Design of Class E Resonant Inverter Incorporating Piezoelectric Transformer

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Abstract

A novel single-stage backlight inverter constituted by class E topology for simplifying the circuit structure and promoting the system efficiency is proposed in this paper. Moreover, in order to overcome the drawbacks of the conventional electromagnetic transformer and miniaturize the backlight module, the piezoelectric transformer (PT) is adopted for driving the cold cathode fluorescent lamp. A simple model and design procedure are constructed by reasonable parameter simplification and combination schemes. Complete analysis and design considerations are discussed in detail. Experimental results agree with the theoretical prediction.

Keywords: backlight, resonant inverter, CCFL, LCD, piezoelectric Transformer

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I. INTRODUCTION

Due to the popularity of information apparatus, the requirements for thin and flat display panels are increasing drastically to replace conventional CRT displays used in many products with different panel sizes. The backlight module is a crucial component for driving light source in flat display panel (FDP) technologies, and its performance will influence the display quality of FDP. At present, there are two major backlight sources available, one is white LED, and the other is cold cathode fluorescent lamp (CCFL). Because of the high cost and small lighting area, the white LED is usually used in small-sized devices, such as PDA, GPS, and portable instruments. The CCFL is normally suitable for the larger display panels. The LCD combined with CCFL satisfies the increasing demand on display performance, size, and efficiency [1]. In the past, the conventional electromagnetic transformers were generally adopted in the backlight module design to boost output voltage for driving CCFL. However, the conventional transformer will cause the magnetic hysteresis loss and electromagnetic interference (EMI) and thus decrease the efficiency of energy conversion. In 1956, Rosen presented the direct and converse piezoelectric effect of piezoelectric components as the mechanism of mechanical and electrical energy interchange, which provides high conversion efficiency by transferring electrical energy via the polarized and mechanical vibration medium. Therefore, piezoelectric components can be employed in promoting and demoting power voltages as well. As a sinusoidal alternating voltage with frequency close to the resonant frequency is applied to the input terminal of a piezoelectric component, because the converse piezoelectric effect causes the piezoelectric component to yield the mechanical strain resonance and then the direct piezoelectric effect causes stress deformation of the resonance, the mechanical energy is transformed to electrical energy and outputted from the piezoelectric component, and hence the voltage transformation is achieved [3, 6]. Since PT can sustain its inherent high voltage gain while matching high impedance load with optimal efficiency, it is applicable to driving high-voltage high-impedance lamps, such as CCFL. In addition, PT possesses many advantages, namely, compact size, high efficiency, low power losses and free of EMI. Moreover, PT not only can provide high voltage gain but also is easy to realize the miniature design of FDP, in which the backlight module is joined with PT [5-6]. Section II depicts and analyzes the main circuit configuration employed in this paper. Section III introduces the model integration of class E resonant inverter with piezoelectric transformer. See-
tion IV discusses the design considerations and demonstrates the experimental results. Finally, Section V is conclusions.

II. ANALYSIS OF CLASS E RESONANT INVERTER

Fig. 1 shows the basic configuration of class E resonant inverter, which consists of choke inductor $L_1$, a power switch $S$, a shunt capacitor $C_1$, and a $L_x$-$C_x$-$R_L$ series–resonant circuit. The following assumptions are essential to simplify the analysis procedures: 1) The power switch and diode are both ideal. 2) The choke inductor is high enough so that its ac component is much lower than the dc component of the input current. 3) The quality factor $Q_L$ of the $L_x$-$C_x$-$R_L$ series–resonant circuit is high enough so that the current $i$ through the resonant circuit is sinusoidal.

File 1. Basic circuit scheme of class E converter

Fig. 2 depicts the equivalent circuits of class E resonant inverter. When the switch is on as Fig. 2a, the resonant circuit consists of $L_x$, $C_x$ and $R_L$ because $C_1$ is short-circuited by the switch. Hence, the resonant frequency $f_{o1}$ is obtained as follows:

$$f_{o1} = \frac{1}{2\pi \sqrt{L_xC_x}}$$

(1)

However, when the switch is off as Fig. 2b, the resonant circuit consists of $C_1$, $L_x$, $C_x$ and $R_L$ connected in series. The resonant frequency $f_{o2}$ can be expressed as below

$$f_{o2} = \frac{1}{2\pi \sqrt{L_xC_x/(C_x+C_1)}}$$

(2)

From the preceding statement, $f_{o1}$ is lower than $f_{o2}$, i.e., $f_{o1} < f_{o2}$. In general, the operating frequency $f$ is selected between $f_{o1}$ and $f_{o2}$ to acquire an approximate sinusoidal load current, i.e. $f_{o1} < f < f_{o2}$.

(3)
For $f > f_{01}$, the series-resonant circuit represents an inductive load shown as Fig. 3, where $L_a$ can be divided into $L_a$ and $L_b$. Assuming that the resonant frequency formed with $L_a$ and $C_a$ is equal to the operating frequency $f$, that is

$$\omega = \frac{1}{\sqrt{L_aC_a}}$$

(4)

The quality factor $Q_L$ is

$$Q_L = \frac{\omega L_a}{R_i} = \frac{\omega(L_a + L_b)}{R_i} = \frac{1}{\omega C_a R_i} + \frac{\omega L_a}{R_i}$$

(5)

Then the resonant current $i$, which through the resonant circuit, can be represented as sinusoidal form.

$$i = I_m \sin(\omega t + \phi)$$

(6)

where $I_m$ and $\phi$ are amplitude and phase angle of the resonant current $i$, respectively.
Therefore, the currents in switch S and shunt capacitor $C_1$ can be expressed as

$$i_1 + i_{c1} = I - i = I - I_n \sin(\omega t + \phi)$$  \hspace{1cm} (7)

For the time interval $0 < \omega t \leq 2\pi D$, the switch is on and therefore $i_{c1} = 0$. Oppositely, for the time interval $2\pi D < \omega t \leq 2\pi$, the switch is off, which implies that $i_1 = 0$. Hence, the current through the shunt capacitor $C_1$ is given by

$$i_{c1} = \begin{cases} 0 & \text{for } 0 < \omega t \leq 2\pi D \\ I - I_n \sin(\omega t + \phi) & \text{for } 2\pi D < \omega t \leq 2\pi \end{cases}$$  \hspace{1cm} (8)

The voltage across the shunt capacitor $C_1$ and the switch is found as

$$v_s = \frac{1}{\omega C_1} \int_0^{2\pi} i_{c1} d(\omega t)$$

$$v_s = \left\{ \begin{array}{ll}
0 & \text{for } 0 < \omega t \leq 2\pi D \\
\frac{1}{\omega C_1} \{I(\omega t - 2\pi D) + I_n[\cos(\omega t + \phi) - \cos(2\pi D + \phi)]\} & \text{for } 2\pi D < \omega t \leq 2\pi
\end{array} \right.$$  \hspace{1cm} (9)

In optimum operation [12], substituting the boundary condition $v_s = 0$ at $\omega t = 2\pi$ into (9) yields the amplitude of resonant current

$$I_n = I \frac{2\pi (1 - D)}{\cos(\omega t + \phi) - \cos \phi}$$  \hspace{1cm} (10)

Substitution of (10) into (8) and (9), we can obtain $i_{c1}$ and $v_s$ as follows:

$$i_{c1} = \left\{ \begin{array}{ll}
0 & \text{for } 0 < \omega t \leq 2\pi D \\
\frac{1}{\omega C_1} \{I(\omega t - 2\pi D) + \frac{2\pi (1 - D)\{\cos(\omega t + \phi) - \cos(2\pi D + \phi)\}}{\cos(2\pi D + \phi) - \cos \phi}\} & \text{for } 2\pi D < \omega t \leq 2\pi
\end{array} \right.$$  \hspace{1cm} (11)

$$v_s = \left\{ \begin{array}{ll}
0 & \text{for } 0 < \omega t \leq 2\pi D \\
\frac{I}{\omega C_1} \{\cos(\omega t + \phi) - \cos(2\pi D + \phi)\} & \text{for } 2\pi D < \omega t \leq 2\pi
\end{array} \right.$$  \hspace{1cm} (12)

Rearranging (12), the input voltage $V_D$ is

$$V_D = \frac{1}{2\pi} \int_{\omega t}^{\pi} v_s d(\omega t) = \frac{I}{\omega C_1} \left\{ \frac{(1 - D)\{\pi(1 - D)\cos \pi D + \sin \pi D\}}{\tan(\pi D + \phi) \sin \pi D} \right\}$$  \hspace{1cm} (13)
Substituting the boundary condition for optimum operation [12], \(\text{div}d(\alpha t)=0\) at \(\alpha t=2\pi\) into (13) yields the relationship between phase \(\phi\) and duty cycle \(D\).

\[
\phi = \pi + \arctan \left( \frac{\cos 2\pi D - 1}{2\pi(1-D) + \sin 2\pi D} \right) \tag{14}
\]

Since the resonant current \(i\) can be regarded as sinusoidal form, the fundamental component of \(v_r\) at operating frequency is

\[
v_r = V_{nl} \sin(\omega t + \phi) + V_{ls} \cos(\omega t + \phi) \tag{15}
\]

Combining \(V_{nl}\) in (15) with (12) by Fourier formula yields

\[
V_{nl} = \frac{1}{\pi} \int_{0}^{\pi} v_r \sin(\omega t + \phi) d(\omega t) = \frac{2 \sin \pi D \sin(\pi D + \phi)}{\pi(1-D)} V_b \tag{16}
\]

And combing \(V_{ls}\) in (15) with (13) by Fourier formula yields

\[
V_{ls} = \omega L_s I_s = \frac{1}{\pi} \int_{0}^{\pi} v_r \cos(\omega t + \phi) d(\omega t)
\]

\[
= \frac{1 - 2(1-D)^2 \pi^2 - 2 \cos \phi \cos(2\pi D + \phi)[\cos 2\pi D - \pi(1-D)\sin 2\pi D]}{2(1-D)\pi \cos(\pi D + \phi)[(1-D)\pi \cos \pi D + \sin \pi D]} V_b \tag{17}
\]

Combining (10), (13) and (16) yields

\[
C_i = \frac{2 \sin \pi D \cos(\pi D + \phi) \sin(\pi D + \phi)[(1-D)\pi \cos \pi D + \sin \pi D]}{\pi^2(1-D)\omega R_i} \tag{18}
\]

By combing (10), (13), (17) and (18) yields

\[
\frac{\omega L_s}{R_i} = \frac{2(1-D)^2 \pi^2 - 1 + 2 \cos \phi \cos(2\pi D + \phi) - \cos 2(\pi D + \phi) [\cos 2\pi D - \pi(1-D)\sin 2\pi D]}{4 \sin \pi D \cos(\pi D + \phi) \sin(\pi D + \phi)[(1-D)\pi \cos \pi D + \sin \pi D]} \tag{19}
\]

Assuming that the main circuit operates in optimum operation, that is, duty cycle \(D=0.5\) [12], one can obtain the phase \(\phi=147.52^\circ\) from (14). Substituting above-mentioned parameters into (18) and (19), then (18) can be simplified as

\[
C_i = \frac{8}{\pi(\pi^2 + 4)\omega R_i} \tag{20}
\]
And (19) is rearranged as
\[
\frac{\omega L_a}{R_c} = \frac{\pi(\pi^2 - 4)}{16}
\]  
(21)

From (5) and (21), we can derive \(C_s\) as follows:
\[
C_s = \frac{1}{\omega R_c \left[ Q_c - \frac{\pi(\pi^2 - 4)}{16} \right]}
\]  
(22)

The resonant inductor \(L_x\) is calculated from (5).
\[
L_x = \frac{Q_c R_c}{\omega}
\]  
(23)

III. MODEL INTEGRATION OF CLASS E INVERTER WITH PIEZOELECTRIC TRANSFORMER

A Rosen type PT, as shown in Fig. 4, is composed of two poled piezo-ceramic plates with equal cross-sections, or a single piezo-ceramic plate with both ends poled separately. Moreover, PT is based on the use of both piezoelectric effects, in which an applied electric energy is transformed into mechanical vibrations and then these vibrations are transformed back into electric energy [13]. In other words, the input terminals are supplied with an alternating electric field that produces a time varying strain in the ceramic material. Then this strain is electromechanically coupled along the length of the plate, and induced a polarization field between the two output terminals [14]. In this paper, PT is employed to integrate class E resonant inverter for simplifying the circuit structure and promoting the system efficiency. Fig. 5 shows the employed class E resonant inverter, which incorporates a piezoelectric transformer, for driving CCFL. When the operating frequency of the circuit is close to the resonant frequency of the employed inverter, the harmonics and DC component of the current flowing in the circuit will be filtered out due to the high quality factor QL of the main circuit [3-4]. Therefore, sinusoidal voltage and current can be produced to start and drive CCFL. Fig. 5a illustrates the equivalent main circuit of Fig. 1 in steady state, in which a PT equivalent model, external resonant components, and a secondary reflected impedance are included to form a
resonant tank \[1, 8\]. The secondary reflected impedance comprises lamp impedance and parasitic capacitance. The mathematical models of derived circuits presented in \[4-5\] are too complicated due to numerous internal equivalent parameters of PT. Therefore, reasonable parameter simplification and combination schemes are utilized to decide the desired parameters of circuit components and to derive a simple mathematical model. To simplify the circuit model, the parasitic capacitance is neglected and the CCFL is replaced by RCCFL. The secondary components of PT are converted equivalently to the primary ones of PT and the resultant circuit is shown in Fig. 5b. When PT is operating in its resonant frequency, R, L, and C elements in the circuit model are considered as short-circuited due to the voltages of inductor L and capacitor C having the same magnitudes but opposite polarities. Thus, we can further simplify the circuit to be equivalent to a resonant circuit structure as shown in Fig. 5c. Fig. 5d illustrates the reduction circuit model, where

\[
C_R = C_{01} + N^2 C_{02} \tag{24}
\]

Fig. 5d also can be converted into Figs. 5e or 5f by circuit transformation technique.

From the (24), \(R_{\text{RCCFL}}\) and \(N\) in Fig. 5c, the equivalent load \(R_L\) is found as

\[
R_L = \frac{R_{\text{RCCFL}}}{1 + \left(\frac{R_{\text{RCCFL}}}{N^2 X_{\text{cc}}}\right)^2} \tag{25}
\]
IV. DESIGN CONSIDERATION AND EXPERIMENT RESULTS

Fig. 5. (a) Equivalent circuit model of class E resonant inverter with PT, (b) equivalent circuit of 5a, (c) simplified circuit of 5b, (d) simplified circuit of 5c, (e) simplified circuit of 5d, (f) simplified circuit of 5e.

Substituting $R_L$ into (20), (22) and (23), we can acquire $C_1$, $C_2$ and $L_{\infty}$, respectively. From Figs. 5d and 5e, the impedance of $C_{K2}$ can be expressed as

\[ X_{ct2} = \frac{X_{ct}}{1 + \left(\frac{X_{ct}}{R_{CT}}\right)^2} \]  

(26)
Owing to the $C_{K2}$ can be expressed as

$$C_{K2} = \frac{1}{\omega X_{K2}}$$

(27)

Hence, the equivalent capacitor $C_s$ can be calculated from Figs. 5e and 5f.

$$C_s = \frac{C_s C_{K1}}{C_s + C_{K2}}$$

(28)

Therefore, the resonant capacitor is obtained from (28) as follows:

$$C_s = \frac{C_s C_{K1}}{C_{K2} - C_s}$$

(29)

The quality factor $Q_L$ is an important parameter to determine the resonant components in this paper. The reasonable value of $Q_L$ will be decided in the following discussion. First, due to $C_s > 0$ in (29), the equivalent capacitor $C_{K2}$ should higher than $C_s$. Substituting $C_{K2}$ and $RL$ into (22) yield $Q_{LA}$, which is the lower boundary value.

$$Q_{LA} = \frac{1}{\omega R_s C_{K2}} + 1.1525$$

(30)

Moreover, quality factor $Q_L$ will also directly affect the resonant components $L_x$ and $C_x$. Hence, we can define a parameter $\gamma$ by combing (22) and (23) as follows:

$$\gamma = \frac{1}{L_x C_x \omega} = \frac{Q_L - 1.1525}{Q_L}$$

(31)

The relationship of $\gamma$ and $Q_L$ can be depicted by Matlab simulation.

From Fig. 6, it is noted that $\gamma$ seems unaffected by $Q_L$ when $\gamma > 0.9$, in other words, $\gamma$ is almost at saturation state; However, it is apparent that $\gamma$ is susceptible to $Q_L$ when $\gamma < 0.9$. Then we can set the corresponding value of quality factor ($Q_{LB}$) at $\gamma = 0.9$ as the higher boundary value. Therefore, we can acquire a reasonable range of quality factor, that is, $Q_{LA} < Q_L < Q_{LB}$. 
IV. DESIGN CONSIDERATION AND EXPERIMENT RESULTS

CCFL, which is characterized by its length, diameter, and structure, is inherently a nonlinear load and hence will influence the design of driving circuit. In general, the best lamp current for driving CCFL is sinusoidal, i.e., its crest factor near 1.414, which does not only reduce EMI but also raises efficiency [1]. Although other wave shapes may provide higher luminance, nevertheless the lamp life will be shortened [10]. Additionally, the conventional electromagnetic transformer has not only bulky volume, which is not adequate for designing thin and flat display panel, but also causes magnetic hysteresis and electromagnetic interference, which decrease the efficiency of energy conversion [8]. Therefore, a PT is employed to overcome the preceding disadvantages. Moreover, PT possesses high efficiency, high gain of conversion performance and low power dissipation [5, 6, 8, 11]. The PT utilized in the paper is EFTU14R0M02, which has rated power of 4W, resonant frequency of 57 kHz, input voltage of 22 Vrms(max), input current of 500 mA rms(max), output voltage of 820 Vrms(max), output current of 7 mA rms(max), and operating temperature range of -10°C~60°C. The CCFL used is FL-30266AE, which has rated power of 3.4 W, operating voltage ν₀ of 620 Vrms, operating current i₀ of 5±1 mA, and starting voltage νstart of 930 Vrms. The operating frequency of CCFL is generally within the range of 20~80 kHz. In this paper, we select the operating frequency f₀ of 57 kHz and input voltage V₀ = 12VDC. The most important parameters in the proposed circuit structure are the internal equivalent impedance R₀ of PT as well as resonant components C₁, L₀, Cₓ of class E resonant inverter. The design considerations and steps of these parameters are discussed below.
1) Measuring the parameters of the equivalent PT model

Fig. 7 Equivalent internal model of PT.

Fig. 7 shows the equivalent circuit model of PT operating in its resonant frequency. The parameters of the employed PT are measured with an impedance analyzer HP-4194A. The related parameter values measured are shown below

\[
\begin{align*}
C_{01} & = 99.66 \text{ nF} \\
L & = 1.22 \text{ mH} \\
R & = 1.26 \Omega \\
C_{02} & = 9.09 \text{ pF} \\
C & = 6.21 \text{ nF} \\
N & = 55
\end{align*}
\]

2) Calculating proper values of \( R_1 \) and \( C_{K2} \)

In steady state, the lamp voltage and lamp current employed in this paper are \( 620 \text{ v}_{\text{rms}} \) and \( 5 \text{ mA}_{\text{rms}} \) respectively. Hence, the equivalent lamp impedance is \( R_{\text{CCFL}} = 620v_{\text{rms}}/5mA_{\text{rms}} =124 \text{ k}\Omega \). In addition, \( f = 57\text{kHz} \) is a better operating frequency of the employed PT under room temperature. According to (24)-(27) and above-mentioned parameters, we can obtain \( R_1 = 9.17 \Omega \), and \( C_{K2} = 163.5 \text{ nF} \).

3) Estimating the quality factor \( Q_1 \) and \( C_s \)

From (30) and \( C_{K2} \) estimated in previous step, we can obtain the lower boundary value \( Q_{LA} = 3.0148 \). According to the previous statement, the higher boundary value \( Q_{LB} = 11.525 \) is found by substituting \( \gamma = 0.9 \) into (31). Therefore, the reasonable value of \( Q_1 \) is in the range of \( 3.0148 < Q_1 < 11.525 \), we choose \( Q_1 = 8 \) in the practical design. Substituting \( R_1 \) and \( Q_1 \) into (22) yields \( C_s = 45.8 \text{ nF} \).

4) Determining parameters of resonant circuit

Substituting \( Q_{LA}, C_{ss}, R_L \) and parameters of PT into (20), (23) and (29), we can calculate the shunt capacitor \( C_1 = 55.9 \text{ nF} \) (use 57 nF), the resonant components \( L_N = 204.8 \mu\text{H} \) (use 201 \( \mu\text{H} \)) and \( C_N = 61 \text{ nF} \) (use 68 nF), respectively. All the estimated parameters are substituted into (1) and (2) to yields \( f_0 = 52.72 \text{ kHz} \) and \( f_0 = 71.11 \text{ kHz} \). Hence, the operating frequency \( f = 57 \text{ kHz} \) employed in this paper is satisfied the requirement of (3).
Fig. 8 demonstrates the measured waveforms of gating signal $v_{gs}$ and the switch voltage $v_{dr}$ under ideal operating mode ($D=0.5$) with zero voltage switching (ZVS) [12]. Fig. 9 shows the measured waveforms of switch current $i_s$ and the shunt capacitor current $i_{cl}$, which forms an approximate sinusoidal current to drive the load. The choke inductor current $i_L$ and resonant current $i$ are illustrated in Fig. 10, the measured phase difference is $\phi=139^\circ$, which is close to the derived phase difference $\phi=147.52^\circ$, with error of 8° (below 6%). From the experimental results, the measured waveforms of the input current of PT and lamp current $i_{CCFL}$ are both sinusoidal type as shown in Fig. 11.

Fig. 8. Measured waveforms of $v_{gs}$, $v_{dr}$ (Ver: 5V/div for $v_{gs}$; Ver: 10V/div for $v_{dr}$; Hor: 4\mu s/div)

Fig. 9. Measured waveforms of $v_{gs}$, $i_s$ and $i_{cl}$ (Ver: 5V/div for $v_{gs}$; 1A/div for $i_s$; Ver: 500mA/div for $i_{cl}$; Hor: 4\mu s/div)
V. CONCLUSIONS

In this paper, a class E resonant inverter is incorporated with a piezoelectric transformer to drive the cold-cathode fluorescent lamp with zero voltage switching. The model integration and simplification are both adopted to derive the optimal parameters of backlight system, so as to promote the system efficiency and stability. In addition, the overall volume is reduced and thin-shaped to miniature the backlight module. The overall system efficiencies are all above 91% in all test conditions.

Fig. 10. Measured waveforms of inductor current $I$ and resonant current $I$ (Ver: 100mA/div for $I$ ; Ver: 500mA/div for $I$ ; Hor: 4μs/div)

Fig. 11. Measured waveforms of input current of piezoelectric transformer $i_{PT}$ and lamp current $i_{CCFL}$ (Ver: 1A/div for $i_{PT}$ ; Ver: 5mA/div for $i_{CCFL}$ ; Hor: 4μs/div)
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E類諧振換流器結合壓電變壓器之設計

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摘要

本文中提出一種由E類架構所組成之新型的單級背光換流器，可簡化電路架構及提升系統效率。其次，為了克服傳統電磁式變壓器之諸多缺點，並達到背光模組小型化的目的，採用壓電變壓器來驅動冷陰極管。再者，本文中利用合理的參數簡化及合併的技巧來建構簡單的系統模型，而完整的分析及設計基礎也將詳細地討論。最後，以實際量測結果來驗證理論的正確性。

關鍵詞：背光、諧振換流器、冷陰極管、液晶顯示器、壓電變壓器