A Low Cost Single-Switch Bridgeless Boost PFC Converter

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ABSTRACT
This paper proposes the single-switch bridgeless boost power factor correction (PFC) converter to achieve high efficiency in low cost. The proposed converter utilizes only one active switching device for PFC operation as well as expecting higher efficiency than typical boost PFC converters. On the other hand, the implementation cost is less than traditional bridgeless boost PFC converters, in where two active switching devices are necessary. The operational principle, the modeling, and the control scheme of the proposed converter are discussed in detail. In order to verify the operation of the proposed converter, a 500W switching model is built in PSIM software package. The simulation results show that the proposed converter perfectly achieves PFC operation with only a single active switch.

IC. INTRODUCTION
A power factor correction (PFC) converters have been employed in many applications such as power supplies, battery chargers, motor drive applications, and so on [1]-[9]. Traditional PFC converters usually employs a diode bridge and an active switching device such as IGBTs and MOSFETs. Recently, bridgeless PFC converters which do not require having a diode bridge have been studied, because the converter efficiency can be improved by eliminating the diode bridge. However, typical bridgeless PFC converters require two active switches to conduct an input current according to the polarity of the input voltage. This increases the implementation cost including the gate driver circuitries and snubbers. Moreover, increasing the number of active switching devices also decreases the reliability of the entire power stage. In terms of electromagnetic interference (EMI), it has been known that bridgeless topologies are worse than traditional boost PFC topologies with a diode bridge. In order to overcome this disadvantage, semi-bridgeless PFC converters have been proposed [10], [11]. These semi-bridgeless PFC converters have equivalent EMI characteristics with traditional boost PFC converter with a diode bridge, but still they employ two active switches.

In this paper, a low cost single-switch bridgeless PFC converter is proposed. The proposed converter utilizes a single active switch, and it operates in entire electrical cycle. So, the implementation cost can be reduced, and the utilization of the switch can be increased. Compared to the traditional semi-boost PFC converter, the proposed converter employs two more diodes to avoid a short circuit condition, but reduce the number of the active switch whose realization cost is higher than several passive switching components such as a diode. So the total cost saving can be achieved. The EMI characteristics of the proposed converter are basically identical to traditional semi-bridgeless PFC topologies. This paper consists of following sections. In section 2, the single-switch bridgeless PFC topology is proposed, and its operation mode and the inductor current equation are analyzed in detail. In section 3, the control model of the proposed converter is discussed.
and its derivation procedure is explained. The control strategy is also introduced in this section. The numerical transfer function and the simulation model of an example case are discussed in section 4. Also the loop-gain analysis with the designed controllers are performed. The simulation results are also shown in the section to verify the performance of the proposed converter. Finally, the conclusion is made in the last section.

2. PROPOSED SINGLE-SWITCH BRIDGELESS BOOST PFC CONVERTER

Figure 1 shows the topology of the proposed single-switch bridgeless boost PFC converter. For the front-end stage, only one active switching device $Q_1$ is employed, and it conducts in full electrical cycle. Since $Q_1$ operates in the entire cycle, the blocking diodes $D_3$ and $D_4$ are necessary to avoid confliction of positive and negative half cycles.

![Proposed single-switch bridgeless boost PFC converter](image)

Figure 1. Proposed single-switch bridgeless boost PFC converter

Figure 2 illustrates the operation modes of the proposed converter according to the polarity of the input voltage $v_g$ and the status of $Q_1$. In figure 2(a), $v_g$ is positive, and $Q_1$ is turned on. In this case, the input current $i_g$ flows through $L_1$, $D_3$, $Q_1$, and $D_6$, and $D_1$, $D_2$, $D_4$, $D_5$ are blocked. The input energy is stored in $L_1$ while the load $R_o$ is supplied from the energy charged in the dc-link capacitor $C_o$. By ignoring the voltage drops induced by $D_3$, $D_6$, and $Q_1$, the voltage across $L_1$ is written as follows:
For the switch off-stage in positive $v_g$, the equivalent circuit is shown in Figure 2(b). Here, $i_g$ flows via $L_1$, $D_1$, $C_o$, $R_o$, and $D_6$. The active switch $Q_1$ and other diodes do not conduct. In this stage, the energies stored in $L_1$ and from the source $v_g$ are simultaneously transferred to the dc-link and the load. This operation is exactly same to typical boost converters. As similar to the previous one, the inductor voltage is represented as (2).

$$v_{L1} = v_g - V_o = L_1 \frac{di_{L1}}{dt}$$

(2)

Note that the inductor current $i_{L1}$ is the same to $i_{g}$ for the positive half cycle. For the negative half cycle, $D_4$ or $D_2$ is turned on according to the status of $Q_1$. Unlike the previous two cases, $D_5$ and $L_2$ are conducting as in Figures 2(c) and 2(d). In these modes, the magnitude of $i_{L2}$ is identical to $i_{g}$, but its direction is opposite according to the definition in Figure 2. The expressions of $i_{L2}$ are represented in (3) and (4).

$$v_{L2} = -v_g = L_2 \frac{di_{L2}}{dt}$$

(3)

$$v_{L2} = -v_g - V_o = L_2 \frac{di_{L2}}{dt}$$

(4)

For the capacitor voltage, only the status of $Q_1$ is considered without referring the polarity of the input voltage. When $Q_1$ is turned on, the capacitor current is written as (5).

$$i_c = C_o \frac{dv}{dt} = -I_o = -\frac{v}{R}$$

(5)

The capacitor current in the off stage of $Q_1$ is represented as below:

$$i_c = C_o \frac{dv}{dt} = (i_{L1} \text{ or } i_{L2}) - I_o = i_{g} - \frac{v}{R}$$

(6)

3. MODELING AND CONTROL OF THE PROPOSED CONVERTER

3.1. Modeling of the Duty-to-input Current

For PFC converter control, usually two control loops, the input current and the dc-link voltage, are necessary. For the input current control loop design, the duty-to-inductor current model should be evaluated. In order to simplify the model, let’s assume that $L_1$ and $L_2$ have the same values as $L_g$. Then, Equation (1) and (3) are the same as well as Equations (2) and (4) are identical. Then, the well-known state-space averaging technique can be applied to obtain the control models [12]. By using Equations (1) and (5), the state equation when $Q_1$ is turned on can be written as follows:

$$\begin{bmatrix}
\frac{di_g}{dt} \\
\frac{dv_c}{dt}
\end{bmatrix} = A_0 \begin{bmatrix}
i_g \\
v_g
\end{bmatrix} + B_0 v_g, \\
A_0 = \begin{bmatrix}
0 & 0 \\
0 & -\frac{1}{R_o C_o}
\end{bmatrix}, \\
B_0 = \begin{bmatrix}
\frac{1}{L_g} \\
0
\end{bmatrix}$$

(7)

On the other hand, the state equation when $Q_1$ is turned off is also derived as (8) by using (2) and (6).

$$\begin{bmatrix}
\frac{di_g}{dt} \\
\frac{dv_c}{dt}
\end{bmatrix} = A_i \begin{bmatrix}
i_g \\
v_g
\end{bmatrix} + B_i v_g, \\
A_i = \begin{bmatrix}
0 & -\frac{1}{R_o} \\
\frac{1}{C_o} & -\frac{1}{R_o C_o}
\end{bmatrix}, \\
B_i = \begin{bmatrix}
\frac{1}{L_g} \\
0
\end{bmatrix}$$

(8)
Let’s define constant matrices $A$ and $B$ for state-space analysis as:

$$A = A_0 d + A_1 (1-d), \quad B = B_0 d + B_1 (1-d)$$  \hspace{1cm} (9)

Where $d$ represents the duty reference. After that the dc components of the states $i_g$ and $v_c$ can be derived as follows by utilizing the matrices $A$ and $B$.

$$X = -A^{-1}BV_g$$  \hspace{1cm} (10)

Where $X$ is the dc component vector, and $V_g$ is the peak value of $v_g$. By applying Laplace transform, (11) is obtained as follows:

$$\begin{bmatrix} i_g(s) \\ d(s) \\ v_c(s) \\ d(s) \end{bmatrix} = \left(sI - A\right)^{-1}\left[\left(A_0 - A_1\right)X + \left(B_0 - B_1\right)V_g\right], \quad I = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$$  \hspace{1cm} (11)

By solving (11), the duty-to-inductor current and the duty-to-capacitor voltage models are derived as follows:

$$\frac{i_g(s)}{d(s)} = \frac{V_g \left(R_C s + 2\right)}{L \left(1-d\right) \left(R_C s^2 + s + \left(1-d\right)^2\right)}$$  \hspace{1cm} (12)

$$\frac{v_c(s)}{d(s)} = \frac{RV_g \left(1-d\right)^2 - s}{L \left(1-d\right)^2 \left(R_C s^2 + s + \left(1-d\right)^2\right)}$$  \hspace{1cm} (13)

### 3.2. Control Strategy

As described before, the voltage and the current control loops as in Figure 3 are necessary for the PFC converter. Two proportional-integral (PI) controllers are employed for each control loop. In order to improve the current control performance, the duty feed-forward term is applied. Thanks to the feed-forward term, the integral portion in the PI current controller is reduced, so that the dynamic property can be improved. For the voltage controller, a 120Hz bandstop filter is employed to filter out 120Hz periodic voltage ripple caused by the single-phase power fluctuation phenomenon [13]. By doing so, the dynamic property of the voltage control loop can be improved without introducing unnecessary 120Hz component in the current reference.

![Figure 3. The control scheme of the proposed converter](image)

### 4. SIMULATION STUDY

#### 4.1. Power Circuit Switching Model

Figure 4 shows the developed switching model in PSIM. In the switching model, both the 15 percent and 85 percent of the rated load are configured. In order to see the dynamic performance of the entire control system, the 85 percent load can be connected or disconnected in step. The parameters of the power circuit is summarized in Table 1. All active and passive switching devices are assumed as ideal elements, so their voltage drops are ignored. The time step for the simulation is selected as 250nsec.
Table 1. The Parameters for the Simulation

| Contents                        | Values          |
|---------------------------------|-----------------|
| Input inductance $L_g$          | 1 mH            |
| Output capacitance $C_o$        | 330 μF          |
| Input root-mean-square (RMS) voltage $V_{rms}$ | 220 V          |
| Rated output power $P_o$        | 500 W           |
| Output voltage reference        | 400 V           |
| Operating frequency             | 60 Hz           |
| Switching frequency             | 200 kHz         |

4.2. Controller Design and Implementation

For the current controller design, the transfer function in (12) is utilized. By substituting the parameters into the transfer function, the numerical model in (14) is obtained.

\[
\frac{i_c(s)}{d(s)} = \frac{27.38s + 622.3}{6.842 \times 10^{-5}s^2 + 0.0007775s + 0.00047}
\]  

(14)

In (14), the duty reference $d$ was selected as 0.2225 which corresponds to the required duty reference to produce 400V output at the peak of the input voltage. By using the MATLAB SISOTOOL, the proportional and the integral gains of the PI current controller are designed as 0.1556 and 2103, respectively. These values give 78 deg of phase margin at 10kHz, and it may be an enough controller design specification for the PFC current control.

In order to design the voltage controller, the plant model is derived as (15) rather than using (13), because of the consideration of the current control loop.

\[
\frac{v_c(s)}{i_c(s)} = \frac{R_c}{R SC_p s + 1}
\]  

(15)

The 120Hz bandstop filter is implemented as follows:

\[
G_{bs}(s) = \frac{s^2 + (2\pi \times f_c)^2}{s^2 + 2\pi \times f_ps + (2\pi \times f_c)^2}
\]  

(16)

Where $f_c$ and $f_p$ represent the cut-off frequency and the passband of the bandstop filter. In the simulation, $f_c$ and $f_p$ are selected as 120Hz and 10Hz. Again, the MATLAB SISOTOOL is utilized to determine the
proportional and the integral gains of the PI voltage controller as 0.1 and 5. With the gains, the entire voltage loop has 80.6 deg of phase margin at 72.5Hz. Figure 5 shows the open-loop gain of the current and the voltage control loops. From the figure, it is confirmed that the open-loop gain results with the designed controllers satisfy the design specifications.

(a) the duty-to-inductor current model  
(b) the duty-to-capacitor voltage model

Figure 5. The open-loop gains

(a) Input current  
(b) Output voltage

Figure 7. The simulation result at the full load steady-state condition

4.3. Simulation Results

The simulation result using the developed switching model at the full load steady-state condition is shown in Figure 7. As shown in the figure, the input current is regulated sinusoidally, and the output voltage is controlled to 400V. In the output voltage, the well-known double frequency ripple, here 120Hz, appears.
Figure 8 shows the simulated waveforms of the devices for positive and negative half cycles. The waveforms correspond the analysis taken in the previous section. Note that the switch current $i_{Q1}$ flows both in the positive and the negative cycles whereas other devices conducts in each half cycle as analyzed before.

(a) For positive half cycle
(b) For negative halfcycle

Figure 8. The waveforms of devices

Figure 9. The transient response of the developed switching simulation model
The transient response of the developed simulation model is shown in Figure 9. In the simulation, the load is changed at \( t = 0.2 \) s from 15 percent to 100 percent in step. As shown in the figure, the current and the voltage controls are stable, and the current is very well regulated. At \( t = 0.4 \) s, the load is changed to 15 percent. Even in that condition, there is no significant transient problem, and the overshoot of the dc-link voltage is less than 20V. At 15 percent load, the current is slightly distorted, because the converter operates in discontinuous conduction mode in such a small output power [14]-[15]. However, the effect of the current distortion is so little, that it is not a big problem in the entire system performance.

In sum, through the simulations, the operation of the single-switch bridgeless PFC converter has been verified.

5. CONCLUSION

The single-switch bridgeless PFC converter has been proposed in this paper. For the proposed converter, the topology, the operation modes, the modeling and controls have been dealt in detail. The numerical transfer function of the proposed converter has been evaluated using the state-space averaging technique, and the simulation model matched with the numerical function have been tested. In order to see the performance of the proposed converter, a 500W switching model was built in PSIM software package. By using the developed simulation model, both the transient and the steady-state operations have been evaluated. Through the simulations, it has been confirmed that the operation of the proposed converter is very well matching the theoretical analyses. Since the proposed converter employs only one single active switch, it is expected that the implementation cost can be reduced, and the reliability of the entire power stage can be improved.

REFERENCES

[1] L Huber, Y Jang, MM Jovanovic. Performance Evaluation of Bridgeless PFC Boost Rectifiers. IEEE Transactions on Power Electronics. 2011; 23: 1381-1390.
[2] F Musavi, W Eberle, WG Dunford. A High-Performance Single-Phase Bridgeless Interleaved PFC Converter for Plug-in Hybrid Electric Vehicle Battery Chargers. IEEE Transactions on Industry Applications. 2011; 47: 1833-1843.
[3] Y Cho, H Mok, JS Lai. Analysis of the Admittance Component for Digitally Controlled Single-Phase Bridgeless PFC Converter. Journal of Power Electronics. 2013; 13: 600-608.
[4] MKH Cheung, MHL Chow, YM Lai, KH Loo. Effect of Imperfect Sinusoidal Input Currents on the Performance of a Boost PFC Pre-Regulator. Journal of Power Electronics. 2012; 12: 689-698.
[5] GG Park, KY Kwon, TW Kim. PFC Dual Boost Converter Based on Input Voltage Estimation for DC Inverter Air Conditioner. Journal of Power Electronics. 2010; 10: 293-299.
[6] P Das, M Pahlevaninezhad, J Drobnik, G Moschopoulos, PK Jain. A Nonlinear Controller Based on a Discrete Energy Function for an AC/DC Boost PFC Converter. IEEE Transactions on Power Electronics. 2013; 28: 5458-5476.
[7] YK Lo, CY Lin, HJ Chiu, SJ Cheng, JY Lin. Analysis and Design of a Push-Pull Quasi-Resonant Boost Power Factor Corrector. IEEE Transactions on Power Electronics. 2013; 28.
[8] SA Khan, Md I Hossain, M Aktar. Single-Phase PFC Converter for Plug-in Hybrid Electric Vehicle Battery Chargers. International Journal of Power Electronics and Drive Systems. 2012; 2.
[9] D Lenine, ChS Babu, G Shankaiaish. Performance Evaluation of Fuzzy and PI Controller for Boost Converter with Active PFC. International Journal of Power Electronics and Drive Systems. 2012; 2.
[10] P Kong, S Wang, FC Lee. Common Mode EMI Noise Suppression for Bridgeless PFC Converters. IEEE Transactions on Power Electronics. 2008; 23: 291-297.
[11] F Musavi, W Eberle, WG Dunford. A Phase-Shifted Gating Technique With Simplified Current Sensing for the Semi-Bridgeless AC-DC Converter. IEEE Transactions on Vehicular Technology. 2013; 62: 1568-1576.
[12] RW Erickson, D Maksimovic. Fundamentals of Power Electronics. Norwell, MA: Kluwer Academic. 2002: 213-226.
[13] S Buso, P Mattavelli, L Rossetto, G Spiazzi. Simple Digital Control Improving Dynamic Performance of Power Factor Preregulators. IEEE Transactions on Power Electronics. 1998; 13: 814-823.
[14] L Huber, L Gang, MM Jovanovic. Design-Oriented Analysis and Performance Evaluation of Buck PFC Front End. IEEE Transactions on Power Electronics. 2010; 25: 85-94.
[15] Y Cho, JS Lai. Digital Plug-in Repetitive Controller for Single-Phase Bridgeless PFC Converters. IEEE Transactions on Power Electronics. 2013; 28: 165-175.
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Younghoon Cho was born in Seoul, Korea, in 1980. He received the B.S. degree in electrical engineering from Konkuk University, Seoul, in 2002, the M.S. degree in electrical engineering from Seoul National University, Seoul, in 2004, and the Ph.D. degree from Virginia Polytechnic Institute and State University, Blacksburg, VA, USA, in 2012. From 2004 to 2009, he was an Assistant Research Engineer with the Hyundai MOBIS R&D Center, Yongin, Korea. Since 2013, he is now an Assistant Professor in the Department of Electrical Engineering, Konkuk University, in Seoul. His current research interests include digital control techniques for power electronic converters in vehicle and grid applications, multilevel converters, and high-performance motor drives.