Soft-Switched PFC Boost Rectifier With Integrated ZVS Two-Switch Forward Converter

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

Abstract—A soft-switched continuous-conduction-mode boost power factor correction front-end converter with an integrated zero-voltage-switched two-switch forward second-stage converter is introduced. In the proposed approach, a single transformer is commonly used by the two stages to provide isolation of the power supply and 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, 430-W/12-V, universal-line range prototype converter.

Index Terms—Boost converter, magnetic integration, power factor correction (PFC), two-switch forward converter, zero voltage switching (ZVS).

I. INTRODUCTION

A BOOST power-factor-corrected (PFC) front-end converter followed by a dc-dc two-switch forward converter is one of the most extensively employed converter combinations in off-line power supplies used in low-end computer servers and high-end desk top computers. 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 off-line power supplies, whereas the two-switch forward converter is employed to provide galvanic isolation and tight output voltage regulation. The popularity of the two-switch forward converter topology stems from its maturity, simplicity, robustness, good performance, and low cost.

The continuous-conduction-mode (CCM) boost converter is the preferred topology for implementation of a front end with PFC over the range of medium to high power. In recent years, significant efforts have been made to improve the performance of high-power CCM boost converters [1]–[5]. The majority of these development efforts have been focused on reducing the adverse effects of the reverse-recovery characteristic of the boost diode on the conversion efficiency and electromagnetic compatibility (EMC) [6]. Similar effort has been put in optimizing and improving the performance of the two-switch forward converter [7], [8]. However, so far no circuit that offers soft switching of both the CCM boost converter front end and the downstream two-switch forward converter dc–dc stage has been reported.

In this paper, a novel ac–dc converter that integrates the soft switching circuit of the CCM boost front end with the dc–dc two-switch forward converter is described. The integration of the two power stages is achieved by a magnetic component that is shared by both stages. This approach not only reduces the number of magnetic components, but also makes it possible to achieve a fully soft-switched ac–dc converter. Namely, in the integrated circuit, not only are the switches in the PFC boost converter soft switched, the switches in the two-switch forward converter are also able to achieve soft switching.

II. SOFT-SWITCHED PFC BOOST CONVERTER WITH INTEGRATED TWO-SWITCH FORWARD CONVERTER

The proposed soft-switched PFC boost converter with integrated two-switch forward converter is shown in Fig. 1. The boost converter consists of voltage source $V_{IN}$, boost inductor $L_B$, main switch $S$, boost rectifier $D$, energy-storage capacitor $C_B$, and the active snubber circuit formed by auxiliary switch $S_1$, winding $N_1$ of transformer TR, snubber inductor $L_S$, and blocking diode $D_1$. The two-switch forward converter consists of switches $S_{D1}$ and $S_{D2}$ with associated antiparallel diodes, isolation transformer TR, rectifiers $D_{R1}$ and $D_{R2}$, output inductor $L_F$, and output capacitor $C_F$.

To facilitate the explanation of the circuit operation, Fig. 2 shows a simplified circuit diagram of the proposed converter in Fig. 1. In the simplified circuit, energy-storage capacitor $C_B$ is modeled by 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 $L_B$ and output filter inductor $L_F$ are modeled as constant current sources $I_{IN}$ and $I_C$, respectively, by assuming that the inductance of $L_B$ and $L_F$ are large so that during a switching cycle the currents through $L_B$ and $L_F$ do not change significantly.

In this analysis, the leakage inductance of the transformer is neglected because 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 are 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 the 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. 3 shows the major topological stages of the circuit in Fig. 1 during a switching cycle, whereas Fig. 4 shows its key waveforms. The reference directions of currents and voltages plotted in Fig. 4 are shown in Fig. 2.

As can be seen from the timing diagrams in Fig. 4(a)–(c), the turn on of boost switch $S$ and of forward switches $S_{D1}$ and $S_{D2}$ are synchronized, whereas auxiliary switch $S_1$ is turned on prior to the turn on of switches $S$, $S_{D1}$, and $S_{D2}$. In addition, auxiliary switch $S_1$ is turned off before boost switch $S$ or forward switches $S_{D1}$ and $S_{D2}$ are turned off. Consequently, 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 output current $I_O$ flows through output rectifier $D_{R2}$ in the two-switch forward power stage as shown in Fig. 3(j). Because output rectifier $D_{R2}$ is conducting during this period, voltage $v_3$ and induced voltage $v_1$ across winding $N_1$ of transformer TR is zero, i.e., $v_1 = (N_1/N_3)v_3 = 0$.

### A. Stage 1 [$T_0, T_1$]

After switch $S_1$ is turned on at $t = T_0$, the voltage of energy-storage-capacitor $v_B$ is applied across snubber inductor $L_S$ so that current $I_1$ starts to increase linearly, as illustrated in Fig. 4(g). The slope of current $I_1$ is

$$\frac{di_1}{dt} = \frac{v_B}{L_S}. \quad (1)$$

As current $I_1$ starts flowing through winding $N_1$ of transformer TR, the current in winding $N_3$ also begins to increase, i.e., $I_{DR1} = (N_1/N_3)I_1$, as shown in Figs. 3(a) and 4(l). Because output current $I_O$ is constant and equal to the sum of rectifier currents $I_{DR1}$ and $I_{DR2}$, rectifier current $I_{DR2}$ decreases until it becomes zero when rectifier current $I_{DR1}$ increases. When rectifier current $I_{DR2}$ becomes zero at $t = T_1$, output rectifier $D_{R2}$ turns off, as shown in Fig. 4(m).

### B. Stage 2 [$T_1, T_2$]

Since the current through winding $N_3$ and rectifier $D_{R1}$ is equal to output current $I_O$ after the turn-off of $D_{R2}$, the increasing current in winding $N_1$ makes current $I_2$ in winding $N_2$ begin to flow. This current discharges the output capacitances of forward switches $S_{D1}$ and $S_{D2}$, as illustrated in Figs. 3(b) and 4(i). During this period, voltage $v_2$ across winding $N_2$ of transformer TR starts to increase. After the output capacitances of forward switches $S_{D1}$ and $S_{D2}$ are fully discharged, switch currents $I_{SD1}$ and $I_{SD2}$ continue to flow through the antiparallel diodes of forward switches $S_{D1}$ and $S_{D2}$, as shown in Figs. 3(c) and 4(i). To achieve ZVS of forward switches $S_{D1}$ and $S_{D2}$, switches $S_{D1}$ and $S_{D2}$ should be turned on while their antiparallel diodes are conducting. To simplify the control circuit timing diagram, the turn-on of forward switches $S_{D1}$ and $S_{D2}$ is synchronized with the turn-on of boost switch $S$. While the antiparallel diodes of forward switches $S_{D1}$ and $S_{D2}$ are conducting, voltage $v_2$ across winding $N_2$ is equal to $v_B$ so that induced voltage $v_1$ on winding $N_1$ is

$$v_1 = \frac{N_1}{N_2}v_B = nV_B. \quad (2)$$

Because $v_1$ is constant, voltage applied across snubber inductor $L_S$ is also constant so that current $I_1$ increases linearly with a slope of

$$\frac{di_1}{dt} = \frac{v_B - v_1}{L_S} = \frac{v_B - nV_B}{L_S} = \frac{(1-n)v_B}{L_S}. \quad (3)$$

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

$$\frac{di_M}{dt} = \frac{v_B}{L_M}. \quad (4)$$

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

$$\frac{di_D}{dt} = -(1-n)\frac{v_B}{L_S}. \quad (5)$$

can be controlled by the proper selection of the inductance value of snubber inductor $L_S$ and turns ratio $n$ of transformer TR.
Fig. 3. Topological stages: (a) \([T_0 - T_1]\), (b) \([T_1 - T_2]\), (c) \([T_2 - T_3]\), (d) \([T_3 - T_4]\), (e) \([T_4 - T_5]\), (f) \([T_5 - T_6]\), (g) \([T_6 - T_7]\), (h) \([T_7 - T_8]\), (i) \([T_8 - T_9]\), and (j) \([T_9 - T_{10}]\)

Typically, for today’s fast-recovery rectifiers, the turn-off rate \(\frac{di_D}{dt}\) should be kept around 100 A/\(\mu\)s. With the selected turn-off rate, the reverse-recovery current of the rectifier and the related power losses and EMI problems are minimized.

C. Stage 3 \([T_2, T_3]\)

After \(t = T_2\), 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. 3(c). If the turns ratio of transformer \( T \) is selected so that \( n < 0.5 \), the energy stored in \( L_S \) is sufficient to completely discharge the output capacitance of boost switch \( S \) regardless of the load and line conditions.

### D. Stage 4 \([T_3, T_4]\)

Once the capacitance is fully discharged at \( t = T_3 \), current \( I_S \) continues to flow through the antiparallel diode of boost switch \( S \), as shown in Figs. 3(d) and 4(h). During this period, voltage \( v_{\text{B}} \) is applied in the negative direction across snubber inductor \( L_S \). Therefore, current \( I_1 \) starts to decrease linearly at the rate given by

$$\frac{di_1}{dt} = \frac{nV_B}{L_S} \quad (6)$$

as illustrated in Fig. 4(g). The current in auxiliary switch \( S_1 \) also starts to decrease, whereas boost-switch current \( I_S \) starts to increase from the negative peak value, as shown in Fig. 4(g) and (h). To achieve ZVS of boost switch \( S \), it is necessary to turn on boost switch \( S \) before its current becomes positive at \( t = T_4 \) i.e., during the period when current \( I_S \) still flows through the antiparallel diode of switch \( S \), as illustrated in Fig. 4(h). To design the gate signals of the prototype circuit, fixed delay time \( T_{D1} \) should be introduced between the turn-on instance of auxiliary \( S_1 \) and the turn-on instance of boost switch \( S \). The approximate value of delay time \( T_{D1} \) can be calculated by

$$T_{D1} = \frac{I_{\text{IN-Peak}}}{rac{nL_S}{V_B}} \quad (7)$$

with the condition that turns ratio \( n \) is much smaller than 1 and the period between \( T_2 \) and \( T_3 \) is much shorter than the period between \( T_0 \) and \( T_2 \). \( I_{\text{IN-Peak}} \) is the maximum input current at low line and full load.

### E. Stage 5 \([T_4, T_5]\)

After \( t = T_4 \), current \( I_1 \) continues to decrease until it reaches zero at \( t = T_5 \), as shown in Fig. 4(g). Shortly after \( t = T_5 \), auxiliary switch \( S_1 \) is turned off to achieve zero-current switching (ZCS). The gate signal of auxiliary switch \( S_1 \) of the prototype circuit can be a constant on-time pulse signal. Fixed turn-on time \( T_{ON-S1} \) of auxiliary switch \( S_1 \) is approximately given by

$$T_{ON-S1} = T_{D1} + \frac{I_{\text{IN-Peak}}}{nV_B L_S} \quad (8)$$

### F. Stage 6 \([T_5, T_6]\)

After switch \( S_1 \) is turned off shortly after \( t = T_5 \), the entire input current \( I_{\text{IN}} \) flows through boost switch \( S \). As a result, the front-end boost converter stage is completely decoupled from the two-switch forward converter stage, as shown in Fig. 3(f). For the rest of the switching cycle, the two-switch forward converter stage continues to operate as a conventional two-switch forward converter.

### G. Stage 7 \([T_6, T_7]\)

After forward switches \( S_{D1} \) and \( S_{D2} \) are turned off at \( t = T_6 \), magnetizing current \( I_M \) starts to charge the output capacitances of forward switches \( S_{D1} \) and \( S_{D2} \). When voltages \( v_{SD1} \) and \( v_{SD2} \) reach \( v_B \), the magnetizing current is diverted from forward switches \( S_{D1} \) and \( S_{D2} \) to clamp diodes \( D_{D1} \) and \( D_{D2} \), as shown in Fig. 3(g). At the same time, the reset of the transformer is initiated by bulk voltage \( v_B \) applied across winding \( N_2 \). During the reset time of the transformer, forward switch voltages \( v_{SD1} \) and \( v_{SD2} \) are equal to \( v_B \), whereas the voltage across auxiliary switch \( S_1 \) is \( v_{NB} \) due to the magnetic coupling of winding \( SN_1 \) and \( N_2 \), as illustrated in Fig. 4(d) and (e).

### H. Stage 8 \([T_7, T_8]\)

After boost switch \( S \) is turned off at \( t = T_7 \), voltage across switch \( S \) starts to increase linearly because constant input current \( I_{\text{IN}} \) begins charging the output capacitance of boost
switch \( S \), as shown in Fig. 3(h). The increasing boost-switch voltage causes an equal increase of voltage \( v_{\text{S1}} \) across auxiliary switch \( S_1 \). This stage ends when boost-switch voltage \( v_{\text{S}} \) reaches \( v_B \) at \( t = T_8 \).

I. Stage 9 \([T_8, T_9] \)

When boost-switch voltage \( v_{\text{S}} \) reaches \( v_B \) at \( t = T_8 \), boost diode \( D \) begins to conduct, as shown in Fig. 3(i). At the same time, auxiliary-switch voltage \( v_{\text{S1}} \) reaches its maximum value of \((1 + n)v_B\). The circuit stays in the topological stage shown in Fig. 3(i) until magnetizing current \( I_M \) decreases to zero at \( t = T_9 \). The next switching cycle is initiated at \( t = T_{10} \).

In summary, the major feature of the proposed circuit in Fig. 1 is the soft-switching of all semiconductor devices. Specifically, boost switch \( S \) and forward switches \( S_{D1} \) and \( S_{D2} \) are turned on with ZVS, whereas auxiliary switch \( S_1 \) is turned off with ZCS. It should be noted that boost switch \( S \) and forward switches \( S_{D1} \) and \( S_{D2} \) are also turned off with soft switching because the output capacitances of the switch devices do not allow immediate increase of the switch voltages when they are turned off. Energy that is stored in the output capacitances of the switch devices when they are turned off is recovered just before the switches are turned on with ZVS. To reduce the turn-off losses even further, additional capacitors can be connected in parallel with the switches. The turn-on loss of the auxiliary switch \( S_1 \) is also minimized by the series connected snubber inductor \( L_S \) because the current of \( S_1 \) slowly increases during the period when it is turned on as indicated in (1). The turn-on loss that is a product of the current and voltage of the switch when it is turned on can be further reduced by using a larger snubber inductor \( L_S \). 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 [6].

However, it should be noted that complete ZVS of forward switches \( S_{D1} \) and \( S_{D2} \) can only be achieved if input current \( I_{IN} \) is large enough to produce a negative current through primary winding \( N_2 \) and discharge the output capacitances of switches \( S_{D1} \) and \( S_{D2} \) completely, as shown in Fig. 3(b). According to Fig. 3(b), to have a negative current flowing through winding \( N_2 \) after \( t = T_1 \), reflected current \( I_1 \) into winding \( N_3 \) has to be greater than output current \( I_{OL} \). If this condition is not met, forward switches \( S_{D1} \) and \( S_{D2} \) operate 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. However, by adding an extra capacitor across boost switch \( S \), forward switches \( S_{D1} \) and \( S_{D2} \) can achieve complete ZVS near the zero crossing of the line voltage.

Due to the ZVS of boost switch \( S \) and forward switches \( S_{D1} \) and \( S_{D2} \), the most suitable switch component is a MOSFET device. 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 boost switch \( S \), forward switches \( S_{D1} \) and \( S_{D2} \), and boost rectifier \( D \) are identical to the corresponding stresses in the conventional converters. However, the voltage stress of auxiliary switch \( S_1 \) is

\[
v_{\text{S1(MAX)}} = (1 + n)v_B. \tag{9}\]

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 two-switch forward converter.

III. EXPERIMENTAL RESULTS

The performance of the proposed converter was evaluated on a 430-W prototype circuit that was designed to operate from a universal ac line input and deliver up to 36 A at 12-V output. Switches \( S \), \( S_1 \), \( S_{D1} \), and \( S_{D2} \) operate at 150 kHz.

The experimental circuit was implemented with the following components: boost switch \( S \) and two-switch forward switches \( S_{D1} \) and \( S_{D2} \)—SPP20N60C2; auxiliary switch \( S_1 \)—SPA11N80C3; boost diode \( D \) and snubber diode \( D_1 \)—RHRP1560; output diodes \( D_R1 \) and \( D_R2 \)—S60SC6M; bulk capacitor \( C_B = 470 \mu F/450 \text{V} \); and output capacitor \( C_F = 2 \times 2200 \mu F/16 \text{V} \).

To build boost inductor \( L_B \), a toroidal core (MS130060) from Arnold and 71 turns of magnet wire (AWG # 19) were used. External snubber inductor \( L_S \) was connected in series with the auxiliary winding of transformer TR, as shown in Fig. 1. The required inductance is approximately 2.7 \( \mu \text{H} \) at full load. Snubber inductor \( L_S \) was built using a toroidal core (A189043) and nine turns of magnet wire (AWG #19). Transformer TR was built using a pair of ferrite cores (PIJ0-3C94). Three magnet wires (\( N_1 = \text{seven turns} \); \( N_2 = 26\) turns : \( N_3 = \text{two turns} \)) were used.

Fig. 5 shows the measured efficiencies of the experimental converter with (solid lines) and without (dashed lines) the active snubber circuit as functions representing output current. As can be seen in Fig. 5, the active snubber improves the conversion efficiency in the entire measured power range. The active snubber improves the efficiency by approximately 1.5% at full load.

Figs. 6 and 7 show the oscillograms of key waveforms of the experimental converter at full and light loads, respectively. As can be seen from the corresponding waveforms in Fig. 4, there is a good agreement between the experimental and theoretical waveforms. As can be seen from Figs. 6 and 7, switches \( S \) and \( S_{D2} \) are turned on with ZVS since their voltages \( v_S \) and \( v_{S2} \) fall to zero before gate-drive signals \( v_{GS} \) and \( v_{GS2} \) become high. Moreover, auxiliary switch \( S_1 \) achieves soft-switching turn-off because switch current \( I_1 \) becomes zero before auxiliary switch \( S_1 \) is turned off. Also, it should be noted that the rising slope of current \( I_1 \) is approximately \( di_1/dt = 80 \text{ mA/\mu s} \), which is proportional to boost diode current \( I_D \) during the period when boost diode \( D \) is turned off as shown in Figs. 6 and 7. The reverse-recovery loss of boost diode \( D \) is dramatically reduced by the controlled \( di_D/dt \). Figs. 8 and 9 show the ac
Fig. 5. Measured efficiency of 150 kHz, 430-W experimental converter with (dashed line) hard switching and (solid line) soft switching at \( v_{IN} = 90 \, v_{AC} \), \( v_B = 380 \, V \), and \( v_O = 12 \, V \) as functions representing output current.

Fig. 6. Measured waveforms of experimental circuit at \( v_{IN} = 90 \, V \), \( v_B = 380 \, V \), \( I_O = 2 \, A \), \( v_O = 12 \, V \). Time base: 1 \( \mu s/\text{div} \).

Fig. 7. Measured waveforms of experimental circuit at \( v_{IN} = 90 \, V \), \( v_B = 380 \, V \), \( I_O = 2 \, A \), \( v_O = 12 \, V \). Time base: 1 \( \mu s/\text{div} \).

Fig. 8. Measured waveforms of experimental circuit at \( v_{IN} = 90 \, V \), \( v_B = 380 \, V \), \( I_O = 36 \, A \), \( v_O = 12 \, V \). Time base: 2 ms/\text{div}.

Fig. 9. Measured waveforms of experimental circuit at \( v_{IN} = 90 \, V \), \( v_B = 380 \, V \), \( I_O = 3.6 \, A \), \( v_O = 12 \, V \). Time base: 2 ms/\text{div}.

Waveforms of input voltage \( v_{IN} \) and input current \( I_{IN} \) at full load and 10% load, respectively. The total harmonic distortions (THD) of input current \( I_{IN} \) shown in Figs. 8 and 9 are approximately 8.5% at full load and 26% at 10% load. The power factor (PF) of the prototype circuit at full load and 10% load are approximately 98% and 89%, respectively.
IV. CONCLUSION

A soft-switched boost PFC front-end converter with an integrated ZVS two-switch forward second-stage converter has been introduced. By using a single magnetic device which is mutually shared by the PFC boost converter and the two-switch forward converter, boost switch \( S_1 \) and forward switches \( S_{D1} \) and \( S_{D2} \) are turned on with ZVS, auxiliary switch \( S_2 \) 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 forward 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 approach was evaluated on a 150-kHz, 430-W, universal-line range prototype converter delivering 12-V/36-A output. The proposed technique improves the efficiency by approximately 1.5% at full load.

REFERENCES


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

From 1982 to 1988, he was a Design Engineer at Hyundai Engineering Co., Korea. Since 1996, he has been a Senior Member of R&D Staff at the Power Electronics Laboratory, Delta Products Corporation, Research Triangle Park, NC (the U.S. subsidiary of Delta Electronics, Inc., Taipei, Taiwan, R.O.C.). He holds 16 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 of the Power Systems Business Group, Delta Electronics, Inc., Taipei, Taiwan, R.O.C.

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 Power Electronics Laboratory, Delta Products Corporation, 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 an IBM Outstanding Innovation Award for Development and Implementation of the Piece Part Control Program for Navy contractors.