A New Soft-Switched PFC Boost Rectifier With Integrated Flyback Converter for Stand-by Power

Yungtaek Jang, Senior Member, IEEE, David L. Dillman, and Milan M. Jovanović, Fellow, IEEE

Abstract—This paper presents a magnetic integration approach that reduces the number of magnetic components in a power supply by integrating magnetic components in two conversion stages. Specifically, in the proposed approach, a single transformer is used to implement the continuous-conduction-mode boost power-factor-corrected (PFC) converter and the dc/dc flyback converter. The integrated boost and flyback converters offer soft switching of all semiconductor switches including a controlled di/dt turn-off rate of the boost rectifier. The performance of the proposed approach was evaluated on a 150-kHz, 450-W, universal-line range boost PFC converter with 12-V/2.2-A integrated stand-by flyback converter.

Index Terms—Boost converter, constant-frequency, flyback converter, magnetic integration, power-factor-corrected (PFC), stand-by power, zero-voltage switching (ZVS).

I. INTRODUCTION

The majority of today’s ac/dc power supplies used in modern data processing and telecom equipment have a boost power-factor-corrected (PFC) front end and a low-power stand-by power supply. The block diagram in Fig. 1 shows the typical structure of an offline power supply for those applications. The front-end boost rectifier is employed to reduce the line-current harmonics and to provide compliance with various worldwide specifications governing the harmonic limits of the line current in ac/dc power supplies. The main purpose of the stand-by power supply is to provide housekeeping power and to ensure system functionality when the system is in low-power (stand-by or sleep) mode. The majority of stand-by power supplies are implemented with a flyback converter due to its low parts count and its ability to operate efficiently in a wide input-voltage range.

To meet the challenges of the ever-present requirement to decrease the size of power conversion equipment, power supplies operating at higher switching frequencies and utilizing advanced packaging and thermal management techniques have been introduced. Specifically, in recent years, significant efforts have been made to reduce switching losses of continuous-conduction-mode (CCM) boost converters since the CCM boost converter is the preferred topology for implementation of a front end with PFC over the range from medium to high power. So far, a number of soft-switched boost converters and their variations have been proposed [1]–[16]. All of them employ an auxiliary active switch with a few passive components to form an active snubber that is used to control the di/dt rate of the boost rectifier current and to create conditions for zero-voltage switching (ZVS) or zero-current switching (ZCS) of the boost switch.

A further size reduction can be achieved by minimizing the number of components through component integration. As already have been demonstrated, the number of components can be reduced by integrating semiconductor switches with drive, control and/or supervisory circuits and/or by integrating magnetic components such as transformers and inductors on the same core.

This paper presents a new magnetic integration approach where the reduction of the number of magnetic components in a power supply is achieved by utilizing the same magnetic component in two conversion stages of the power supply. Specifically, in the proposed approach, a single transformer is used to implement the integration of the CCM PFC boost converter and the flyback stand-by converter. The proposed magnetically integrated boost and flyback converters feature soft-switching of all semiconductors. The boost switch and the primary side flyback converter switch are turned on at zero voltage, whereas the active-snubber switch of the boost converter turns off at zero current. In addition, the boost rectifier is turned off softly with a controlled di/dt rate so that reverse-recovery-related losses of the boost rectifier are virtually eliminated.

II. SOFT-SWITCHED PFC BOOST CONVERTER WITH INTEGRATED FLYBACK CONVERTER

The proposed soft-switched boost converter magnetically integrated with a stand-by flyback dc/dc converter is shown in Fig. 2. The boost converter consists of voltage source $V_{IN}$, boost inductor $L_B$, main switch $S_B$, boost rectifier $D$ energy-storage capacitor $C_B$, and the active snubber circuit formed by auxiliary switch $S_1$, winding $N_3$ of transformer TR, snubber inductor $L_S$, and blocking diode $D_1$. The stand-by
flyback converter consists of switch $S_D$ with an associated antiparallel diode, isolation transformer TR, and the secondary side circuit that consists of rectifier $D_R$ and output capacitor $C_f$.

To facilitate the explanation of the circuit operation, Fig. 3 shows a simplified circuit diagram of the proposed converter in Fig. 2. In the simplified circuit, energy-storage capacitor $C_B$ is modeled with voltage source $V_B$ by assuming that the value of $C_B$ is large enough so that the voltage ripple across the capacitor is small in comparison to its dc voltage. In addition, boost inductor $L_B$ is modeled as constant current source $i_{IN}$ by assuming that the inductance of $L_B$ is large so that during a switching cycle the current through $L_B$ does not change significantly. In this analysis, the leakage inductance of the transformer is neglected since it does not have a significant effect on the operation of the circuit. Moreover, since snubber inductor $L_S$ and primary winding $N_1$ of transformer TR is connected in series, the leakage inductance of the transformer is absorbed by $L_S$. As a result, transformer TR is modeled by magnetizing inductance $L_M$ and a three-winding ideal transformer. Finally, it is assumed that in the on state, the semiconductors exhibit zero resistance, i.e., they are short circuits. However, the output capacitance of the switches, as well as the junction capacitance and the reverse-recovery charge of the boost rectifier are not neglected in this analysis.

To further facilitate the analysis of operation, Fig. 4 shows the major topological stages of the circuit in Fig. 2 during a switching cycle, whereas Fig. 5 shows its ideal waveforms. The reference directions of currents and voltages plotted in Fig. 5 are shown in Fig. 3.

As can be seen from the timing diagrams in Fig. 5(a)–(c), the turn on of boost switch $S$ and of flyback switch $S_D$ are synchronized, whereas auxiliary switch $S_1$ is turned on prior to the turn on of switches $S$ and $S_D$. In addition, auxiliary switch $S_1$ is turned off before boost switch $S$ or flyback switch $S_D$ is turned off, i.e., the proposed circuit operates with overlapping gate drive signals for the active snubber switch and the converter switches.

Prior to the turn on of switch $S_1$ at $t = T_0$, all switches are open. As a result, the entire input current $i_{IN}$ flows through boost rectifier $D$ into energy-storage capacitor $C_B$ in the boost power stage, while reflected magnetizing current $(N_2/N_3)i_M$ flows through output rectifier $D_R$ in the flyback power stage as shown in Fig. 4(l). Because output rectifier $D_R$ is conducting during this period, the reflected output voltage is induced across winding $N_1$ of transformer TR, i.e.,$v_1 = -(N_1/N_3)V_O$. After switch $S_1$ is turned on at $t = T_0$, the voltage of energy-storage capacitor $V_B$ and induced voltage $(N_1/N_3)V_O$ is applied across snubber inductor $L_S$ so that current $i_1$ starts to increase linearly, as illustrated in Fig. 5(g). The slope of current $i_1$ is

\[
\frac{di_1}{dt} = \frac{V_B - v_1}{L_S} = \frac{V_B + N_1V_O}{L_S}
\]

where $N_1 = N_1/N_3$.

As current $i_1$ starts flowing through winding $N_1$ of transformer TR, the current in winding $N_3$ starts to decrease, i.e., $i_{DR} = (N_2/N_3)i_M - (N_1/N_3)i_1$, as shown in Figs. 4(a) and 5(k). Current $i_{DR}$ decreases until it becomes zero and output rectifier $D_R$ turns off at $t = T_1$. Since the current through winding $N_3$ is zero after the turn-off of $D_R$, the increasing current in winding $N_1$ makes current $i_2$ in winding $N_2$ larger than magnetizing current $i_M$. This excessive current discharges the output capacitance of switch $S_D$, as illustrated in Figs. 4(b) and 5(d). During this period, voltage $v_2$ across winding $N_2$ of transformer TR starts to increase. After the output capacitance of switch $S_D$ is fully discharged at $t = T_2$, current $i_{DS}$ continues to flow through the antiparallel diode of switch $S_D$, as shown in Figs. 4(c) and 5(i). To achieve ZVS of $S_D$, switch $S_D$ should be turned on while its antiparallel diode is conducting. To simplify the control circuit timing diagram, the turn-on of switch $S_D$ is synchronized with the turn-on of boost switch $S$. When the antiparallel diode of switch $S_D$ is conducting, voltage $v_2$ across winding $N_2$ is equal to $V_B$ so that induced voltage $v_1$ on winding $N_1$ is

\[
v_1 = \frac{N_1}{N_2}V_B = nV_B.
\]

Since $v_1$ is constant, voltage applied across snubber inductor $L_S$ is also constant so that current $i_1$ increases linearly with a slope of

\[
\frac{di_1}{dt} = \frac{V_B - v_1}{L_S} = \frac{V_B - nV_B}{L_S} = (1 - n)\frac{V_B}{L_S}.
\]

During the same period, magnetizing inductance $i_M$ increases with a slope given by

\[
\frac{di_M}{dt} = \frac{V_B}{L_M}.
\]
As current $i_1$ linearly increases, boost rectifier current $i_D$ linearly decreases at the same rate since the sum of $i_1$ and $i_D$ is equal to constant input current $i_{IN}$, i.e., $i_1 + i_D = i_{IN}$. Therefore, in the proposed circuit, the turn-off rate of the boost rectifier

$$\frac{di_D}{dt} = -(1 - n) \frac{V_B}{L_S}$$

(5)

can be controlled by the proper selection of the inductance value of snubber inductor $L_S$ and turns ratio $n$ of transformer TR. Typically, for today's fast-recovery rectifiers, the turn-off rate $\frac{di_D}{dt}$ should be kept around 100 $A/\mu S$. With such a selected turn-off rate, the reverse-recovery current of the rectifier and the related power losses and electromagnetic interference (EMI) problems are minimized.

The topological stage in Fig. 4(c) ends at $t = T_3$ when the current of boost rectifier $D$ becomes zero. During the time period between $t = T_3$ and $t = T_5$, the reverse-recovery current of boost rectifier $D$ flows through snubber inductor $L_S$. After $t = T_4$, current $i_1$ starts to discharge the output capacitance of boost switch $S$ and charge the junction capacitance of boost rectifier $D$, as shown in Fig. 4(e). If the turns ratio of transformer TR is selected so that $N < 0.5$, the energy stored in $L_S$ is sufficient to completely discharge the output capacitance of boost
As shown in Fig. 5(g), current $i_2$ continues to decrease until it reaches zero at $t = T_7$. Shortly after $t = T_7$, auxiliary switch $S_1$ is turned off to achieve ZCS. After switch $S_1$ is turned off, the entire input current $i_{IN}$ flows through boost switch $S$. As a result, the front-end boost converter stage is completely decoupled from the stand-by flyback converter stage, as shown in Fig. 4(h). For the rest of the switching cycle, the flyback converter stage continues to operate as a conventional flyback converter.

After flyback-converter switch $S_{D}$ is turned off at $t = T_k$, magnetizing current $i_M$ starts to charge the output capacitance of switch $S_{D}$, as shown in Fig. 4(i). When voltage $V_{SD}$ reaches $V_B + n_2V_O$ at $t = T_9$, diode $D_{R}$ starts to conduct, which forces the commutation of the magnetizing current from switch $S_{D}$ to output diode $D_{R}$, as shown in Fig. 4(j). At the same time, the reset of the transformer is initiated by applied output voltage $V_O$ across winding $N_3$. During the reset time of the transformer, voltage $V_{SD}$ across the flyback-converter switch is equal to $(N_2/N_3)V_O + V_B$, whereas the voltage across auxiliary switch $S_1$ is $(N_1/N_3)V_O$ due to the magnetic coupling of windings $N_1$ and $N_2$, as illustrated in Fig. 5(d) and (e).

After boost switch $S$ is turned off at $t = T_{10}$, voltage across switch $S$ starts increasing linearly because constant input current $i_{IN}$ starts charging the output capacitance of boost switch $S$, as shown in Fig. 4(k). The increasing boost-switch voltage causes an equal increase of voltage $v_{S1}$ across auxiliary switch $S_1$. When boost-switch voltage $v_{S}$ reaches $V_B$ at $t = T_{11}$, boost diode $D$ begins to conduct, as shown in Fig. 4(l). At the same time, auxiliary-switch voltage $v_{S1}$ reaches its maximum value of $(N_1/N_3)V_O + V_B$. The circuit stays in the topological stage shown in Fig. 4(l) until the next switching cycle is initiated at $t = T_{12}$.

In summary, the major feature of the proposed circuit in Fig. 2 is the soft-switching of all semiconductor devices. Specifically, boost switch $S$ and flyback-converter switch $S_{D}$ are turned on with ZVS, whereas auxiliary switch $S_1$ is turned off with ZCS. In addition, boost diode $D$ is turned off with a controlled turn-off rate of its current. Because all semiconductor components of the proposed converter operates with soft switching, the overall switching losses are minimized, which maximizes the conversion efficiency. In addition, soft switching has a beneficial effect on EMI and may result in a smaller size input filter.

However, it should be noted that complete ZVS of flyback-converter switch $S_{D}$ can only be achieved if input current $i_{IN}$ (which is being commutated to winding $N_1$ when auxiliary switch $S_1$ is closed) is large enough to produce a negative current through primary winding $N_2$ and discharge the output capacitance of switch $S_{D}$, as shown in Fig. 4(b). According to Fig. 4(b), to have a negative current flowing through winding $N_2$ after $t = T_1$, reflected current $i_1$ into winding $N_2$ has to be greater than magnetizing current $i_M$. If this condition is not met, switch $S_{D}$ operates with partial ZVS. This mode of operation typically occurs near the zero crossing of the line voltage in a PFC boost converter. Since the input current is proportional to the line voltage, input current $i_{IN}$ is small near the zero crossing of the line voltage.

Due to the ZVS of the boost switch and the flyback switch, the most suitable implementation of the circuit in Fig. 2 is with the boost switch and the flyback switch consisting of MOSFET (Metal Oxide Semiconductor Field Effect Transistor) devices.
Similarly, due to the ZCS of auxiliary switch $S_1$, an insulated gate bipolar transistor (IGBT) is suitable for the auxiliary switch.

In the proposed circuit, the voltage stresses on switches $S$, $S_D$, and boost rectifier $D$ are identical to the corresponding stresses in the conventional boost converter without a snubber. However, the voltage stress of auxiliary switch $S_1$ is

$$v_{S1\text{(MAX)}} = V_B + n_1 V_O.$$  \hspace{1cm} (7)

The control of the proposed circuit is performed by two independent controllers that are synchronized. Specifically, one controller is used to regulate the output voltage of the front-end boost stage, i.e., voltage $V_B$ across the energy-storage capacitor $C_B$. The other controller is used to regulate output voltage $V_O$ of the flyback converter. Any control strategy can be used to control these two voltages, including multiloop control strategies such as various current-mode control implementations.

### III. EXPERIMENTAL RESULTS

The performance of the proposed soft-switched converter was evaluated on a 150-kHz, 450-W prototype circuit that was designed to operate from a universal ac-line input ($90 V_{\text{RMS}} = 264 V_{\text{RMS}}$) and deliver up to 1.2 A at a 380-V output. The integrated flyback converter was designed to deliver 26-W stand-by power at a 12-V output.

Since the drain voltage of boost switch $S$ is clamped to bulk capacitor $C_B$, the peak voltage stress on boost switch $S$ is approximately 380 V. The peak current stress on switch $S$, which occurs at full load and low line, is approximately 7.8 A. Therefore, an SPP20N60C2 MOSFET ($V_{\text{DSS}} = 600 V$, $i_D=25 = 20 A$, $R_{\text{DS}} = 0.19 \Omega$) from Infineon was used for the boost switch. The maximum drain voltage of flyback switch $S_D$ is $V_{\text{SD(MAX)}} = V_B + n_2 V_O = 380 + (32/4)12 = 536 V$, as shown in Fig. 5(d). The peak current stress on flyback switch $S_D$ is approximately 1.3 A. An SPA111N80C3 MOSFET ($V_{\text{DSS}} = 800 V$, $i_D=25 = 11 A$, $R_{\text{DS}} = 0.45 \Omega$) from Infineon was used for the flyback switch. Finally, a high speed HGTG12N60A4 IGBT ($V_{\text{RRM}} = 600 V$, $i_f = 12 A$) from Fairchild was used as auxiliary switch $S_1$ since its maximum drain voltage is $V_{\text{S1(MAX)}} = V_B + n_1 V_O = 380 + (12/4)12 = 416 V$, as shown in (7).

Since boost diodes $D$ must block the bulk voltage and must conduct the peak input current, which is approximately 7.8 A, an RHRP1560 diode ($V_{\text{RRM}} = 600 V$, $i_F=25 = 15 A$) from Fairchild was used as boost diode $D$. An RHRP860 diode ($V_{\text{RRM}} = 600 V$, $i_F=25 = 8 A$) was used as diode $D_1$, whereas a 16CTQ060 diode ($V_{\text{RRM}} = 60 V$, $i_F=25 = 16 A$) from IRF was used as output diode $D_R$.

To obtain the desired inductance of boost inductor $L_B$ of approximately 250 $\mu$H at full load, the boost inductor was built using a toroidal core (MS130060) from Arnold and 71 turns of magnet wire (AWG #19).

External snubber inductor $L_S$ was connected in series with winding $N_1$ of transformer $TR$, as shown in Fig. 2. The required inductance is approximately 2.4 $\mu$H at full load. Snubber inductor $L_S$ was built using a toroidal core (A189043) from Arnold and eight turns of magnet wire (AWG #19).

Transformer $TR$ was built using a pair of ferrite cores (Philips, ER35-3F3) with an air gap of 0.102 mm, 12 turns of magnet wire (AWG# 21) for winding $N_1$, 52 turns of magnet wire (AWG# 29) for winding $N_2$, and four turns of magnet wire (AWG# 21) for winding $N_3$. Magnetizing inductance $L_M$ measured across winding $N_2$ of the transformer is approximately 4 mH. A high voltage aluminum capacitor (470 $\mu$F, 450 VDC) was used for bulk capacitor $C_B$ to meet the hold-up time requirement. A low voltage aluminum capacitor (220 $\mu$F, 16 VDC) was used for output capacitor $C_f$.

An average-current-mode PFC controller (UC3854) from TI is employed for the PFC front end rectifier. The measured total harmonic distortion (THD) of the input current is less than 5% at full load and low line.

Figs. 6 and 7 show the oscillograms of key waveforms in the experimental converter when it delivers full power from the low line input voltage. It should be noted that the waveforms were captured when the ac input voltage is near its peak. As can be seen from the corresponding waveforms in Fig. 5, there is a good agreement between the experimental and theoretical waveforms. As can be seen from Figs. 6 and 7, switches $S$ and $S_D$ are turned on with ZVS since their voltages $V_S$ and $V_{SD}$ fall to zero before gate-drive signals $V_{GS}$ and $V_{CSD}$ become high. Moreover, auxiliary switch $S_1$ achieves soft-switching turn off because switch current $i_S$ becomes zero before auxiliary switch $S_1$ is turned off. It should be also noted that the slope of rectifier current $i_D$ is approximately $\frac{di}{dt} = 80 \ A/\mu s$ during the
period when boost diode $D$ is turned off. The rectifier-current slope is controlled by snubber inductance $L_{S}$, as indicated in Figs. 6 and 7. With this $di/dt$ rate, peak reverse-recovery current $i_{RR}$ is reduced to approximately 2 A.

Fig. 8 shows the measured efficiencies of the experimental converter with the active snubber circuit (solid line) and without the active snubber circuit (dotted line) as functions of the output power of PFC front end. To measure the efficiency of the experimental converter without the active snubber circuit, switch $S_1$, diode $D_1$, inductor $L_S$, and winding $N_1$ of the proposed converter shown in Fig. 2 are disconnected. For both experimental circuits, the integrated flyback converter delivers 26 W at 12-V output. To obtain the efficiencies, the summation of the output powers of the PFC boost converter and the flyback converter is divided by the ac input power. As can be seen in Fig. 8, the active snubber improves the conversion efficiency in the entire measured power range. The efficiency improvement is more pronounced at higher power levels where the reverse-recovery losses are greater. The active snubber improves the efficiency by approximately 5% at 450 W, which translates into approximately 50% reduction of losses. Since all switches operate with zero-voltage or zero-current switching, the rectifier reduces switching losses and is also expected to improve EMI [17].

IV. CONCLUSION

A new PFC boost converter with an integrated stand-by flyback converter that can achieve soft-switching of all semiconductor devices in the power stages has been introduced. By using a single magnetic device which is mutually shared by the PFC boost converter and the stand-by flyback converter, boost switch $S$ and flyback switch $S_D$ are turned on with ZVS, auxiliary switch $S_1$ is turned off with ZCS, and boost diode $D$ is turned off softly using a controlled $di/dt$ rate. As a result, the turn-on switching losses in the boost and flyback switches, the turn-off switching loss in the auxiliary switch, and reverse-recovery-related losses in the boost diode are eliminated, which maximizes the conversion efficiency. The performance of the proposed converter was verified on a 150-kHz, 450-W prototype circuit that was designed to operate from a universal ac-line input. The PFC boost converter and the stand-by flyback converter of the prototype circuit deliver up to 1.2 A at a 380-V output and 2.2 A at a 12-V output, respectively. The proposed technique improves the efficiency by approximately 5% at 450 W.

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Dr. Jang received the IEEE TRANSACTIONS ON POWER ELECTRONICS Prize Paper Award for best paper published in 1996.

David L. Dillman received the A.S.E.E. degree from DeVry Technical Institute of Technology, Chicago, IL, in 1966.

He is currently a Support Engineer at Delta Product DEPL Laboratory, Research Triangle Park, NC. He has 30 years of switching power supply and analog component experience, including 25 years in aerospace and military power supply development. He was Group Leader for a thick and thin film hybrid engineering team that developed RF and analog hybrid modules used in aerospace and military systems.

Mr. Dillman received the IBM Outstanding Innovation Award for Development and Implementation of Piece Part Control Program for Navy Contractors.

Yungtaek Jang (S’92–M’95–SM’01) was born in Seoul, Korea. He received the B.S. degree from Yonsei University, Seoul, in 1982, and the M.S. and Ph.D. degrees from the University of Colorado, Boulder, in 1991 and 1995, respectively.

Since 1996, he has been a Senior Member of R&D Staff with the Delta Power Electronics Laboratory, Delta Products Corporation, Research Triangle Park, NC. He has authored more than 40 papers published in power electronics journals and conference proceedings and is the holder of 14 U.S. patents.

Dr. Jang received the IEEE TRANSACTIONS ON POWER ELECTRONICS Prize Paper Award for best paper published in 1996.

Milan M. Jovanović (F’01) was born in Belgrade, Serbia. He received the Dipl.Ing. degree in electrical engineering from the University of Belgrade, Serbia.

Presently, he is the Chief Technology Officer (CTO) of Power Systems Business Group of Delta Electronics, Inc., Taipei, Taiwan, R.O.C.