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고효율의 PDP 유지 구동 전원단을 위한 새로운 펄스폭 제어방식의 쿼지 공진 컨버터

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A New PWM-Controlled Quasi-Resonant Converter for High Efficiency PDP Sustaining Power Module

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ABSTRACT

A new PWM-controlled quasi-resonant converter for high efficiency PDP sustaining power module is proposed in this paper. The load regulation of the proposed converter can be achieved by controlling the ripple of the resonant voltage across the resonant capacitor with bi-directional auxiliary circuit, while the main switches are operating at the fixed duty ratio and fixed switching frequency. Hence, the waveform of currents can be expected to be optimized on the conduction loss. Furthermore, the proposed converter shows the good ZVS capability, simple control circuits, no high voltage ringing problem of rectifier diodes, no DC offset of the magnetizing current and low voltage stress of power switches. In this paper, operational principles, analysis and design considerations are presented. Experimental results demonstrate that the output voltage can be controlled well by the auxiliary circuit as PWM method.

1. INTRODUCTION

A plasma display panel (PDP) has been considered as the best candidate for flat panel display because of its wide view angle, high contrast ratio, and long life time. Due to the existence of dielectric layer in each cell of PDP, it is purely capacitive load in respect to circuit operation [1]. Since, not only it features pure capacitive load characteristics but also it is driven by the address display separation (ADS) method, the load variation of the sustaining power module is very wide and abrupt. In ADS method, the operation of PDP can be divided into three periods such as resetting, addressing, and sustaining period. In real PDP TVs, the load condition is strongly dependent on the average pixel level (APL)such as a concept that defines the total light output of a given TV image as a percentage of the total light output of a full-white image. Since TV signals typically have an APL of 20% or less [4], the sustaining power module is usually operating on light load conditions. Although the sustaining power module is operating such load conditions, the power dissipated during sustaining period is still the most of power driving the PDP compared with that dissipated during resetting and addressing periods. Therefore, the sustaining power module is mainly responsible for overall system efficiency [1-3]. In addition, when the PDP is operating on TV signals, the high efficiency is needed significantly at light load conditions. Until now, several DC/DC converters which can realize the high efficiency and low cost, have been proposed for sustaining power module of PDP. Among them, the resonant converters have been investigated to realize prominent features of miniaturization, high efficiency, and low noise.[6,7] However, since a large variation of switching frequency is needed to control the output voltage, those converters have some difficulties from the view-points of size reduction and noise problem.[8] To overcome above problems, recently a half bridge LLC resonant converter has been discussed because it shows many unique characteristics and improvements over previous topologies.[4,5] However, this converter has a small magnetizing inductor in order to have a narrow variation of switching frequency. This results in not only considerably higher current stresses of primary power switches, but also more conduction losses especially at the above resonance mode. In addition, a variable frequency control method makes the control circuits much more complicated than those of the PWM control method. To resolve these problems effectively, we propose a new PWM-controlled quasi-resonant converter which has the simpler control circuits and less conduction losses compared with those of half bridge LLC resonant converter. As shown in Fig. 1, the proposed converter is similar to the half bridge LLC resonant converter except to the auxiliary circuit which is needed to control the output voltage. In the proposed converter, the output voltage can be regulated by controlling the voltage across the resonant capacitor while two main switches are operating at fixed duty ratio and fixed switching frequency. Therefore, the shape of both primary and secondary current can be expected to be optimized on conduction loss and current stress. Simultaneously, since the auxiliary circuit controls the voltage ripple on the resonant capacitor, the output voltage can be regulated well at whole load condition. Thus, keeping the good characteristics of half bridge LLC resonant converter, the proposed converter is expected to overcome effectively the above mentioned problems such as the higher current stresses and high circulating energy, and can realize the high power density, high performance, and high efficiency.

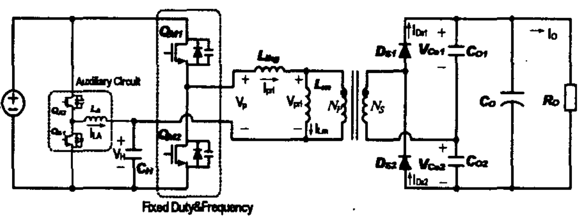

Fig. 1 Circuit diagram of proposed converter

2. OPERATIONAL PRINCIPLES

Fig. 2 shows the key waveforms of the proposed converter. The operation of the proposed converter can be divided into ten modes. Since the operational principles of two half cycles are symmetric, only the first half cycle is explained. A half cycle can be divided into 5 modes. To illustrate the steady state operation, several assumptions are made as follow:

■ QM1, QM2, QA1, and QA2 are ideal except for their internal diodes and output capacitors.

■ DS1 and DS2 are ideal except for their junction capacitors.

■ Vo is constant during a switching period.

After the ZVS turned on of QM1 is achieved and the commutation between DS1 and DS2 is completed at t0, The primary current Ipri, which raises with resonance between the leakage inductor and resonant capacitor, is given by

$$I_{pri}(t) = \frac{1}{nZo}\left[\frac{V_S}{n} - \frac{V_H(t_0)}{n} - V_{Co1}(t_0)\right]\sin\omega_r(t - t_1) + I_{Lm}(t) \quad (1)$$

Concurrently, the magnetizing current, ILm, also raises as follows:

$$I_{Lm}(t) = I_{Lm}(t_1) + \frac{nV_{Co1}(t)}{L_m}(t - t_1) \quad (2)$$

where, $\omega_r = \frac{1}{\sqrt{L_r C_r}}$, $Zo = \sqrt{\frac{L_r}{C_r}}$, $L_r = \frac{L_{lkg}}{n^2}$, $C_r = \frac{n^2 C_H \times C_{o1}//C_{o2}}{n^2 C_H + C_{o1}//C_{o2}}$, $n = \frac{N_p}{N_s}$.

The current of rectifier diode DS1, IDS1, is flowing through Coland equivalent load resistor, while the rectifier capacitor Col and Co2 is charged and discharged respectively. When QA2 is turned on at tl, the resonant capacitor, CH, is additionally charged from the input source, VS, through the auxiliary inductor LA operating in the discontinuous conduction mode (DCM). After QA2 is turned off at t2, ILA starts to charge and discharge the output capacitors of QA2 and QA1 respectively. When the voltage across QA1 becomes 0V, ILA begins to flow through the internal diode of QA1. Since the voltage across the CH, VH is applied to LA reversely, ILA is decreased. During t2~t3, CH is still charged. After QM1 is turned off at t3, since Ipri starts to charge and discharge the output capacitors of QM1 and QM2 respectively, the voltage across the primary side of the transformer VP, is decreased to - VS. Since the rectifier diode DS1 is still conducting, the voltage across Lm, Vpri is maintained to be nVCol. Thus, the negative voltage which is the same as the difference between VP and Vpri, is applied to the leakage inductor, Llkg. Thereby Ipri is decreased rapidly. Also, in t3~t4, CH is continuously charged until ILA becomes 0A. After the voltage across QM2 becomes 0V, Ipri starts to flow through the internal diode of QM2. Thus, the ZVS condition of QM2 is satisfied. When Ipri is smaller than ILm, the current of secondary side of the transformer flows reversely through the junction capacitors of DS1 and DS2. Since the voltage across each diode is increased and decreased respectively, the commutation between DS1 and DS2 is started. During t4~t5, the ZVS turned on of QM2 can be achieved. After DS2 is fully conducting, first half cycle is finished.

3. ANALYSIS & DESIGN CONSIDERATION

During the half switching period TS/2, the charging current of Col is equal to the load current IO. In addition, since the sum of VCol and VCo2 is always equal to the output voltage VO, the bias voltage of VCol and VCo2 is VO/2. Thus, by using the equation IC=C(dVC/dt), VCol(t0) can be obtained. Also, since both QM1 and QM2 are operating with the constant duty ratio (D=0.5), the bias voltage of VH is the same as VS/2. Thus, VCol(t0) and VH(t0) are represented respectively as follows:

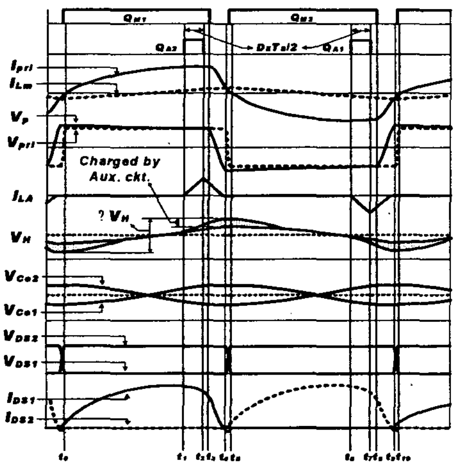

Fig. 2 Key waveforms of proposed converter

$$V_{Co1}(t_0) = \frac{V_O}{2} - \frac{T_S I_O}{4C_{o1}}. \quad (3) \qquad V_H(t_0) = \frac{V_S}{2} - \frac{\Delta V_H}{2}. \quad (4)$$

By using the equations (3) and (4), the current through DS1, IDS1 can be easily obtained as follows:

$$I_{DS1}(t) = \frac{1}{Z_O}\left[\frac{V_S}{n} - \frac{1}{n}\left(\frac{V_S}{2} - \frac{\Delta V_H}{2}\right) - \left(\frac{V_O}{2} - \frac{T_S I_O}{4C_{o1}}\right)\right]\sin\omega_r t \quad (5)$$

Turns ratio of the transformer can be determined at no load condition in order to maintain VO to be desired value without the operation of auxiliary circuit. Thus, it can be obtained by using the equation (5) with these conditions such as IO=0A, IDS1=0A, ΔVH=0V, and $\sin\omega_r t \neq 0$.

$$n = \frac{V_S}{V_O}. \quad (6)$$

Since both the charging current of Co1 and the discharging current of Co2 flow through DS1, the averaging value of IDS1 during TS/2 is equal to the two times of IO. From this fact, IO can be represented as follows:

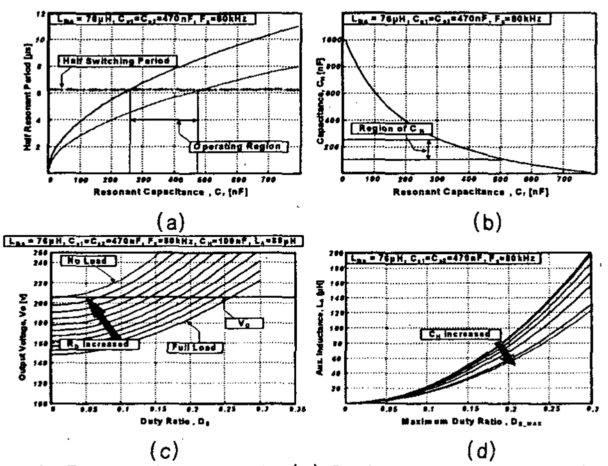

Fig. 3 Figures for analysis,(a) Desirable operating region (b) Selection of CH (c) Output voltage (d) Selection of LA

$$I_O = \frac{1}{Z_O T_S \omega_r}\left[\frac{V_S}{n} - \frac{1}{n}\left(\frac{V_S}{2} - \frac{\Delta V_H}{2}\right) - \left(\frac{V_O}{2} - \frac{T_S I_O}{4C_{o1}}\right)\right]\left[1 - \cos\left(\frac{T_S\omega_r}{2}\right)\right]. \quad (7)$$

The waveform of Ipri should be similar with that of below resonance mode (BRM), so as to reduce the conduction loss. Therefore, the resonant frequency, which is decided by the resonant inductance and capacitance, should be selected according to the switching frequency. Fig 3 (a) shows the desirable region of resonant frequency when the switching frequency is fixed as 80kHz. To be optimized about the conduction loss, the resonant frequency should be selected within the operating region. After the resonant frequency is selected, CH can be decided with the resonant inductance 76μH, as shown in Fig. 3 (b). ΔVH, which is controlled by the auxiliary circuit, can be obtained in similar way as mentioned above.

$$\Delta V_H = \frac{1}{C_H L_A}\left[\frac{V_S}{2} - \frac{T_S I_O}{2nC_H}\right](T_S D_E)^2. \quad (8)$$

From the equation (7) and (8), the steady state voltage conversion ratio of the overall system can be derived.

$$\frac{V_O}{V_S} = \frac{\frac{1}{2} + \frac{(T_S D_E)^2}{4C_H L_A}}{\frac{T_S}{R_O}\left[\frac{nZ_O\omega_r}{1-\cos(T_S\omega_r/2)} - \frac{n}{4C_{o1}} + \frac{(T_S D_E)^2}{2C_H L_A}\right] + \frac{n}{2}}. \quad (9)$$

Using the equation (9), the output voltage can be plotted as shown in Fig. 3 (c). In this figure, the output voltage can be obtained without the operation of auxiliary circuit at no load (DE=0). As the load goes to full load, DE is increased to get the desired output voltage. From this figure, the output voltage can be regulated well at whole load condition by the auxiliary circuit. Fig. 3 (d) shows the proper auxiliary inductance to achieve the load regulation according to the maximum duty ratio DE-MAX. As shown in this figure, when LA is selected as the larger value to reduce the peak value of ILA, the more maximum duty ratio is needed.

4. EXPERIMENTAL RESULTS

A 450W prototype of the proposed converter has been built for the experiment. Fig. 4 shows experimental waveforms. As can be seen in this figure the waveform of Ipri is similar with that of BRM. This results in less conduction loss and lower peak values of Ipri, IDS1, and IDS2. In addition, CH is additionally charged or discharged by ILA to regulate the output voltage. The ZVS operation of QM1 and QM2 at 10% and full load is also shown. In order to have the waveforms of current in BRM, the leakage inductance is rather large compared with that of half bridge LLC resonant converter. Due to the large leakage inductance, the ZVS operation of QM1 and QM2 is easily achieved even 10% load. Moreover, the ZCS operation of DS1 and DS2 can be achieved. The voltage across the DS1 and DS2 can be clamped to the output voltage without the high voltage ringing.

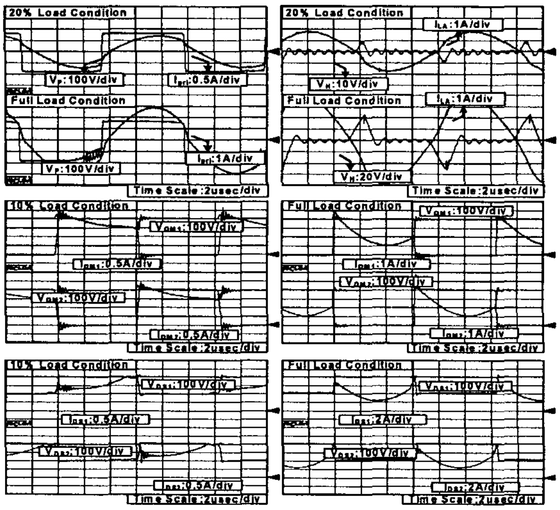

Fig. 4 Experimental Wavefomrs

5. CONCLUSION

A new PWM-controlled quasi-resonant converter for high efficiency PDP sustaining power module is proposed in this paper. Since the load regulation of the proposed converter can be achieved by the auxiliary circuit, the waveforms of current can be optimized on conduction loss especially at the light load conditions. Moreover, employing the voltage doubler type rectifier, additional resonant ripple of the voltage across the rectifier capacitors helps the operation of the auxiliary circuit. Besides, DC offsets of the magnetizing current and magnetic flux can be completely blocked. From the experimental results, good ZVS capability of the power switches QM1 and QM2 is also proved. Fig. 5 shows the measured efficiency. The measured efficiency with 10%~40% load range, is higher than that of the half bridge LLC resonant converter. As mentioned in introduction, when the PDP is operating on TV signals, the sustaining power module is usually operating on light load conditions. Thus, the proposed converter is expected to be suitable for the sustaining power module of PDP. Additionally, the measured efficiency along wide load ranges shows as high as around 94%. Therefore, the proposed converter demonstrates its suitability as a sustaining power module owing to its simple control circuits, low noise and high efficiency.

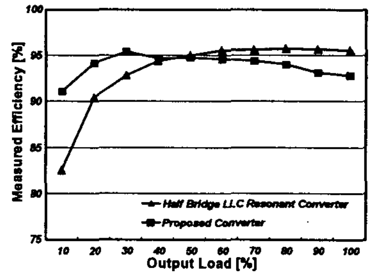

Fig. 5 Measured Efficiency References

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