Design Considerations for Achieving ZVS in a Half Bridge Inverter that Drives a Piezoelectric Transformer with No Series Inductor

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Abstract **A general procedure for maintaining soft switching in inductor-less half-bridge piezoelectric transformer (PT) inverter was analyzed by applying the equivalent circuit of the PT device. Soft switching capability of the PT was delineated and detailed guidelines are given for the load and frequency boundaries, voltage transfer function and the output voltage that will keep the operation under ZVS conditions. The analysis takes into account the maximum power dissipation of the PT, which is used to bind the permissible power transfer through the device. The analytical results for the half-bridge inductor-less PT inverter where verified by simulation and experiments for a radial vibration mode PT.**

I. Introduction

As Piezoelectric Transformer (PT) technology is developing, PTs may become a viable alternative to magnetic transformers in various applications [1, 2]. Power supplies that employ PTs, rather than the classical magnetic transformers, could be made smaller in size - an attribute that is important in a number of applications such as battery chargers, laptop computers supplies, fluorescent lamp drivers etc. However most of earlier designs of PT based converters/inverters used additional series inductors to achieve Zero Voltage Switching (ZVS) condition [3 - 7]. By this, the PT advantages of small size were inadvertently lost. It was already shown in [8] that by using specific characteristics of the PT, ZVS could be achieved without any additional elements. This can be accomplished when the circuit is operating at a frequency that is higher than the resonance frequency of the PT and sufficient energy is available to charge and discharge the input capacitance of PT during the switching dead time. Thus, by utilizing the characteristics of the PT, the switches of the inverter will operate under ZVS conditions reducing significantly the turn-on switching - without the need to include a series inductor. In addition, the inherent input capacitance of the PT works as a turn-off snubber for the power switches. This further decreases the turn-off voltage spikes and thus the turn-off losses of the switches.

This paper presents a comprehensive analysis of the of inherent soft switching capability of PTs. Closed form equations estimate the load and the frequency boundaries that allows soft switching in power inverter/converter built around a given PT.

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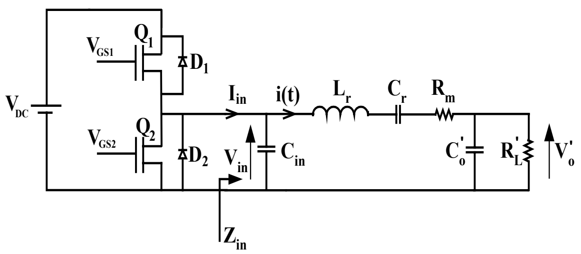

Fig. 1. Equivalent circuit of a half bridge inverter driving a PT around the resonant frequency.

II. Analysis and Design of ZVS PT Power Inverter

A. *Analysis of ZVS Condition for Half-Bridge Inverter*

The analysis is carried out on a half bridge inverter shown Fig.1. It includes two bi-directional switches Q_1 and Q_2 including anti-paralleled diodes D1, D2, and a PT, that is represented by an equivalent series-parallel resonant circuit $R_m - L_r - C_r - C_O{}' - R_L{}' - n$. the parameters L_r, C_r, R_m represent the mechanical behavior of the PT, C_{in} is the input capacitance of PT plus the output capacitance of the switches, $C_O{}' = \frac{C_O}{n^2}$ is the reflected output capacitance, where C_O is the output dielectric capacitance and *n* is the gain, $R_L{}' = n^2 R_L$ is the reflected load resistance R_L [3].

The switches Q_1 and Q_2 (Fig. 1) will normally be power MOSFETs and will include an inherent anti-parallel diodes D_1 and D_2. The switches are driven alternately by rectangular voltages V_{GS1} and V_{GS2} with a sufficiently long dead time. Fig. 2 depicts steady-state current and voltage waveforms in the inverter for an operating frequency that is higher than the resonance frequency.

The charging process of the capacitor is considered with reference to Fig. 2. At the instant t_O the drive voltage V_{GS2} turns OFF Q_2. Transistor Q_1 is still kept in the OFF state by the drive voltage V_{GS1}. Both diodes are reversed biased and remain in the OFF condition.

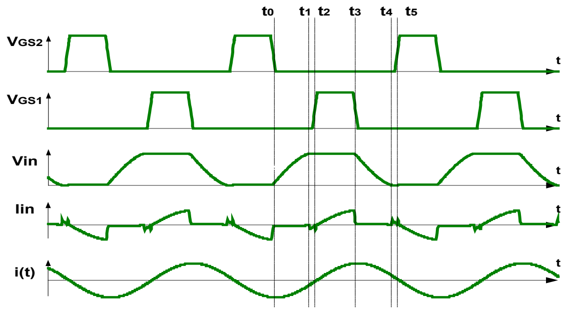

Fig. 2. The steady-state current and voltage waveforms of the ZVS PT inverter.

Assuming that the Q of the network is high, the sinusoidal current waveform of the resonant circuit can be represented by:

$$i(t) = I_m \sin(\omega t - \psi) \tag{1}$$

where I_m and ψ are the current peak and the initial phase respectively (referred to the phase of the first harmonic of the input voltage). When transistor Q_2 is turned OFF at the instant t_0 this current is diverted from transistor Q_2 to the capacitor C_{in}. Thus, the current through the shunt capacitor during $t_0 < t < t_1$ is:

$$i_{C1}(t) = -i(t) = -I_m \sin(\omega t - \psi) \tag{2}$$

This current charges the capacitor C_{in} and the voltage across the capacitor (and hence across the switch Q_2) gradually increases from zero to V_{DC}. If the current i_{S2} through transistor Q_2 is forced to drop quickly to zero (by a proper gate driver) switching losses in the transistor Q_2 will be low. At t_1, the voltage across the capacitor C_{in} reaches V_{DC}, therefore the diode D_1 turns ON and the current $i(t)$ is diverted from the shunt capacitor C_{in} to the diode D_1. The voltage across the top switch becomes zero. The diode D_1 conducts during the interval $t_1 t_2$. When the voltage V_{in} decreases to V_{DC} at the instant t_2, the drive voltage V_{GS1} turns the upper transistor ON. The transistor Q_1 is ON during the time interval $t_2 t_3$. At t_3, it turns OFF again. Since transistor Q_2 still remains OFF, the current of the resonant circuit discharges the shunt capacitor C_{in}, decreasing V_{DS2} and thereby increasing V_{DS1}. The discharging process of C_{in} is taking place during the time interval from t_3 to t_4. When the voltage across C_{in} reaches zero, the diode D_2 starts to conduct at t_4 and the voltage across the top switch S_2 becomes zero. Since the charging-discharging process takes place when the switch current is zero (both switches are OFF) the switching losses could be made small. In fact, the capacitor C_{in} works as a turn-off snubber for the switches Q_1 and Q_2 of the half-bridge inverter.

B. The Main Assumptions

The analysis is carried out under the following approximations [9]:

1) The capacitor charging time is shorter than the switching dead time.
2) The capacitor is charged by a constant current.
3) The input voltage $V_{in}(t)$ is assumed to be a symmetrical rectangular waveform (instead of trapezoidal). (It can be seen, that this assumption is relevant because the difference between the first harmonic amplitude of the rectangular and trapezoidal waveforms is small).
4) The power losses on PT are limited to 5-10% of the output power.

In order to ensure ZVS for the switches, the input capacitor has to be charged-discharged within the switching dead time which duration is less than T/4, where T is the period of the resonant current that developed during the dead time (see below).

The charging process begins at t_0. If the charging time t_r is much shorter than the cycle T=1/f, the charging current can be assumed to be constant and given approximately by:

$$I = i_{Cin}(0) = I_m \sin\psi \tag{3}$$

The charging process ends when the capacitor voltage reaches V_{DC}. Hence, the charging time is approximately [9]:

$$\begin{aligned} t_r &= \frac{C_{in} V_{DC}}{I_m \sin\psi} = \frac{\pi}{2}\frac{C_{in} V_{in(1)p}}{I_m \sin\psi} = \\ &= \frac{\pi}{2} C_{in} \frac{|Z_{in}|}{\sin\psi} < \frac{1}{4} T \end{aligned} \tag{4}$$

where $V_{in(1)p}$ is the fundamental component of the rectangular waveform, and Z_{in} is the input impedance of the resonant tank (not including C_{in}).

In order to transfer sufficient energy to the output the inverter has to operate close to the frequency of maximum output power and under high efficiency conditions [8]. The power dissipated by the PT has to be limited to 5-10% of the output power, to achieve efficiencies in the order of 90-95%. For example, if the power dissipation of PT is limited to be 1W, the output power of 10W will obtained with 90% efficiency (assuming here a lossless inverter).

C. Normalized Model for Soft Switching PT Inverter

In order to generalize the analysis, we developed a normalized model for PT inverter that is applicable to any PT that can be described by the resonant network of Fig. 1. All parameters of the inverter considered in this study are normalized as follows:

The main initial parameters are defined as:

$$\begin{aligned} & a = \frac{C_o'}{C_r};\ b = \frac{C_{in}}{C'};\ \omega_r = \frac{1}{\sqrt{L_r C_r}} \\ & Q = \omega_r C_o' R_L';\ Q_m = \frac{1}{\omega_r C_r R_m} \end{aligned} \tag{5}$$

The normalized input impedance $()\Delta_{Zin}$ is defined as the ratio of the input impedance of the PT - Z_{in} to the reflected load resistance R_L':

$$\Delta_{Zin}=\frac{Z_{in}}{R_L'}=\frac{R_m+j\omega L_r+\frac{1}{j\omega C_r}+\frac{R_L'}{1+j\omega C_O'R_L'}}{R_L'}=\tag{6}$$
$$=\frac{A+jB}{1+j\frac{\omega}{\omega_r}Q}$$

where: $\begin{cases} A=1+\frac{a}{QQ_m}-a\left[\left(\frac{\omega}{\omega_r}\right)^2-1\right] \\ B=\frac{\omega_r}{\omega}\frac{a}{Q}\left[\left(\frac{\omega}{\omega_r}\right)^2-1\right]+\frac{\omega}{\omega_r}\frac{a}{Q_m}\end{cases}$

The normalized charging time Δ_r is defined as the ratio of the charging time t_r to the switching period T. (Note that Δ_r has to be less than ¼ - [8]-[9]):

$$\Delta_r=\frac{t_r}{T}=\frac{\pi}{2T}C_{in}\frac{|Z_{in}|}{\sin\psi}=\frac{1}{4}\frac{\omega}{\omega_r}bQ\frac{|\Delta_{Zin}|}{\sin\psi}\tag{7}$$

where ψ is the normalized input impedance phase angle (that is opposite to the initial current phase (1):

$$\psi=\arg\left(\frac{Z_{in}}{R_L'}\right)\tag{8}$$

The voltage transfer ratio ko is the ratio of the output voltage V_{out} to the peak of the first harmonic of the input voltage $V_{in(1)p}$ (Fig. 1):

$$ko=\frac{V_{out}}{V_{in(1)p}}=\frac{1}{|\Delta_{Zin}|\sqrt{\left[1+\left(\frac{\omega}{\omega_r Q}\right)^2\right]}}\tag{9}$$

The normalized power dissipated by the PT, Δ_{PD} is defined as the ratio of the PT power dissipated by the P_{PD} to the output power P_{out}:

$$\Delta_{PD}=\frac{P_{PD}}{P_{out}}=\frac{a}{QQ_m}\left[1+\left(\frac{\omega}{\omega_r}Q\right)^2\right]\tag{10}$$

The inverter efficiency is the ratio of the output power P_{out} to the input power P_{in}:

$$\eta=\frac{P_{out}}{P_{in}}=ko^2\frac{\Delta_{Zin}}{\cos(\Delta_{Zin})}\tag{11}$$

The maximum output voltage is reached at the equivalent resonant frequency ω_m. Since the equivalent resonant frequency is close to the series resonant frequency ω_r one can replace the normalized resonant frequency $\frac{\omega_m}{\omega_r}$ by the factor $1+\varepsilon$, where ε is a small number that represents the deviation from the normalized series resonant frequency. By taking the derivative of (9) and equating it to zero we obtain an approximate expression for ε:

$$\varepsilon\approx\frac{1}{2a\left(1+\frac{1}{Q^2}\right)}\tag{12}$$

The normalized operating frequency $\frac{\omega}{\omega_r}$ (the ratio of the operating frequency ω to the series resonant frequency ω_r) can now be expressed as:

$$\frac{\omega}{\omega_r}=\frac{\omega}{\omega_m}\frac{\omega_m}{\omega_r}=k(1+\varepsilon)\tag{14}$$

where the normalized frequency factor $k=\omega/\omega_m$ is the ratio of the operating frequency to the frequency of the maximum output power.

D. Design guidelines

Given: the PT inverter output voltage V_{out} and the PT parameters - $L_r, C_r, C_{in}, C_o, R_m, n$.

To be evaluated: the frequency range, the output power and the load boundaries for soft switching.

The general design steps:

1) On the basis of the specifications of the given PT we calculate the parameters a, b, Q_m (5).
2) For different Q we plot $\Delta_r(k)$ (6), (7), (13), (14).
3) For the same Q we calculate $\Delta_{PD}(k)$ (10).
4) Soft Switching is achieved in the k range where $\Delta_r(k)<0.25$. The upper boundary for Q is stated by the requirement $\Delta_r(k)=0.25$ and the lower boundary for Q is bounded by the PT power dissipation limit Δ_{PD}.
5) From the parameter Q and parameters of PT we calculate the load resistance R_L':

$$R_L'=\frac{Q}{\omega_r C_O'}\tag{15}$$

6) Based on the soft switching boundaries of the normalized frequency factor k, the series resonant frequency ω_r and the normalized load factor Q, we calculate the frequency boundaries for ZVS:

$$f=\frac{1}{2\pi}k\left(1+\frac{1}{2a\left(1+\frac{1}{Q^2}\right)}\right)\omega_r\tag{16}$$

For $k=const$ one can calculate the transfer function $ko=f(R_L)$ or for $R_L=const$ - the transfer function $ko=f(k)$.

Example 1.

Given: The PT is a radial vibration mode piezoelectric transformer (T1-2, Transoner[R]) [10], the power dissipation of PT is limited to Δ_{PD}=10% and the required peak output voltage is $V_{out(p)}$=30V.

To be evaluated: the frequency range, the input voltage range and the load range that ensures soft switching.

This PT has one layer at the input side and one layer at the output side. The diameter of the PT is 19mm; the thickness of the input layer is 1.52mm, the thickness of the output layer is 2.29mm.

Applying the HP4395A Impedance Analyzer, the parameters of the simplified electrical equivalent circuit for narrow frequency range around its mechanical resonant frequency were estimated to be:

$$R_M = 11.6\Omega, C_{in} = 2.19nF, C_O = 1.547nF, C_r = 120pF,$$
$$L_r = 15.1mH, f_{res} = 118.3kHz, n \approx 1$$

For these circuit parameters the normalized model parameters are calculated to be: a=12.9, b=1.416, Q_m=966.5.

As a preparation for the design we generate the following plots:

a) Fig. 3, based on equation (7), shows the plots of the normalized charge time Δ_r as a function of the normalized frequency factor k for different normalized load factor Q. It can be seen that the upper limit for Q to comply with $\Delta_r < 0.25$ is $Q \approx 0.37$. The lower boundary for the load factor is determined by $\Delta_{PD} < 0.1$ ($Q \approx 0.13$) (10).

b) Fig. 4, based on equation (9), shows the transfer function $ko = \frac{V_{out}}{V_{in}}$ as a function of the normalized frequency factor k for the same normalized load factors Q.

These plots are then used to calculate the input voltage range and power range for which ZVS can be achieved taking into account the design constraints and the maximum power dissipation on PT.

For any given load (Q) the corresponding plots from Fig. 3 and Fig. 4 can be combined into a single plot. For example Fig. 5 is for $Q = 0.15$ (The same plots can be built for different R_L values, to cover the desired power range). In this case soft switching frequency boundaries are $1.003 < k < 1.025$ (or, from (17), $118770 < f < 121360Hz$). For this frequency region the transfer function ko that can be achieved is $0.8 > ko > 0.22$ approximately. The range of the peak input voltage $V_{in(p)} = \frac{V_{out(p)}}{ko}$ will thus be $37.5V < V_{in} < 136V$ respectively (Fig. 6).

Example 2.

Given: the same PT as in the Example 1, the power dissipation on PT is limited to 10% of the output power, the peak input voltage $V_{in(p)}$=50V and load resistance $R_L = 130\Omega$.

To be evaluated: the boundaries of the output voltage and output power for soft switching operation.

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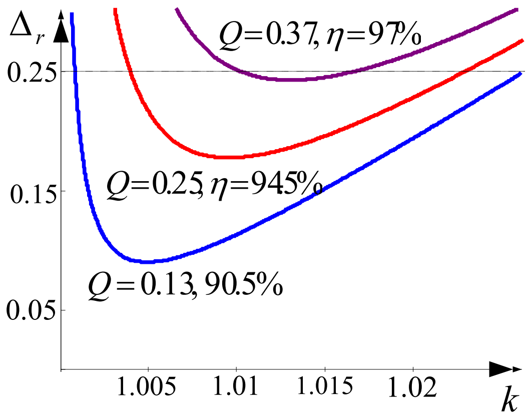

Fig. 3. Normalized charging time as a function of the normalized frequency factor k for different normalized load factors Q in the range Q=0.13 to Q=0.37. Data is for T1-2, Transoner[R] [10].

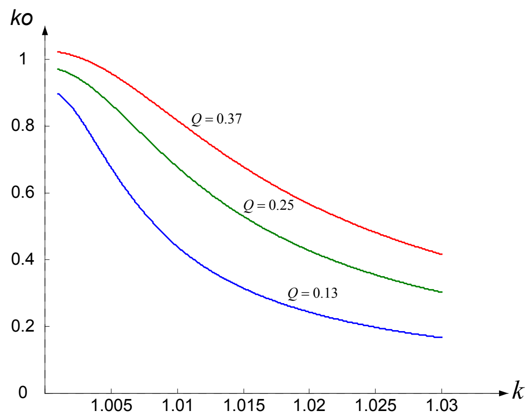

Fig. 4. Voltage transfer function ko as a function of the normalized frequency factor k for different normalized load factors Q in the range of Q=0.13 to Q=0.37. The PT under the calculations is (T1-2, Transoner[R]) [10].

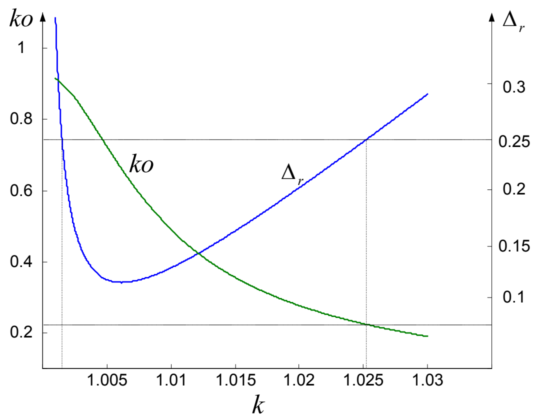

Fig. 5. Voltage transfer function ko and the normalized charging time Δ_r as a function of the normalized frequency factor k for normalized load factor Q=0.15 ($R_L = 130\Omega$). The PT under the calculations is (T1-2, Transoner[R]) [10].

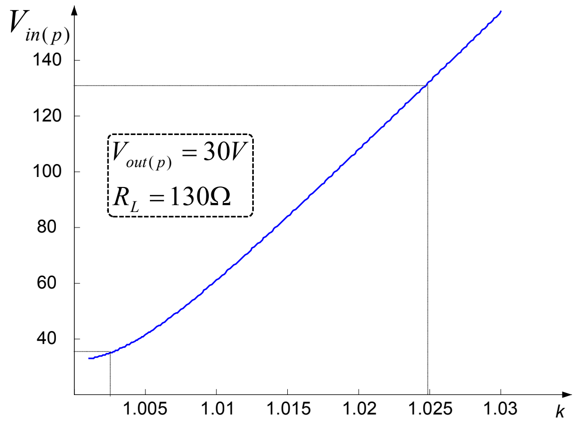

Fig. 6. Peak input voltage $V_{in(p)}$ as a function of the normalized frequency factor k for the constant output peak voltage $V_{out(p)} = 30V$ and the load resistance $R_L = 130\Omega$. Data is for T1-2, Transoner[R]) [10].

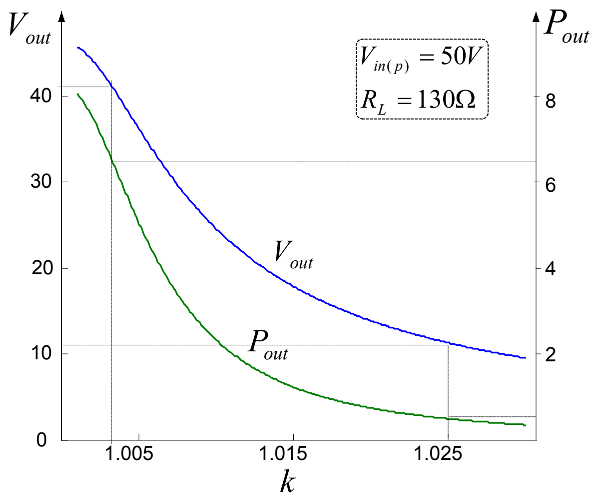

Fig. 7. The peak output voltage $V_{out(p)}$ and the output power P_{out} as a function of the normalized frequency factor k for peak input voltage $V_{in(p)} = 50V$ and load resistance $R_L = 130\Omega$. Data is for T1-2, Transoner[R]) [10].

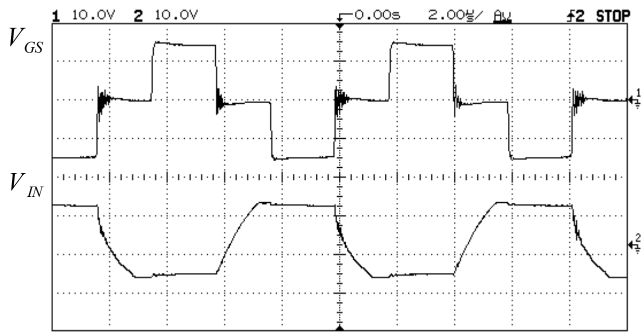

Fig. 8. Experimental voltage waveforms of the half-bridge inductor-less PT inverter: V_{GS} (upper plot), V_{IN} (lower plot). Operating frequency f=120kHz, load resistance R_L=130 Ohm. Data is for T1-2, Transoner[R]) [10].

The frequency boundaries for soft switching is avaluated from Fig. 5 to be $1.003 < k < 1.025$ (or, from (17), $118770 < f < 121360\text{Hz}$). The power dissipation of the PT for this load factor and in this frequency range is 9% approximately

The power dissipation of the PT for this load factor and in this frequency range is 9% approximately.

The transfer function ko that corresponds to this frequency region is $0.8 > ko > 0.22$ (the same as in Example 1.).

Concequently, the range of peak output voltage that can be obtained (under soft switching conditions) $V_{out(p)} = ko \cdot V_{in}$ is $40V > V_{out(p)} > 11V$, and the range of the output power $P_{out} = \frac{V_{out(p)}^2}{2R_L}$ will be $6.1W > P_{out} > 0.46W$ respectively (Fig. 7).

The same prossedure can be carried out for different load resistences in the soft swithing range to obtain the whole range of output voltages and powers.

III. Simulation and Experimental Results

The proposed model was verified by simulations and experiments. Fig. 2 shows the switching timing diagram of the half-bridge inverter (Fig. 1) obtained by simulation. The operation frequency was assumed to be $f = 120kHz$ $(k = 1.014)$ and the load resistance $R_L = 130\Omega$ (Q=0.15). The calculated normalized charging time was $\Delta_r = 0.16$ while the simulated normalized charging time was $\frac{t_r}{T} \approx 0.167$ (Fig. 2).

Fig. 8 shows the experimental voltage curves under the same conditions as above. The experimental normalized charging time was found to be $\frac{t_r}{T} \approx 0.175$.

IV. Conclusions

This paper presents a comprehensive analysis of the of inherent soft switching capability of PTs. Closed form equations, developed in this study, estimate the load and the frequency boundaries that allow soft switching in power inverter built around a given PT.

A general procedure for maintaining soft switching in inductor-less half-bridge piezoelectric transformer inverter was studied analytically and verified by the simulation and the experiment. The analytical results were found to be in a good agreement with simulation and experiment.

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