاربردی در کاربردهای فرکانس بالا است. نتایج شبیه‌سازی نشان داده که تغییرات آماده در یک محدوده محدود می‌باشد و با تغییرات بار و ولتاژ ورودی، مبدل را از طریق تغییر فرکانس کلیدزنی در یک محدوده معین، همواره در مد هدایت مرزی نگاه دارد.

کلمات کلیدی: کنترل فرکانس، کلیدزنی در صفر جریان، تغییر فرکانس کلیدزنی، مد هدایت مرزی.