Cost-Effective Sustain Driver Employing a New Four-Quadrant Switch Cell for AC Plasma Display

J. H. Park, Member, IEEE, Jongwon Shin, Student Member, IEEE, Woosup Kim, Student Member, IEEE, and B. H. Cho, Senior Member, IEEE

Abstract — A new energy-recovery circuit (ERC) for the Plasma Display Panel (PDP) using a switched transformer is proposed. The proposed ERC is very cost effective due to the reduction of device count and circuit complexity such as high-side gate drivers and diodes in the resonant path. Also, the leakage inductances of the transformers are used as resonant inductors for soft-switching operation of the main switches. The operation and performance of the proposed energy recovery circuit were validated through waveform and power-efficiency measurement tests with prototype hardware in 200kHz switching frequency.

Index Terms — Plasma display panel (PDP), switched-transformer, cost-effective, energy-recovery, sustain driver.

I. INTRODUCTION

In current large-size and flat panel color display (FPD) markets, plasma display panel (PDP) industry and liquid crystal display (LCD) vendor have occupied a dominant portion of the production [1]-[2]. The PDP is a powerful candidate for emissive display markets of future FPD industry due to some advantages such as large screen size, wide view angle, thinness, high brightness and rapid response time [3]. However, competitors such as LCDs and organic light emitting diodes (OLEDs) are extending the market share more and more in display panels by emphasizing low power consumption and high contrast ratios. Therefore, research on power-efficient and cost-effective PDP driving circuits is vital to surviving in the consumer electronics market.

A practical and feasible circuit implementation requires a precise understanding of PDP’s photoelectric load characteristics. Cells of the PDP are electrically modeled as an equivalent panel capacitance $C_P$ because the dielectric and MgO layers fully cover the discharge electrodes in the unit cell [4]. However, when the sustain voltage $V_s$, which has an alternating current (AC) square waveform with a high peak-to-peak value greater than 200V, is applied to the sustaining and scanning electrodes, discharge energy as much as $C_p V_s^2$ is dissipated in every operating cycle. This reactive energy discharge with hard-switching causes not only the power dissipation but also a peak surge current and significant electromagnetic interference (EMI). To minimize this capacitive energy loss and improve the power efficiency of the PDP system, an energy recovery circuit (ERC) is necessary.

In general, there are two kinds of ERCs: the series-resonant type and the parallel-resonant type. Weber’s ERC, shown in Fig. 1, is the most representative series-resonant type ERC [4]. It is widely used by PDP makers in the market due to its high power efficiency and good circuit flexibility. Many other researchers such as Ohba et al. have analyzed and proposed various type of ERCs including a parallel-resonant type to improve circuit performance and power efficiency [5]-[23]. However, they have some drawbacks such as complex circuit configuration, high device count and high cost. The conventional PDP drivers have been realized based on a few expensive components such as a high-speed bootstrapping circuitry with high-voltage and high-current ratings in order to drive gates/bases elevated to several hundred volts of Metal Oxide Semiconductor Field Effect Transistors (MOSFETs) or Insulated Gate Bipolar Junction Transistors (IGBTs) [24]. In addition to the basic driver component, several peripheral functions such as short-circuit, under-voltage, and thermal protection should be added to the high-side gate drivers, which result in the increase of overall PDP module costs [2].

In this paper, a new series-resonant type ERC is proposed to overcome these aforementioned disadvantages. Study of this plasma display research focuses on the gate driver which can be a primary factor of the PDP module manufacturing cost. The proposed ERCs replace the conventional floating-MOSFET/IGBT circuit configuration to a ground-referenced switching network, which eliminates the complicated and expensive high side gating driver composed of a RS-flop buffered by a high-current driver, with bootstrapping circuitry to provide gate-voltage drive above the supply rail [24].

In the following sections, the proposed ERC will be explained and analyzed using operational mode analysis, and will be compared with Weber’s ERC with respect to device count and conduction power loss. Experimental results will be described to verify the characteristics of the proposed ERC.

II. BI-DIRECTIONAL SWITCH CELL

A. Switching Cells for Energy-Recovery

Since energy recovery under PDP sustain driving mode requires bi-directional current flow and bi-polar voltage...
Figure 1. AC PDP sustain driver with energy recovery circuit, proposed by Weber [4].

Figure 2. Previous 4-quadrant switch cells

blocking, a four-quadrant switch cell is necessary to process the energy. Widely-used conventional switch cells for high speed power processing are shown in Fig. 2 [25]. Weber’s ERC in Fig. 1 also utilizes one of the bi-directional switch cells.

In this paper, a new four-quadrant switch cell using a switched transformer is proposed. The cell is shown in Fig. 3(a). It is composed of a transformer and two MOSFETs. Close examination of the cell shows that since the switches are connected in cascade through the current balance transformer, the switch cell can block both of bi-directional current flow and bi-polar voltage enhancement. There are two more extra connecting nodes at the MOSFET’s source terminal in the proposed cell. The terminals are used to connect with common ground, eliminating the expensive and complicated high-side gate drivers.

The proposed ERC, employing the switched transformer cell in Fig. 3(a) is shown in Fig. 3(b). It features similar operating waveforms to those of Weber’s ERC with a smaller number of devices and a simple circuit structure. The ERC consists of a transformer \( T_1-T_2 \), auxiliary switches \( S_{A1}-S_{A4} \), and energy-recovery capacitors \( C_1-C_4 \). The conventional resonant inductors \( L_{lk1} \) and \( L_{lk2} \) were replaced with leakage inductances \( T_1 \) and \( T_2 \) respectively.

The proposed ERC does not include high-side gate drivers because all the auxiliary switches operate with their sources connected to a common ground. The leakage inductances of the transformers are used as the resonant inductors for soft-switching operation of the main switches as well as for sustain energy-recovery. Furthermore, the proposed ERC has no diodes in the resonant path because the auxiliary switches themselves \( S_{A1}, S_{A2}, S_{A3}, S_{A4} \) block the resonant path by the simultaneous turning-off of both switches.

III. ENERGY RECOVERY CIRCUIT WITH SWITCHED TRANSFORMER

The proposed ERC has six modes per switching cycle in a sustaining period; these are similar to Weber’s scheme.

A. Operation Mode Analysis

The key waveforms and equivalent circuits in each mode are displayed in Figs. 5 and 6, respectively. Since the modes from \( t_1 \) to \( t_6 \) are symmetric with respect to those from \( t_0 \) to \( t_3 \), the mode analysis was only performed for the first half-cycle.
All the circuit elements, including the magnetizing inductance of transformer, were assumed to be ideal to simplify the analysis. A detailed derivation including the magnetizing inductance is presented in the Appendix.

Mode 1 [t₀ ≤ t ≤ t₁]: The current through the transformer resonant inductance \( i_{l_{lk1}}(t) \) and the voltage across the panel capacitance \( v_{Cp}(t) \) are zero just before \( t₀ \). When \( S_{A1}, S_{A2}, \) and \( S_{YG} \) turn on at \( t₀ \), \( L_{lk1} \) and \( C_p \) starts to resonate and \( v_{Cp}(t) \) increases from zero to \( V_S \) under the resonance. \( i_{l_{lk1}}(t), v_{Cp}(t) \) in Mode 1 and the duration time of Mode 1, \( T_{d1} \), are defined as:

\[
i_{l_{lk1}}(t) = \frac{V_S}{2} \sqrt{\frac{C_{eq}}{L_{lk1}}} \sin \omega_0 (t - t_0),
\]

\[
v_{Cp}(t) = \frac{V_S}{2} \left[ 1 - \cos \omega_0 (t - t_0) \right],
\]

where \( T_{d1} = \frac{\pi}{\omega_0} \), \( \omega_0 = \frac{1}{\sqrt{L_{lk1}C_{eq}}} \), \( C_{eq} = \frac{C_2C_p}{C_2 + C_p} \equiv C_p \).

\( C_{eq} \) is the parallel capacitance of the energy recovery capacitor \( C_2 \) and the panel capacitance \( C_p \). However, it can be approximated as \( C_p \) because \( C_p \) is much smaller than \( C_2 \) in the general case.

Mode 2 [t₁ < t ≤ t₂]: \( v_{Cp}(t) \) reaches \( V_S \) at \( t₁ \). Then, \( S_{A1} \) and \( S_{A2} \) turn off with ZCS and \( S_{XS} \) turns on with ZVS. \( v_{Cp}(t) \) is sustained to be \( V_S \) constantly by \( S_{XS} \) and \( S_{YG} \), and \( L_{lk1} \) carries no current.

Mode 3 [t₂ < t ≤ t₃]: At \( t₂ \), \( S_{A1} \) and \( S_{A2} \) turn on again and \( S_{XS} \) turns off. \( C_p \) is discharged under resonance and the energy stored in \( C_p \) is recovered to \( C_1 \) and \( C_2 \). \( S_{YG} \) turns on with ZVS at \( t₃ \) because \( v_{Cp}(t₃) \) is zero. The duration time of Mode 3, \( T_{d3} \), is the same as \( T_{d1} \), and \( i_{l_{lk1}}(t) \) and \( v_{Cp}(t) \) in Mode 3 are similar to the current and the voltage of Mode 1 as:

\[
i_{l_{lk1}}(t) = -\frac{V_S}{2} \sqrt{\frac{C_{eq}}{L_{lk1}}} \sin \omega_0 (t - t_2),
\]

\[
v_{Cp}(t) = -\frac{V_S}{2} \left[ 1 + \cos \omega_0 (t - t_2) \right].
\]

B. Device Count Comparison

Figure 7(a) shows the practical circuit configuration for auxiliary switch driving in Weber’s ERC. Because the switches operate with a floating ground, an additional switch driver is necessary. The dotted line in Fig. 7(a) shows the high-side gate driver for the auxiliary switches, two blocking diodes, and one air-core inductor. The high-side driver also needs some peripheral passive components to implement charge pump. The complex circuitry for gating and energy recovery enhances both of the system size and manufacturing cost.
Fig. 6 Equivalent circuits of each operating mode: operating current flows on the thick solid line.

(a) Mode 1

(b) Mode 2

(c) Mode 3

TABLE I

<table>
<thead>
<tr>
<th>Component</th>
<th>Weber</th>
<th>Proposed</th>
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<tbody>
<tr>
<td>gate driver</td>
<td>High-side gate driver with buffer.</td>
<td>General MOSFET driver.</td>
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<tr>
<td>blocking diode</td>
<td>Eight discrete diodes including the clamp.</td>
<td>Four discrete diodes including the clamp.</td>
</tr>
<tr>
<td>resonant inductor</td>
<td>Air-core inductor.</td>
<td>Leakage inductance of transformer.</td>
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C. Conduction Loss Analysis

Neither the proposed ERC nor Weber’s ERC guarantees ZVS turn-on of auxiliary switches. Assuming that the switching loss differences of the auxiliary switches between both ERCs are negligible, minimizing the conduction loss is critical the power efficiency improvement of the proposed ERC.

Parasitic resistances, such as on-resistance of the auxiliary switch and diode and other equivalent series resistances (ESRs), must be considered to analyze the conduction loss of the ERC. The parasitic components are shown in Fig. 8. If the charging and discharging current is assumed to be ideally sinusoidal, the ratio of conduction losses of the proposed ERC and Weber’s ERC are:

\[
\frac{P_{\text{proposed}}}{P_{\text{Weber}}} = \frac{I_{pk}}{I_{pk}} \left[ R_{ds} + R_L + R_D \right] + 4 \left( V_F / \pi \right) .
\]

where \( I_{pk} \), \( R_{ds} \), \( R_L \), \( R_D \), \( V_F \), \( R_{ow} \) are the peak value of the charging and discharging current, the on-resistance of auxiliary switch, the ESR of the air-core inductor, the on-resistance and forward voltage drop of the blocking diode, and the ESR of each transformer winding, respectively. To make eq. (5) near or larger than unity, we recommend reducing the value of \( R_{ow} \) by sandwich-winding the Litz wire in the transformer. Table II shows the device parameters used to verify the proposed ERC operation with the load parameters equivalent to practical circuit ones for typical 42 inch standard-definition (SD) PDP TV sets.

TABLE II

<table>
<thead>
<tr>
<th>Device Parameters of Experimental System</th>
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<tbody>
<tr>
<td>Switch</td>
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<td>Transformer</td>
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<tr>
<td>Resonant Inductor</td>
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</tr>
<tr>
<td>Panel Capacitance</td>
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<tr>
<td>Sustain Voltage</td>
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</table>
IV. EXPERIMENTAL RESULTS

To validate the operation of the proposed ERC, a prototype sustain driver with device parameters shown in Table II was built and evaluated. Fig. 9(a) shows the experimental waveforms of the current in the resonant inductance and the voltage across the x-electrode of panel capacitance at a condition of 200 kHz switching frequency. Both the rising time and falling time of the panel voltage are approximately 400 nanoseconds. Switching losses of the main full bridge were reduced by ZVS turn-on of the main switches, as shown in Fig. 9(b).

Figure 10 shows the measured input current drawn by the same sustain voltage supply according to the various numbers of sustain pulses. The proposed ERC shows power consumption similar to Weber’s ERC within the tolerance of measurement error. It is expected that the efficiency of the proposed ERC would be better than a conventional circuit if we included gate driving loss due to the smaller number of gate-driving ICs.

V. TWO-LEVEL SUSTAIN DRIVER

The proposed ERCs can be applied to two-level sustain driver. Two-level driver has a simultaneous energy-recovery operation of the pair of each ERC leg, which can be integrated into single one as shown in Fig. 11(a). The voltage waveform across PDP has only two transient level, Vs and –Vs, as shown in Fig. 11(b). Two-level drivers are very suitable to very high-frequency sustain drivers since the ERC can reduce the transition time of the panel polarity due to the simultaneous recovery of the reactive energy on both side of the panel [26].
The simplified ERC has only single leg, which results in the half of part count and manufacturing cost of the ERC, including an expensive high-side gate driver. The operating principle of the two-level is the same as the aforementioned three-level operation except the operating mode overlap of mode 1 with mode 6 and of mode 3 with mode 4. Clamp circuit is also employed in the ERC as shown in Fig. 11(a) for preventing instant voltage spikes caused by the discontinuous current on resonant inductor.

![Circuit Diagram](image)

(a) Circuit Diagram

![Key waveforms](image)

(b) Key waveforms

Figure 11. Two-level sustain driver with a switched transformer. The ERC can be more simplified into single leg, which leads to the reduction of part count such as a high-side gate driver.

VI. CONCLUSION

In this paper, a new ERC for an AC PDP sustain driver using a switched transformer has been proposed. The proposed ERC overcame the problems of conventional circuits, such as high device count and high manufacturing cost, especially for eliminating high-side drivers. A transformer with a magnetic core replaced the high-side drivers, blocking diodes, and air-core inductor to reduce circuit complexity and the device count.

Analysis of operating principles and the power loss of the proposed ERC were presented. To verify the results of the analyses, prototype hardware for a 42 inch PDP sustain driver was built and evaluated. With the small number of part count such as a transformer and two MOSFETs, the proposed ERC performed the energy recovery in an operating principle similar to the conventional Weber scheme, and achieved the soft-switching operation of the main switch with comparable power consumption.

Extensive ERC version for two-level sustain driver has also been presented. The ERC has single resonant leg, integrated from the two legs of the three-level version. The circuit is very cost-competitive for consumer electronics products in a capacitive-load characteristic such as plasma display.

APPENDIX

- Derivation of the sustain voltage equation

During the charging and discharging transients of panel capacitance such as in Mode 1 and Mode 3, the transformer is involved in the circuit operation. Therefore, the effect of magnetizing inductance must be considered. To simplify the derivation process, parasitic resistances are neglected and the turn’s ratio of the transformer is assumed to be exactly 1:1. The equivalent Laplace resistances are neglected and the turn’s ratio of the transformer is assumed to be exactly 1:1. The equivalent Laplace circuit in Mode 1 is shown in Fig 12. The sustain voltage $V_{cp}(s)$ and the circuit impedance $Z$ are expressed as follows:

$$ s = \text{complex frequency} $$

$$ L_m = \text{magnetizing inductance} $$

$$ L = L_{a1} $$

$$ C = C_1 = C_2 $$

$$ V_{cp}(s) = \frac{V_s}{s} \left( \frac{1}{sC} + Z \right) \left( \frac{1}{sC_p} \right) $$

$$ Z = \frac{s^2 L_p + \frac{L}{C_p}}{s^3 L_m C_p + s \left( \frac{L C_p + L_m + L}{C_p} \right) + \frac{1}{sC_p}} $$

Substituting eq. (a2) into (a1) and rearranging parameters, $V_{cp}(s)$ has the following form:

$$ V_{cp}(s) = \frac{V_s}{s} \left( \frac{s^2 L_p C}{2s^4 L_m C_p + s \left( \frac{L_m C_p + L_m + L}{2C_p} \right) + s \left( L_m C_p + L \right) + 1} \right) $$

$$ (a3) $$
If $L \ll L_m$ and $C_p \ll C$, then eq. (a3) becomes:

$$V_{C_p}(s) = \frac{V_s}{s} \frac{s^2 L C}{s^2 (2L_m L C) + s^2 (2L_m C + L_p) + 1} = \frac{V_s}{2} \left[ \frac{s^2 L C}{1 + s^2 (2L_m C) + L_p} \right].$$

(a4)

By taking the inverse Laplace transform of eq. (a4), panel capacitance voltage in the time domain will be

$$v_{C_p}(t) = \frac{V_s}{2} \left[ \frac{\cos \frac{t}{\sqrt{2L_m C}} - \cos \frac{t}{\sqrt{L_p}}}{1 + \sqrt{2L_m C} \sqrt{L_p}} \right].$$

(a5)

The resonant period of the first term in eq. (a5) is much longer than that of the second term. If $t$ is very small; i.e., $t \ll 2\pi \sqrt{2L_m C}$, then eq. (a5) can be approximated as:

$$v_{C_p}(t) \approx \frac{V_s}{2} \left[ \frac{1 - \cos \frac{t}{\sqrt{L_p}}}{1 + \sqrt{2L_m C} \sqrt{L_p}} \right],$$

(a6)

which is the same as eq. (2). Therefore, the magnetizing inductance is negligible for a short duration of time such as $T_{d1}$ or $T_{d3}$.

**Figure 12.** Equivalent Laplace circuit in Mode 1 and Mode 3. The turn's ratio of the transformer is assumed to be 1:1.

**REFERENCES**


J. H. Park (M’07) was born in Korea 1975. He received his B.S., M.S., and Ph.D. degrees from Electrical Engineering and Computer Science department of Seoul National University, Seoul, South Korea in 1999, 2001, and 2006, respectively. From 2006 to 2009, he was a post doctor in Department of Electrical Engineering and Computer Science, Seoul National Univ., Seoul, Korea. He joined Department of Electrical Engineering, Soongsil University, Seoul, Korea in 2009 and currently a full-time lecturer. His interests include analysis and design of high-frequency switching converter, renewable energy systems, and piezoelectric-transformer power applications, etc.

Jongwon Shin received the B.S. degree in electrical engineering from Seoul National University, Seoul, Korea in 2006, where he is currently working toward his Ph.D. degree in the School of Electrical Engineering. His research interests are in the areas of plasma display panel driver, high-efficiency power converter and power factor correction circuit. Mr. Shin is a Student Member of the Korean Institute of Power Electronics.

Woosup Kim received the B.S. degree in electrical engineering from Kwangwoon University, Seoul, Korea in 2009 and currently a full-time lecturer. His interests include analysis and design of high-frequency switching converter, renewable energy systems, and piezoelectric-transformer power applications, etc.

B. H. Cho (M’89-SM’95) was born in Korea 1952. He received the B.S. and M.E. degrees in electrical engineering from California Institute of Technology, Pasadena, and the Ph.D. degree, also in electrical engineering, from Virginia Polytechnic Institute and State University (Virginia Tech), Blacksburg. Prior to his research at Virginia Tech, he worked for two years as a member of the technical staff of the Power Conversion Electronics Department, TRW Defense and Space System Group, where he was involved in the design and analysis of spacecraft power processing equipment. From 1982 to 1995, he was a professor in the Department of Electrical Engineering, Virginia Tech, Blacksburg, Virginia. He joined the school of Electrical Engineering, Seoul National University, Seoul, Korea in 1995 and he is presently a professor. His main research interests include power electronics, modeling, analysis and control of spacecraft power processing equipment, power systems for space station and space platform, and distributed power systems.