Improve power density with a baby boost converter in a PFC circuit

Sheng-Yang Yu
Systems Applications Engineer

Benjamin Genereaux
Systems Applications Engineer

LiehChung Yin
Systems Applications Engineer

Introduction

In the year 2000, server front end power-supply units (PSUs) – with an AC input to a 12-V/48-V DC rail – achieved around 10 W/in3 of power density, with about 85% peak efficiency [1]. Today, so many server PSUs can meet 80 Plus Platinum (94% peak) and 80 Plus Titanium (96% peak) requirements [2] that the latter requirement, along with very high power density (>90 W/in^3), are both now becoming minimum requirements.

One reason why server PSUs can achieve higher power density levels is because of technological innovations in the semiconductor industry. New semiconductor fabrication processes enable devices to have lower parasitics and better figures of merit [3], significantly improving power dissipation and facilitating higher power density.

Topology and architecture innovations are also behind PSUs with high power density. Applying a totem-pole bridgeless power factor correction (PFC) circuit to the AC/DC rectifier stage of new server PSUs – along with wide bandgap devices such as gallium nitride (GaN) and silicon carbide (SiC) (**Figure 1**) – achieves the best converter efficiency over other bridge or bridgeless PFC topologies [4] [5]. Although higher efficiency does minimize the area needed for heat dissipation, a bulk capacitor (C_{BULK} in **Figure 1**) is still required to hold the output voltage in regulation after AC dropout. In order to accomplish this for more than 10 mS, a 3-kW server PSU would need a total capacitance of over 1.3 mF, which would consume at least 30% of the overall space. To further improve power density, you must reduce the bulk capacitance.

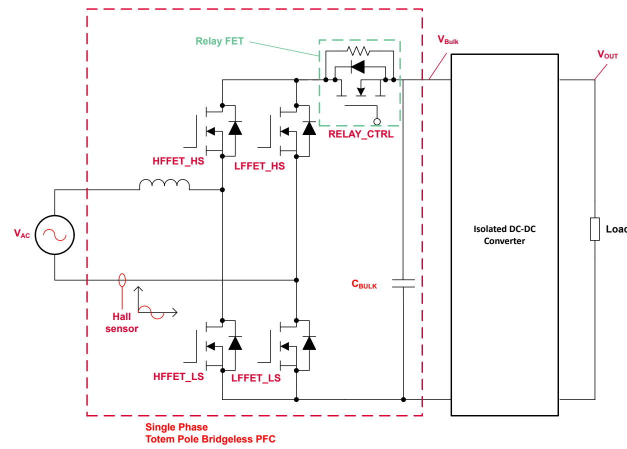

Figure 1. *Server PSU block diagram.*

In this article, the concept and operational principles of a “baby” boost converter (a compact boost converter that only operates during AC dropout events) is introduced to reduce the bulk capacitance. Test results on a PFC reference design [6] with a baby boost converter show that a 910-µF bulk capacitor (as opposed to a 1.3-mF capacitor) is enough to hold the output voltage above 320 V for more than 10 mS after AC dropout with a 3-kW load.

Selecting C_{Bulk} capacitance

The server front end PSU shown in **Figure 1** generally consists of two stages: an AC/DC rectifier stage and an isolated DC/DC converter stage, with peak efficiency targets of >98.5% and >97.5%, respectively. In order for the isolated DC/DC stage to achieve an efficiency target >97.5%, the operational input voltage range of the isolated DC/DC converter (V_{Bulk}) generally has to be limited inside 320 V to 410 V ($V_{Bulk,max}$). Assuming a nominal bulk capacitor voltage ($V_{Bulk,nom}$) of 390 V, **Equation 1** calculates the capacitance required for holding up 3 kW for 10 mS as:

$$C_{Bulk} = \frac{2P_{OUT} \times t_{holdup}}{V_{Bulk,nom}^2 - V_{Bulk,min}^2} = \frac{2 \times 3\,\text{kW} \times 10\,mS}{(390\,V)^2 - (320\,V)^2} = 1.207\,mF \quad (1)$$

Considering V_{Bulk} voltage ripple and capacitance tolerance, the system shown in **Figure 1** would require a capacitor with over 1.3 mF of capacitance. It is notable that the capacitor energy used to hold up the output voltage after AC dropout is only 32.6% of the total energy stored in the bulk capacitor during normal operation.

Inserting a baby boost converter stage in between the AC/DC rectifier stage and the isolated DC/DC converter stage (as shown in **Figure 2**) makes it possible to turn off the bypass field-effect transistor (FET) and enable the baby boost converter to allow the charging of C_{BB} to above 320 V from C_{BULK} after AC dropout. V_{Bulk} can then go much lower than 320 V, thus requiring less capacitance on the bulk capacitor to hold the output voltage for the same amount of time.

Assuming that V_{Bulk} can go down to 240 V ($V_{Bulk,min}$) with the baby boost converter during the AC dropout period, using **Equation 1** equates to a required C_{BULK} of 635 µF, using 62% of the total capacitor energy.

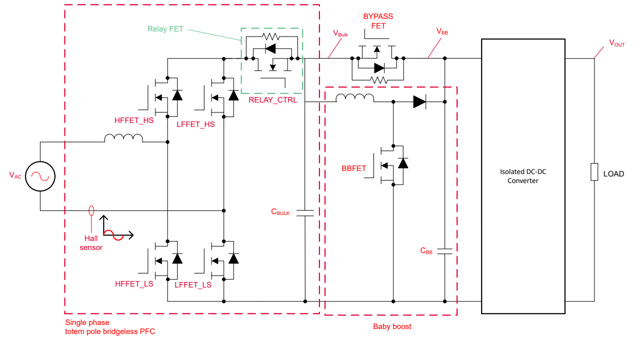

Figure 2. *Server PSU block diagram with a baby boost converter.*

Baby boost converter design considerations

Although a baby boost converter can reduce a bulk capacitor's size and capacitance, minimizing the converter's footprint will help preserve the original high power density goal. Since a baby boost converter operates in a very short amount of time (an AC dropout event), the peak operational current and voltage stresses will determine your power-stage component selections, instead of continuous power dissipation. At $V_{Bulk,min}$, current stress should be at maximum. Select the boost diode and metal-oxide semiconductor FET (MOSFET) to handle current stress at $V_{Bulk,min}$ while rated for $V_{Bulk,max}$. The baby boost inductor needs to handle peak current at $V_{Bulk,min}$.

Equation 2 determines the baby boost inductor inductance:

$$L_{BB} = \frac{V_{Bulk,min} \times \left(V_{BB} - V_{Bulk,min}\right)}{\Delta i_{LBB} \times F_{s,BB} \times V_{BB}} \quad (2)$$

where V_{BB} is the voltage at C_{BB}, Δi_{LBB} is the baby boost inductor peak-to-peak ripple current and $F_{s,BB}$ is the baby boost converter switching frequency.

Since the goal is to minimize the footprint of the inductor, **Equation 3** assumes that the peak-to-peak ripple current equals twice the input current at $V_{Bulk,min}$ and the maximum output power:

$$\Delta i_{LBB} = \frac{2P_{OUT}}{V_{Bulk,min}} = \frac{2 \times 3\,\text{kW}}{240\,\text{V}} = 25\,A \quad (3)$$

With V_{BB} regulating to 390 V and assuming $F_{s,BB}$ = 500 kHz, **Equation 2** calculates LBB as 7.385 µH.

Because the footprint is a higher design priority than the power dissipation, an inductor core with a higher saturation point is preferable; in the case of the baby boost converter, a powder iron core is a better choice than a ferrite core. The soft-saturation characteristic of the powder iron core [7] makes the design of the baby boost inductor design a bit tricky, however. With the core permeability dropping (inductance dropping) as the current increases, you must ensure the LBB calculated in **Equation 2** is the inductance at i_{LBB} peak. **Equation 4** estimates the inductance at a given magnetizing field:

$$L = A_L \times \mu_i\% \times N^2 \tag{4}$$

where A_L is the inductance factor in henry/turns^2, μ_i% is the remaining percentage of the initial permeability at a given magnetizing field, and N is the number of turns applied to the inductor.

Equation 5 expresses the relationship between μ_i% and the magnetizing field, according to the core manufacturer:

$$\mu_i\% = \frac{1}{\left(a + bH^c\right)} \tag{5}$$

where a, b and c are constant coefficients and H is the magnetizing field.

Assuming the application of a Magnetics 0076381A7 [8] (a Kool Mμ H*ƒ* core [9]) for the baby boost inductor, the constant coefficients a, b and c would be 0.01, 4.064·10-7 and 2.131, respectively.

According to Ampere's law, **Equation 6** expresses the relationship between H and N:

$$H = \frac{\mathrm{N} \times \mathrm{I}}{l_e} \tag{6}$$

where I is the current through the winding and l_e is the effective magnetic path length in centimeters.

With **Equation 2** and **Equation 3** calculating L_{BB}, **Equation 4**, **Equation 5** and **Equation 6** determine the N needed to achieve the inductance at a given magnetizing field.

It is also possible to estimate N iteratively. Assuming that a given inductor has an inductance that operates at a certain H with a given current, you can use **Equation 4**, **Equation 5** and **Equation 6** to evaluate whether the calculated H is close to the assumed H.

For example, if your initial guess is that H = 140 Oe with I = 25 A and the inductor has an inductance of 7.385 μH, **Equation 4** calculates μ_i% at 39.65%. Then, taking the L_{BB} calculated from **Equation 2** and **Equation 3**, along with the μ_i% calculated into **Equation 4**, N is then 20.8.

To verify H using **Equation 6** with calculated N, you have H = 125.67 Oe. Since there is still error between the guessed H and the calculated H, you can make a second guess about H and calculate H again until the error becomes negligible. You will find the correct turns (operation point) after a couple of iterations. Using the iterative method, H is 108.75 Oe, with N = 18.009. The inductance is 7.385 μH at 25 A.

Design implementation and test results

Figure 3 shows the Texas Instruments (TI) 3.6-kW Single-Phase Totem-Pole Bridgeless PFC Reference Design with >180-W/in3 Power Density, which uses a baby boost converter [6]. TI's **LMG3522R030** GaN device, which has zero reverse-recovery charges, minimizes switching losses in the totem-pole bridgeless PFC. All of the components are placed in an x and y dimension less than 68 mm by 121 mm, with a maximum component height of 32 mm. The reference design achieves >180-W/in^3 power density and 98.7% peak efficiency. The selected C_{BULK} is a 910-μF 450-V aluminum capacitor.

Although the required capacitance is only 635 μF, the ripple current rating on available capacitors with less than 910 μF of capacitance is not enough to handle the ripple current generated by a single-phase 3-kW PFC. Two 1-μF 450-V ceramic capacitors serve as C_{BB}, which are nicely utilized in the space under the bulk capacitor.

The design applies a Magnetics 0076381A7 core to the baby boost inductor, with 23 turns on the inductor. The inductance at 0 A and 25 A is 22.75 μH and 9.1 μH, respectively. The 9.1-μH inductance allows peak current less than the 25-A target.

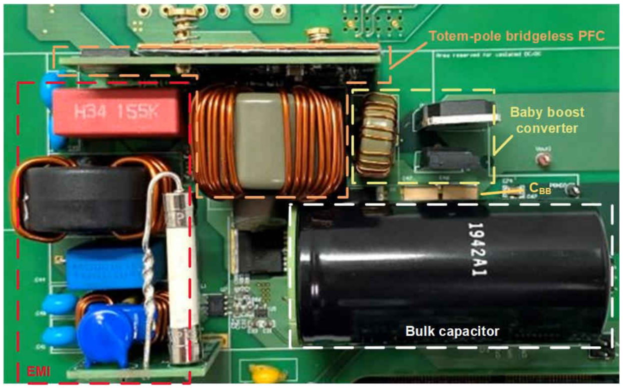

Figure 3. *3.6-kW single-phase totem-pole bridgeless PFC reference design with >180-W/in3 power density.*

Figure 4 shows the measurement of an AC dropout event on the 3-kW PFC reference design. When the AC voltage drops down to 0 V (as does the AC current), C_{BULK} and C_{BB} continuously deliver the energy stored to the load. Once V_{Bulk} drops down to 340 V, the bypass FET turns off and the baby boost converter starts to operate, boosting V_{BB} to 380 V. The baby boost converter continuously operates until V_{Bulk} drops down to 240 V. V_{BB} remains above the targeted isolated DC/DC minimum operational input voltage of 320 V for 14 mS.

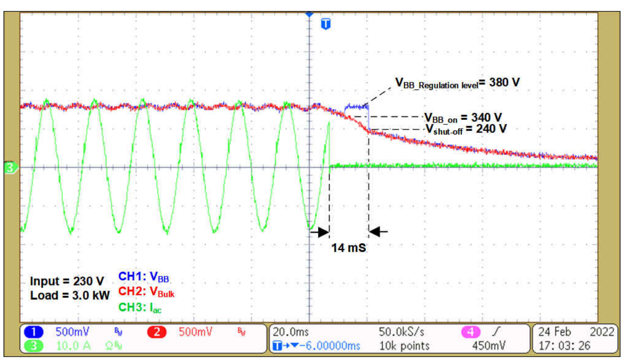

Figure 4. *Waveforms at an AC dropout event.*

References

1. Jovanovic, Milan M. **Power Supply Technology – Past, Present, and Future**. Published in Power Conversion and Intelligent Motion China Conference for Power Electronics (PCIM China), Shanghai, China, March 21-23, 2007.
2. CLEAResult.com. **What is 80 Plus Certified?** Accessed May 6, 2022.
3. Texas Instruments: ***GaN FET-Based CCM Totem-Pole Bridgeless PFC***
4. L. Huber, Y. Jang and M. M. Jovanovic, "Performance Evaluation of Bridgeless PFC Boost Rectifiers," in *IEEE Transactions on Power Electronics*, vol. 23, no. 3, pp. 1381-1390, May 2008.
5. Texas Instruments. n.d. "**4-kW Single-Phase Totem-Pole PFC Reference Design with C2000™ and GaN**." Texas Instruments reference design No. TIDA-010203. Accessed May 6, 2022.
6. Texas Instruments. n.d. "**3.6-kW Single-Phase Totem-Pole Bridgeless PFC Reference Design with a >180-W/in3 Power Density**." Texas Instruments reference design No. PMP23069. Accessed May 6, 2022.
7. Magnetics. "**Learn More about High Flux Cores**." Accessed May 6, 2022.
8. "**0076381A7 Kool Mu Hƒ core.**" Magnetics data sheet. Accessed May 6, 2022.
9. Magnetics. "**Kool Mμ Hƒ Material Curves**." Accessed May 6, 2022.