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Dual Path Magnetic-Coupled AC-PDP Sustain Driver with Low Switching Loss

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ABSTRACT

A cost-effective magnetic-coupled AC-PDP sustain driver with low switching loss is proposed. The transformer reduces current stress in the energy recovery switches which affects circuit cost and reliability. The turns-ratio can be used to adjust the sustain pulse slopes which affect gas discharge uniformity. Dividing the recovery paths prevents abrupt changes in the output capacitance and thereby switching losses of the recovery switches is reduced. In addition, the proposed circuit has a more simple structure because it does not use the recovery path diodes which also afford a large recovery current. By reducing the current stress and device count in the energy recovery circuit, the proposed driver may have decreased circuit cost and improved circuit reliability.

Keywords: Sustain Driver, Energy recovery, Plasma Display Panel(PDP)

1. Introduction

Since AC-PDP was invented at the University of Illinois in 1964, PDP technology has made remarkable developments. It successfully entered the Flat Panel Display (FPD) market thanks to its attractive merits such as wide view angle, large screen, high brightness, and thinness. The PDP market-share has become more widespread in the large size display market, but its cost is still too high to compete with other FPDs. In the PDP field, this is the most important issue left to be solved [1]. PDPs have developed remarkably based on a three-electrode surface-discharge type cell structure. Fig. 1 shows the simplified PDP structure and its driving method which is ADS (Address Display Separation).

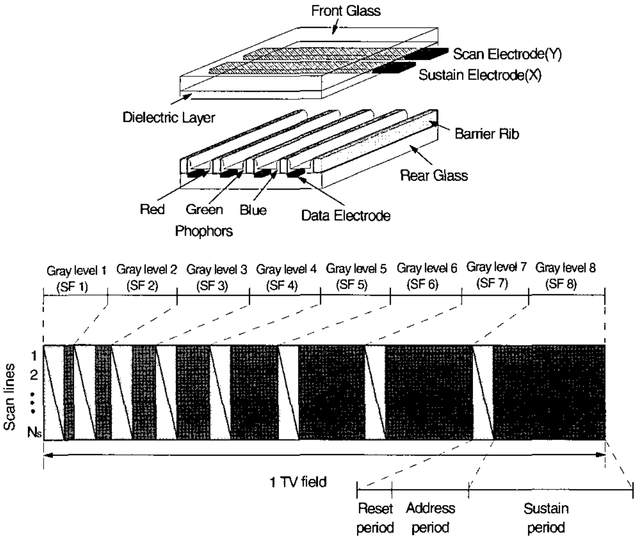

Fig. 1 Simplified PDP cell structure and ADS driving method

Bus electrodes including scan(Y) and sustain(X) electrodes are built on the front glass and data electrodes are built on the rear glass. A dielectric layer is encrusted

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on bus electrodes to limit discharge current while phosphors are pasted on data electrodes to realize colors, which make inherent capacitances among the three electrodes [2]. The operation of the PDP is divided into three periods: setup, addressing, and sustaining. Brightness information is defined by sustain pulse number. During the setup period, all of the PDP cells are erased and prepared to carry out addressing by forming adequate wall charges. Then addressing selective discharges from an image are ignited by applying data and scan pulses to the data and scan electrodes respectively [3]. Since addressing the discharge itself emits insufficient light, high voltage AC square pulses are continuously applied between the sustain and scan electrodes for strong light emission of selected cells. As a result, most of the PDP power is consumed during this sustain period. The high voltage pulses can be simply generated using a full-bridge inverter as shown in Fig. 2.

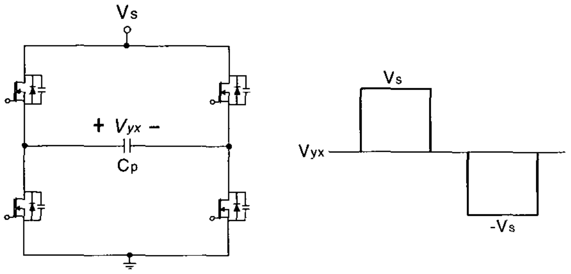


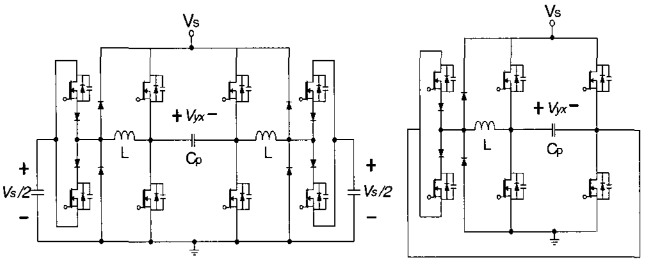
Fig. 2 Basic full-bridge sustain inverter and its wavefom

Due to panel capacitance, a large amount of energy should be dissipated through the switches while a parasitic resistance of the wire occurs during charge and discharge transients. If an average frequency sustain pulse is f_{av}, then the total dissipated power is $2C_pV_s^2f_{av}$ where C_p is panel capacitance and V_s is sustain voltage [4]. Without proper methods to recover the energy, many problems such as heat dissipation of switching devices and EMI emission occurs in addition to excessive power loss. To solve these problems, L. F. Webber, et al. suggests an energy recovery circuit (ERC) using a series-resonant concept [5]. It features high efficiency and good circuit flexibility to cope with various driving methods, which has led many PDP makers such as Samsung, LG, Matsushita, and FHP to adopt this circuit. M. Ohba, et al. have reduced this circuit supporting parallel-resonance and it has been adopted by NEC [6]. PDP power consumption is controlled by changing the total sustain number according to displayed images due to the low panel efficiency. Consequently, the energy recovery circuit of the sustain driver suffers from heavy current stress in cases where maximum sustain pulse number is applied to get maximum brightness in a minimum display area. It results in heat problems, cost increase, and difficulty in component selection. To reduce the current stress of the recovery circuit, a new concept sustain driver using a magnetic-coupled technique was suggested [7]. The net current of recovery switches can be reduced using the energy recovery current reflected through the transformer. An additional advantage is the sustain pulse slope is adjusted by the turns-ratio of the transformer and it can be used for improving discharge uniformity. Unfortunately it suffers from heavy switching loss in the recovery switches and still has a bulky circuit structure. This paper introduces an improved magnetic-coupled method with dual-recovery paths. The operation is similar to the previously suggested circuit with a three-winding transformer. The proposed circuit has respective recovery paths according to panel capacitance charge and discharge operations. Additionally, it prevents the abrupt change of output capacitance charge current so that the switching loss of recovery switches can be reduced. By modifying the switch control timings, normal operation can be guaranteed despite of the removal of the bulky recovery path diodes. This makes the proposed circuit simple. The proposed method has been designed and validated for 42" SD PDPs(852×480).

2. Series and Parallel-Resonant Approaches

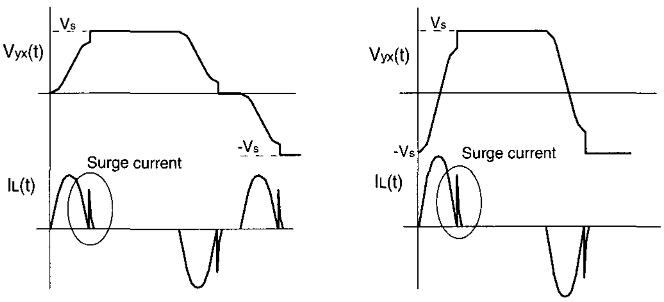
Fig. 3 shows prior approaches using LC resonance and in Fig. 4. their sustain voltage waveforms, applied across panel electrodes, are shown. The sustain voltages go up or ramp down in a resonant manner and the sustain driver switches are turned on to hold a sustain voltage or GND. Figure R is parasitic resistance including the on-resistances of the switches and V_{on} is the on-drop voltage. Based on this figure, the panel voltage V_{yx} can be obtained as follows:

- *Series-resonant type*:



(a) Series-resonant circuit (b) Parallel-resonant circuit

Fig. 3 Prior approaches for AC-PDP sustain driver



(a) Series-resonant circuit (b) Parallel-resonant circuit

Fig. 4 Panel voltage waveforms of prior approaches

$$V_{yx} = \left(\frac{V_s}{2} - V_{on}\right)\left[1 - e^{-t/\tau}\cos\omega_p t\right] \qquad (1)$$

- *Parallel-resonant type*:

$$V_{yx} = -(V_s - V_{on})e^{-t/\tau}\cos\omega_p t - V_{on} \qquad (2)$$

where $\omega_p = 1/\sqrt{LC_p}$ and $\tau = 2L/R$.

Reducing the parasitic components is essential for minimizing hard-switching stress and improving recovery performance. An optimally designed circuit board includes switching devices with small on-resistance and low on-drop voltage. However, it is impossible to get rid of parasitic components completely. Therefore EMI and switching stress caused by the surge current is inevitable [8]. Although the parallel-resonant method has a simpler circuit structure, the circuit costs of the two methods are similar because they require low impedance interconnections across the X and Y drivers. Therefore most PDP makers employ a series-resonant circuit that can utilize various driving waveforms. In the case of 42"PDPs, about 40 semiconductor devices are used to construct a sustain driver because discharge current and recovery current show pulsating waveforms with large peak values over 150A and 60A respectively. It is an important cause of cost increase and power loss in PDP driving system.

3. Magnetic-Coupled Sustain Driver

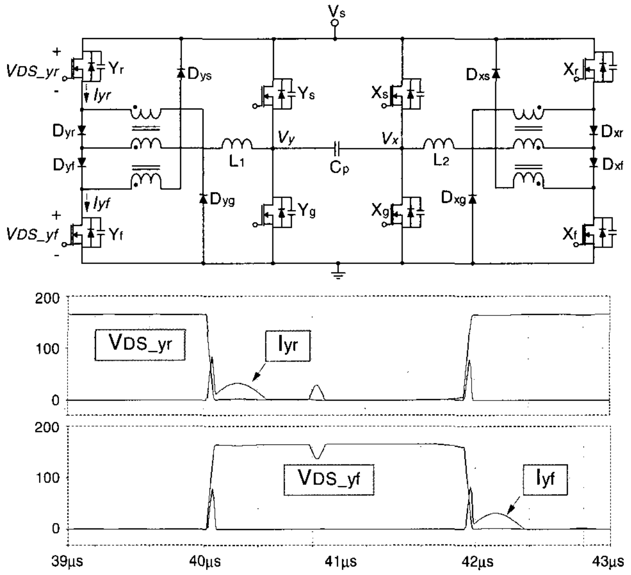


Fig. 5 Magnetic-coupled sustain driver and its simulated switching waveforms of energy recovery switches

Fig. 5 is the magnetic-coupled PDP sustain driver [7]. The driver is composed of eight switch blocks respectively. Y_s and X_s are for applying a high sustain voltage Vs to each electrodes and GND level is connected through Y_g and X_g. Y_r, X_r and Y_f, X_f are in the charge of delivering energy between power source and panel capacitance. Because it does not use the recovery capacitor bank and the recovery switches are directly connected to V_s and GND, the energy charged in the panel capacitance is recovered to sustain voltage source by way of the transformers. During the recovery operation, the reflected resonant current flows through the body diodes of the recovery switches. This reduces the switch current stress, and the amount of reduced current is determined by the transformer turns-ratio. However, it suffers from heavy switching loss of the recovery switches. As shown in the recovery switch waveforms of Fig. 5, a large surge current to charge the output capacitance of the recovery switches flows from Y_r and Y_f when the recovery switches are turned on and it causes the switching loss of the recovery switches.

Because the sustain pulse number varies according to the displayed images, it becomes worse in cases where the maximum number of sustain pulses are generated. To solve this problem an improved magnetic-coupled method with dual recovery paths is proposed to prevent switching loss, which is shown in Fig. 6.

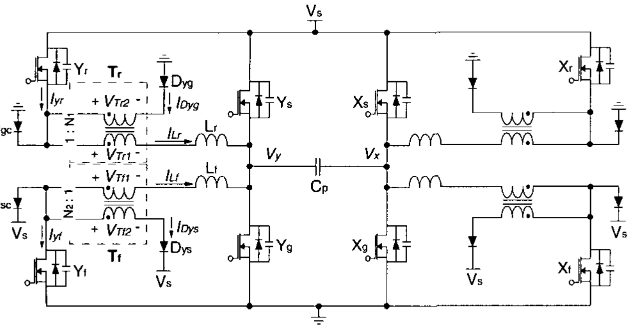


Fig. 6 Proposed sustain driver with dual recovery paths

The charge current of the output capacitance detours through the resonant inductor and the transformer between the recovery switches so that abrupt current changes can be prevented. Another advantage of the proposed method is the circuit structure can be simplified by reducing the semiconductor devices. In conventional 42" PDP sustain drivers, 16 diodes with a 20A rating have been used for recovery path diodes to handle large resonant currents. The recovery path diodes, that occupy considerable circuit volume and cost, can be removed by modifying the recovery switch operation.

4. Mode Analysis

Key waveforms of the proposed circuit are depicted in Fig. 7. Fig. 8 shows the operational mode diagrams. In Fig. 8 the dotted arrows describe the magnetizing current path and the solid arrows describe the energy recovery path. The operation is divided into 6 modes. Mode analysis was only performed during the first half cycle because the operation of the two half cycles is symmetric. Before the detailed analysis, it is assumed that the transformers have no parasitic components without magnetizing inductance and their secondary turns-ratios of N_1 and N_2 are assumed to be more than 2, which will be discussed in section V.

Mode 1 ($t_0 \leq t < t_1$) : Mode 1 begins at t_o when Y_r is turned on.

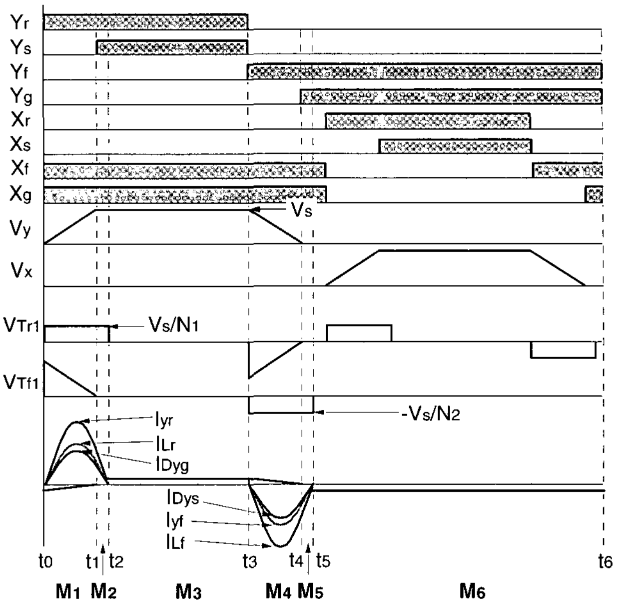


Fig. 7 Key waveforms of the proposed method

The current path forms to include Y_r, primary windings Tr, L_r, and C_p in sequence. Because the dotted end of the transformer T_r is positive, diode D_{yg} begins to conduct and another current path is formed through D_{yg}, secondary winding T_r, and body diodeY_r. Because V_{Tr2} is clamped by V_s, the reflected voltage of primary winding V_{Tr1} is determined as V_s/N_1. The inductor current I_{Lr} flows in a resonant manner and Y electrode voltage V_y increases. The expressions are

$$I_{Lr} = \left(1 - \frac{1}{N_1}\right)\frac{V_s}{Z_r}\sin\omega_r(t - t_o) \tag{3}$$

$$V_y = \left(1 - \frac{1}{N_1}\right)V_s(1 - \cos\omega_r(t - t_o)) \tag{4}$$

where $\omega_r = 1/\sqrt{L_r C_p}$ and $Z_r = \sqrt{L_r / C_p}$. Since the reflected current I_{Lr} scaled by $1/N_1$ flows through the body diode Y_r, the net current Y_r is $(1\text{-}1/N_1)I_{Lr}$. This reduces current stress and power loss Y_r. During mode 1, primary voltage T_f, V_{Tf1} is applied by V_s-V_y so that T_f becomes completely reset.

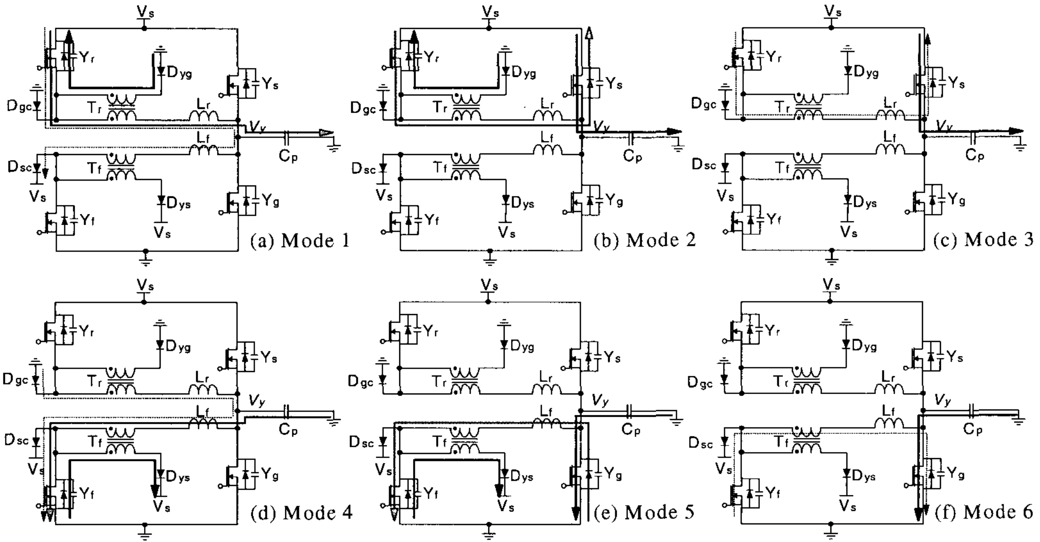


Fig. 8 Operational mode diagrams

Mode 2 ($t_1 \le t < t_2$) : When V_y goes up to V_s, body diode Y_s is turned on. I_{Lr} starts to linearly decrease through Y_r, the primary windings T_r, L_r, and body diode Y_s, while D_{yg} keeps conducting. It is written as

$$I_{Lr} = I_{Lr}(t_1) - \frac{V_s}{N_1 L_r}(t - t_1) \tag{5}$$

If Y_s is turned on to hold a sustain voltage during this mode, zero-voltage-switching (ZVS) is accomplished, which gives helps reduce the switching loss. This mode continues until I_{Dyg} is reduced to zero.

Mode 3 ($t_2 \le t < t_3$) : The gas discharge to emit visible light is induced and this discharge current is supplied through Y_s. During this mode, only magnetizing current T_r remains and it requires a circuit path to ensure the flow of the magnetizing current. This can be accomplished by keeping recovery switch Y_r turned on together with Y_s. Therefore the small magnetizing current can flow through Y_r, primary windings T_r and L_r, and body diode Y_s. This has another advantage in that the bulky path diode can be removed to prevent abnormal operation. Fig. 9 shows circuit operation cases when the recovery switch is turned off after V_y is increased to V_s. The magnetizing inductance of T_r and recovery inductor L_r become resonated with the output capacitance of Y_r through primary windings T_r and L_r, and body diode Y_s as shown in path A. When the resonant operation has ended, the output capacitance of Y_r discharges completely and body diode Y_r is turned on. Accordingly, the current freewheels through path B because Y_s continues conducting until V_y falls to $(N_1\text{-}1)V_s$, that is, V_{Tr2} reaches V_s. When this condition occurs, D_{yg} begins to conduct and I_{Lr}, which has the value $(N_1\text{-}1)I_{Dyg}$, is built up through path C.

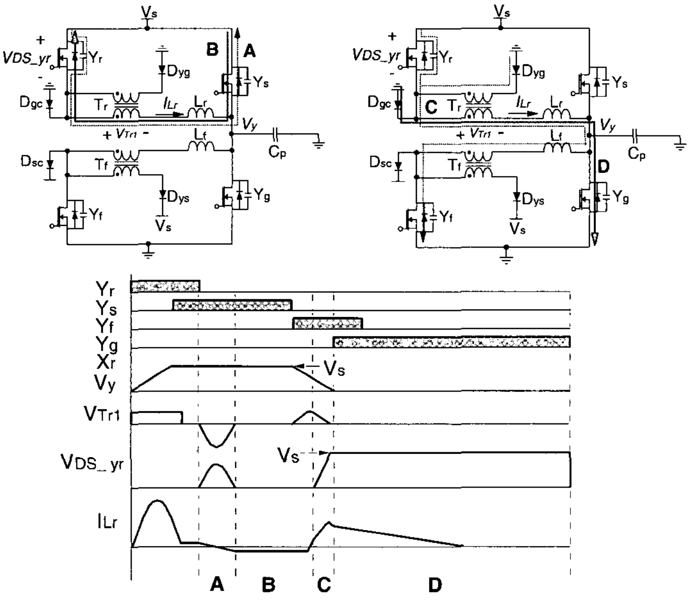


Fig. 9 Circuit operation in case that the recovery switch is turned off after V_y is increased to V_s

This current increases until the output capacitance of Y_r is charged up to V_s. After the drain-source voltage of Y_r

reaches V_s, the abnormal build-up of current starts to flow through D_{ygc}, the primary windings T_r and L_r, and Y_g as shown in path D. It is decreased slowly, dissipating heat through parasitic resistances distributed in the freewheeling path. This undesirable power dissipation happens during the pulse-falling operation as well as during the pulse-rising operation. To avoid this abnormal case, it is necessary to block current path B by installing bulky path diodes. However, the proposed method can remove the abnormal operation by using recovery switches as a magnetizing current path.

Mode 4 ($t_3 \le t < t_4$) : From mode 4 the charged energy in the panel capacitance begins recovery. When Y_f is turned on at t_3, a current path forms through C_p, L_f, and primary windings T_f, Y_f. Because positive voltage is applied to the undotted end of the primary winding, D_{ys} conducts and V_{Tf2} is clamped by V_s through the body diode of Y_f and D_{ys}. This induces V_{Tf1} to have voltage V_s/N_2. It follows that the inductor current I_{Lf} flows in a resonant manner from the panel and the Y electrode voltage V_y starts to decrease. They can be expressed as

$$I_{Lf} = -\left(1-\frac{1}{N_2}\right)\frac{V_s}{Z_f}\sin\omega_f(t-t_3) \tag{6}$$

$$V_y = \frac{V_s}{N_2} + \left(1-\frac{1}{N_2}\right)V_s\cos\omega_f(t-t_3) \tag{7}$$

where $\omega_f = 1/\sqrt{L_f C_p}$ and $Z_f = \sqrt{L_f / C_p}$. The current stress of Y_f is reduced to $(1-1/N_2)I_{Lf}$ in the same manner as Y_r. In the primary voltage of T_r, V_{Tr1} is applied by $-V_y$ and T_r is reset during this mode.

Mode 5 ($t_4 \le t < t_5$) : When V_y is ramped down to GND, the inductor current starts increasing with a slope of $V_s/(N_2L_f)$ through body diodes Y_g, L_f. and primary windings T_y, and Y_f. The expression is

$$I_{Lf} = I_{Lf}(t_4) + \frac{V_s}{N_2 L_f}(t-t_4) \tag{8}$$

The ZVS condition Y_g is also achieved in this mode. Mode 5 continues until L_f and T_y have no currents.

Mode 6 ($t_5 \le t < t_6$) : Similar to mode 3 in that only magnetizing current T_f remains and Y_f continues its conducting state to maintain a magnetizing current path before the new cycle begins.

5. Design Considerations

Gas discharge stability and brightness of PDPs are very sensitive to sustain pulse rising-time T_r which is defined by the time when the sustain voltage increases from GND to V_s. By replacing V_y with V_s in eq. (4), the rising-time is obtained as

$$T_r = \frac{1}{\omega_r}\cos^{-1}\left(\frac{-1}{N_1-1}\right) \tag{9}$$

It shows that T_r can be adjusted by the transformer turns-ratio.

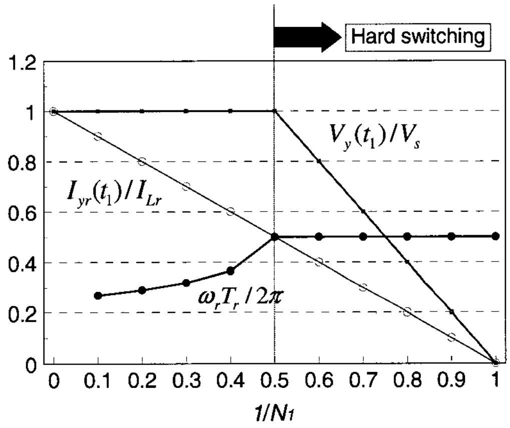


Fig. 10 Current stress and rising-time according to transformer turns-ratio

Fig. 10 is the current stress and rising-time plotted according to $1/N_1$. In cases where $1/N_1$ is greater than 0.5, T_r is defined as a half of the resonant period because the resonant current is not enough for V_y to reach the sustain voltage. In this case, switching loss Y_s occurs while recovery switch current stress is reduced. Conversely, very small values of $1/N_1$ produce a large circulating current because the energy stored in the resonant inductor is too excessive to charge or discharge the panel capacitance. Fig 11 is the recovery current including the resonant current and circulating current. Resonant time T_1, peak resonant current $I_{r,pk}$, and circulation start current $I_{c,pk}$ can be found from eqs. (3) and (4) by replacing V_y with V_s. Circulating time T_2 is calculated with eq. (5) by finding the time when the current is reduced to zero. Referring to Fig. 11 and the mode diagrams, the power dissipated in semiconductor devices from a single recovery operation can be calculated as follows:

- *Power loss due to circulating current during single recovery operation:*

$$P_c = \frac{1}{T_s}\left[V_{on}\left(\frac{(N_1+1)L_r I_{c,pk}^2}{2V_s}\right) + R_{DS\_on}\left(\frac{(N_1-1)L_r I_{c,pk}^3}{3V_s}\right)\right] \quad (10)$$

- *Power loss due to resonant current during single recovery operation:*

$$P_r = \frac{1}{T_s}\left[\frac{V_{on} I_{r,pk}}{\omega_r}\left(\frac{1}{N_1-1}\right) + \frac{R_{DS,on} I_{r,pk}^2}{\omega_r}\left(1-\frac{1}{N_1}\right)^2 \times\left(\frac{1}{2}\cos^{-1}\left(\frac{-1}{N_1-1}\right) + \frac{\sqrt{1-2/N_1}}{2(N_1-1)}\right)\right] \quad (11)$$

where $R_{DS,on}$ describes the on-resistance of the recovery switches and T_s is the sustain pulse period.

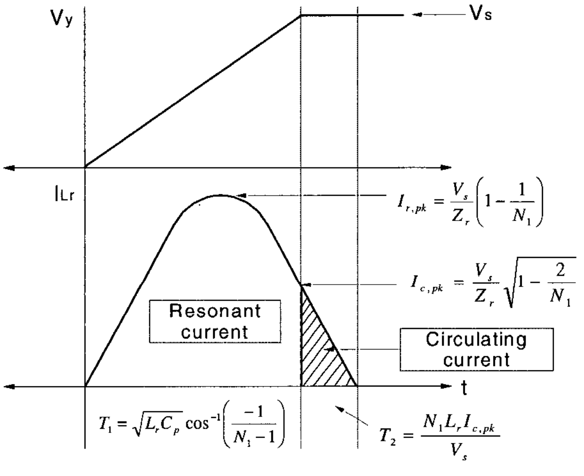


Fig. 11 Current waveform during pulse-rising period

Because the falling-time of the sustain pulse does not affect the discharge if it is too short, the same rising-time value is generally used. With this assumption, the average device power loss of the proposed sustain driver during one TV field period can be written as

- *Power loss due to circulating current during one TV field period:*

$$P_{c,TV} = 2P_c \times \frac{N_p T_s}{T_{TV\_field}} \quad (12)$$

- *Power loss due to resonant current during one TV field period:*

$$P_{r,TV} = 2P_r \times \frac{N_p T_s}{T_{TV\_field}} \quad (13)$$

where N_p is the sustain pulse number and T_{TV_field} is one TV field period. These equations suggest another guideline for selecting circuit parameters. Therefore, it is necessary to select an appropriate value of turns-ratio by considering discharge characteristic, current stress, and power loss.

6. Experimental Results

To validate the proposed method, a sustain driver was designed for a 42" SD PDP with the following specifications:

- ▸ Sustain voltage : V_s = 180V
- ▸ Sustain pulse switching frequency : f_s= 200kHz
- ▸ Panel capacitance : C_p = about 80nF
- ▸ Maximum sustain pulse number : $N_{p,max}$=2300

Using eqs. (12) and (13) and the parameters of the selected devices, device power losses according to turns-ratio can be plotted as shown in Fig. 12.

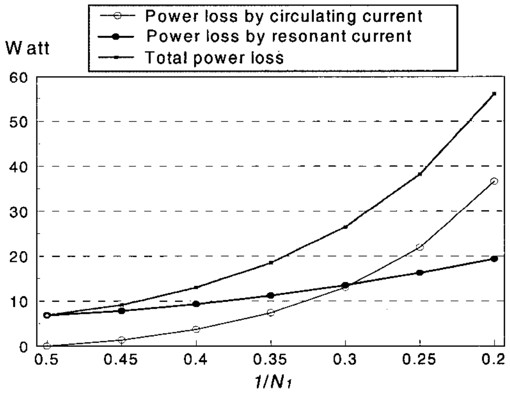


Fig. 12 Calculated device power loss plots according to transformer turns-ratio at sustain pulse number = 2300, V_s = 180V

After considering the discharge stability and power loss of the devices, the transformer turns-ratio is selected as

0.43 and the transformer is constructed with a PQ2625 ferrite core of TDK. The leakage inductance of the transformer is used as a resonant inductor and it is approximately 0.2μH. The prototype driver is implemented with two IXYS 62N25(62A/250V) in parallel to enable the sustain driver to conduct high discharge currents over 150A. One 2SK2995(30A/250V) can be used for the recovery switches due to the reduced current stress. Fig. 13 is the measured waveforms of the proposed driver. It shows that the current stress of Y_r, which is about 60% of the resonant current and the primary voltage of the transformer, concurs with the theoretical analysis. In addition, the clamping current is negligibly small in the proposed method because of the large value of the magnetizing inductance compared with the resonant inductance of the series connected inductor of the conventional method. This may help to suppress the power consumption caused by the freewheeling current.

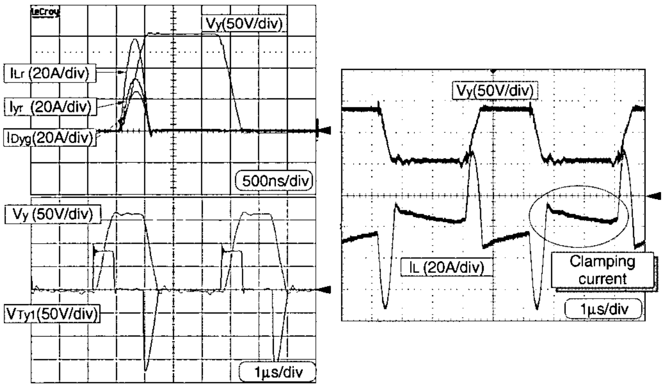


Fig. 13 Measured waveforms of the proposed method and series-resonant method

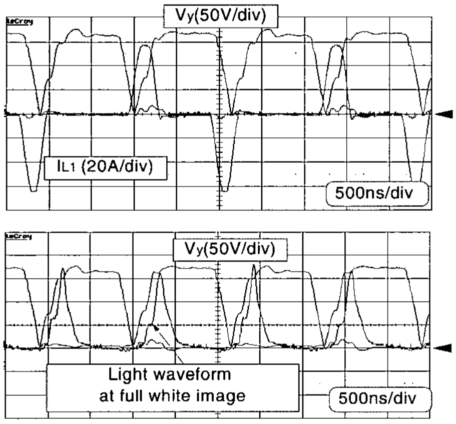


Fig. 14 Light and recovery current waveforms at full white image

As shown in Fig. 14, light emission for a full white image requires a stable maximum discharge current.

Fig. 15 is the enlarged current flowing through Y_r. It shows that no surge currents occur when the recovery switch is turned on and helps to reduce the switching stress. By removing gas discharge to investigate the circuit performance itself, the measured driver loss including transformer and driving circuitry is approximately 17W at maximum sustain number, which is a 70% reduction compared with previous the magnetic-coupled method.

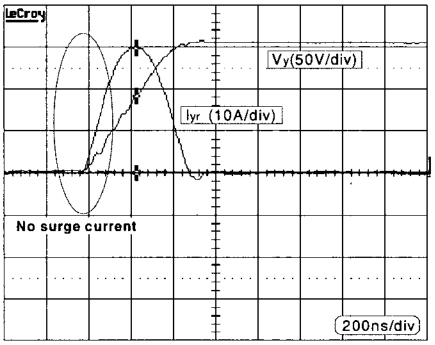


Fig. 15 Current waveform of recovery switch Y_r

Table 1 is the component list comparisons with the series-resonant driver circuit in cases of 42" SD PDP, which shows that the device count of high capacity semiconductors is decreased.

Table 1 Component list comparisons

	Proposed driver	Series-resonant driver
Sustain inverter switches	IXYS 62N25×8	IXYS 62N25×8
Recovery switches	2SK2995×4	2SK2995×8
Diodes	S20LC40×4 S2FL20U×4	SF20LC30×20
Transformers	2	-

7. Conclusions

The proposed method suggests an improved version of a magnetic-coupled method with dual recovery paths. By preventing an abrupt change of output capacitance charge current, by detouring it through components installed in a recovery path, the switching loss of the recovery switches can be reduced. Semiconductor device count can be reduced to 45% compared with the conventional series-resonant method. By changing the recovery switch operation it is possible to maintain its advantages such as recovery current cancellation by transformer and low clamping current. Test results on a 42" SD PDP show more than 70% of circuit loss can be reduced at maximum sustain number compared with the previous magnetic-coupled method and power loss is equivalent to the series-resonant method. Therefore, the proposed method may be another candidate for low cost PDP sustain drivers.

Acknowledgment

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