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Class E ZVS Resonant Inverter with CLC Filter and PLL-Based Resonant Frequency Tracking for Ultrasonic Piezoelectric Transducer

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Abstract

This paper presents a Class E zero-voltage soft-switching (ZVS) resonant inverter integrated with a CLC filter and a digital resonant frequency tracking technique for driving a piezoelectric ceramic transducer (PZT) in ultrasonic cleaning applications. A digital signal processor (DSP) is used to dynamically monitor and adjust the operating frequency in response to slight variations in the cleaning load, employing a phase-locked loop (PLL) control scheme. The proposed method ensures that the inverter maintains ZVS operation across a frequency range from 30.0 kHz to 34.0 kHz, thereby improving energy efficiency and reducing switching losses. The system is capable of delivering a stable power output of 100 W. Both the simulation and experimental results validate the effectiveness of the proposed technique, demonstrating improved performance under varying load conditions. The combination of CLC filtering and frequency tracking offers a compact and robust solution suitable for ultrasonic cleaner systems and similar resonant-load applications.

Keywords: zero-voltage soft switching (ZVS); class E resonant inverter; piezoelectric ceramic transducer (PZT); phase-locked loop (PLL); ultrasonic cleaning system



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1. Introduction

Ultrasonic cleaning is a well-established technology widely used for removing contaminants from the surfaces of objects, particularly in areas that are difficult to access using conventional methods [1–7]. A piezoelectric ceramic transducer (PZT) is typically used to convert electrical energy into high-frequency ultrasonic waves, generally above 20 kHz, which generates microscopic cavitation bubbles in a liquid medium to dislodge and eliminate dirt and particles from the target surface [3,6,8–12]. This technique enables efficient and thorough cleaning without requiring many harmful chemicals, making it suitable for delicate or intricate items.

Ultrasonic cleaners are employed across various industries, including medical device sterilization, electronic component cleaning, and the maintenance of optical instruments, like camera lenses, microscopes, and telescope parts. To efficiently drive the PZTs in such systems, resonant inverters are widely developed and optimized. Several inverter topologies, such as full- and half-bridge and Class D and E resonant inverters, have been investigated for ultrasonic applications [6,13–17]. Among them, the Class E resonant inverter is especially attractive due to its high efficiency, reduced electromagnetic interference

(EMI), low conduction losses, minimal component count, and its ability to maintain ZVS under ideal conditions [13,18–22].

However, maintaining ZVS becomes challenging when the system is subjected to variations in load or changes in the resonant frequency of the PZT. For instance, the predictive current control approach proposed in [4] for a Class E inverter fails to compensate for such frequency shifts, leading to operation outside the ZVS condition and a drop in overall efficiency. This issue is compounded by the narrow resonant and anti-resonant frequency range of the PZT, which limits the effective control margin of conventional inverters. To address this limitation, prior work has proposed extending the operating range of the inverter using a series inductive filter in a Class D topology [6]. More recently, CLC filters have been employed in power electronics to suppress EMI, enhance power quality, and reduce switching losses [11,23–26]. In addition, digital control techniques such as PLL algorithms have proven effective for dynamically tracking resonant frequency variations in real time [27–30].

This paper proposes an enhanced Class E ZVS resonant inverter architecture that incorporates a CLC filter and a PLL-based resonant frequency tracking algorithm implemented on a digital signal processor (DSP). The proposed system is designed to ensure continuous ZVS operation even under dynamic load conditions, thereby improving the reliability and efficiency of ultrasonic cleaners. The remainder of this paper is organized as follows: Section 2 describes the overall system architecture and circuit topology. Section 3 provides a detailed analysis of the resonant circuit. Section 4 presents the hardware implementation along with the simulation and experimental results. Finally, Section 5 concludes the findings and discusses the advantages of the proposed approach.

2. Proposed System

Figure 1 illustrates the overall architecture of the proposed ultrasonic cleaning system. It is composed of three main stages: a full-bridge rectifier, a Class E ZVS resonant inverter, and an ultrasonic cleaning tank equipped with a PZT. The input AC voltage v_i , typically of the mains at a frequency of 50 Hz, is first converted into a DC voltage V_{dc} by the full-bridge rectifier. To reduce the ripple in the DC supply and maintain a steady current, a high-value choke inductor L_1 is placed after the rectifier. This ensures smooth energy delivery to the inverter stage. The DC voltage is then fed into the Class E ZVS resonant inverter, which converts it into a high-frequency AC signal with a switching frequency f_s in the range of 25–35 kHz. This frequency range is carefully chosen to match the resonant characteristics of the PZT, thereby maximizing the efficiency of ultrasonic energy generation. The inverter drives the PZT mounted beneath or within a water tank. When excited at its resonant frequency, the PZT produces ultrasonic vibrations that propagate through the liquid medium and generate cavitation bubbles. These bubbles collapse near the surface of the immersed workpiece, effectively dislodging contaminants in a non-invasive manner.

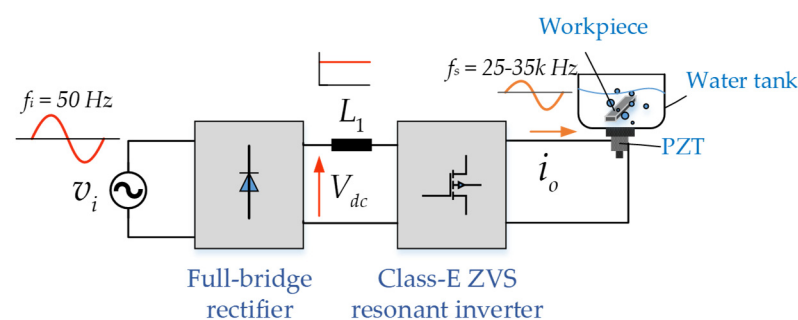


Figure 1. Ultrasonic cleaning system.

The proposed configuration leverages the advantages of Class E ZVS switching—such as reduced switching losses, lower EMI, and high efficiency—while maintaining stable operation under varying load conditions. To achieve this, a PLL algorithm implemented on a DSP continuously tracks the resonant frequency of the PZT and adjusts the switching frequency accordingly. This system is particularly suitable for high-performance ultrasonic cleaning applications where both energy efficiency and cleaning effectiveness are critical.

Figure 2 presents the detailed schematic of the proposed Class E ZVS resonant inverter system integrated with a CLC filter, specifically designed for driving a PZT in ultrasonic cleaning applications. The system is powered by a DC supply voltage V_{dc} , which is typically obtained from a full-bridge rectifier, as shown earlier in Figure 1. A choke inductor L_1 is connected in series to suppress voltage ripple and maintain a relatively constant current I_{dc} flowing into the inverter stage. The inverter stage consists of a single MOSFET switch Q_1 operating under ZVS conditions. A parallel capacitor C_1 is connected across the MOSFET to facilitate soft switching and reduce voltage stress during turn-off transitions. When Q_1 is switched ON and OFF at a high frequency (typically 25–35 kHz), it generates a quasi-sinusoidal current waveform i_o that flows into the resonant-load network. The CLC filter—composed of a series inductor L_2 , a series capacitor C_2 , and another shunt capacitor C_3 —functions as a resonant matching network between the inverter and the PZT. This network not only improves power transfer efficiency, but also shapes the output waveform and filters out unwanted harmonics, resulting in a smoother output voltage v_o across the load.

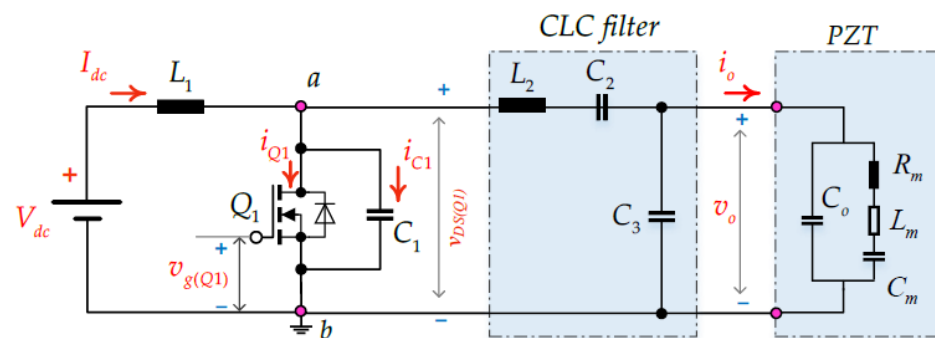


Figure 2. The Class E ZVS resonant inverter.

On the output side, the PZT is modeled as an equivalent circuit containing a motional branch with inductance L_m , resistance R_m , and capacitance C_m , representing the mechanical vibration dynamics of the transducer. A parallel capacitor C_o accounts for the static capacitance of the PZT structure. The interaction of these components determines the resonant or anti-resonant behavior of the transducer. By precisely adjusting the switching frequency of Q_1 to match the PZT's resonant frequency using a DSP-controlled PLL, the system ensures maximum energy transfer and maintains the ZVS condition under varying load scenarios. This circuit configuration combines the simplicity and high efficiency of Class E operation with the filtering and impedance matching benefits of the CLC network, making it well-suited for compact and energy-efficient ultrasonic cleaning systems.

Figure 3 presents physical and simplified electrical models of the piezoelectric ceramic transducer (PZT) used in the proposed system. Figure 3a shows the actual structure of the PZT used in the experimental prototype. The device has a cylindrical form factor, with a diameter of 5.4 cm and a height of 5.3 cm. It is designed to convert high-frequency electrical signals into ultrasonic vibrations for cleaning applications. Figure 3b illustrates the detailed equivalent circuit of the PZT, which includes a motional branch consisting of the elements L_m , R_m , and C_m . These elements represent the mechanical characteristics of the transducer: L_m models inertia (mass), R_m accounts for energy dissipation (mechanical

loss), and C_m corresponds to elasticity (compliance). This motional branch is connected in parallel with static capacitance C_o , representing the dielectric property of the transducer material. Figure 3c shows a simplified steady-state equivalent model. Under resonance, motional impedance becomes minimal, allowing for the PZT to be approximated as a parallel combination of C_o and R_m . This simplification is useful for impedance matching and circuit analysis when designing the resonant inverter and the *CLC* filter.

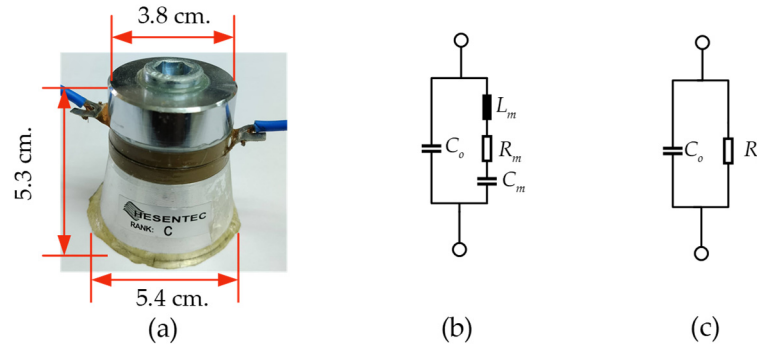


Figure 3. The ultrasonic piezoceramic transducer: (a) physical structure of the PZT used in the prototype, with a diameter of 5.4 cm and a height of 5.3 cm; (b) a full equivalent circuit model including motional components L_m , R_m , and C_m in parallel with static capacitance C_o ; and (c) a simplified steady-state model, where the motional branch is approximated by a parallel combination of C_o and R_m under resonant conditions.

The design of the inverter to control the PZT load must be based on the impedance and phase graph of the piezoceramic transducer (PZT), which was previously presented in reference [31], as shown in Figure 2. Two frequencies occur: resonant (f_r) and anti-resonant frequencies (f_a). When the switching frequency (f_s) of the inverter operates in the range of $f_r < f_s < f_a$, the PZT load is inductive, causing a lagging phase. In this case, the inverter operates under ZVS conditions. However, if the frequency is $f_s < f_r$ or $f_s > f_a$, the PZT load is capacitive, causing a leading phase. In this case, the inverter operates under non-ZVS conditions, which can damage the switches. The measured frequencies of the PZT load are $f_r = 28$ kHz and $f_a = 36$ kHz. Therefore, to ensure the inverter operates under ZVS conditions, the frequency f_s should be in the range from 28 kHz to 36 kHz.

Figure 4 presents the step-by-step equivalent circuit transformation of the *CLC* filter combined with the PZT load, which is necessary for accurate impedance modeling and resonant frequency analysis of the system. Figure 4a shows the original configuration, where the *CLC* filter—comprising L_2 , C_2 , and C_3 —is connected to the PZT load modeled by a parallel combination of static capacitance C_o and motional resistance R_m . For analytical purposes, capacitance C_o is combined with C_3 , as shown in the next stage. Figure 4b simplifies the parallel capacitors C_3 and C_o into a single equivalent capacitance, denoted as C_a , using the following equation:

$$C_a = C_3 + C_o. \quad (1)$$

Figure 4c converts the parallel combination of C_a and R_m into an equivalent series *RC* circuit defined by the parameters R_x and C_x , given by the following equation:

$$R_x = \frac{R_m}{1 + \omega^2 R_m^2 C_a^2} \quad (2)$$

$$C_x = \frac{1 + \omega^2 R_m^2 C_a^2}{\omega^2 R_m^2 C_a}. \quad (3)$$

where $\omega = 2\pi f$ is the angular frequency of operation.

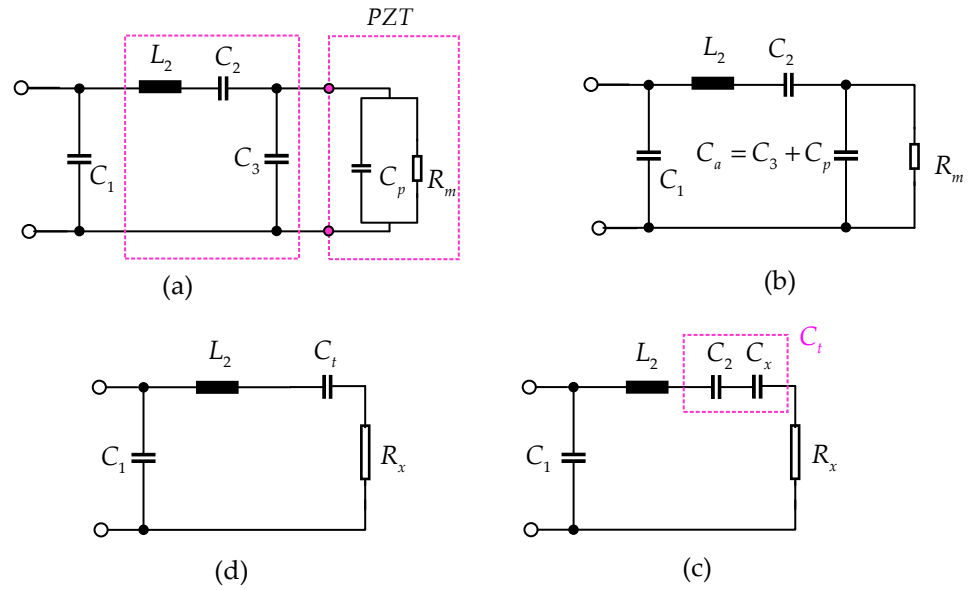


Figure 4. Equivalent circuits: (a) original configuration of CLC filter and PZT; (b) reduction in parallel capacitors C_3 and C_4 ; (c) transformation of parallel C_a and R_m into equivalent series components C_2 and C_x ; and (d) final series-resonant circuit composed of L_2 , C_t , and R_x .

Figure 4d further reduces the combination of series capacitors C_2 and C_x into a single effective capacitance C_t , using the following standard series formula:

$$C_t = \frac{C_2 C_x}{C_2 + C_x}. \quad (4)$$

All the variables represented in Figure 4 are listed in Appendix A.

3. Circuit Analysis

- The analysis of the proposed Class E ZVS resonant inverter is based on the following assumptions:
- The power MOSFETs operate as ideal switches.
- The quality factor Q_L of the series resonant load is sufficiently high, ensuring a nearly sinusoidal output current i_o .
- The input current I_{dc} is considered constant due to the large choke inductor L_1 .

Figure 5 illustrates the equivalent circuit of the Class E ZVS resonant inverter. High-frequency switching is performed by the power switch Q_1 , with its gate signal $v_{g(Q1)}$. A shunt capacitor C_1 enables ZVS by shaping the voltage waveform across the MOSFET. The series LC branch consisting of L_2 and C_t , followed by the resistive component R_x , represents the load network connected to the ultrasonic transducer.

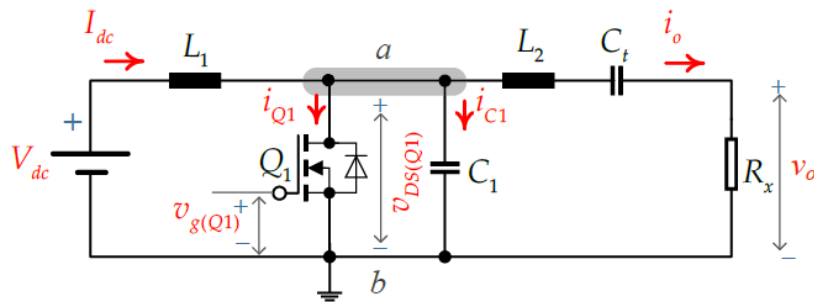


Figure 5. The equivalent circuit of a Class E ZVS resonant inverter.

Figure 6 shows the key voltage and current waveforms within one switching cycle T_s . The waveforms include:

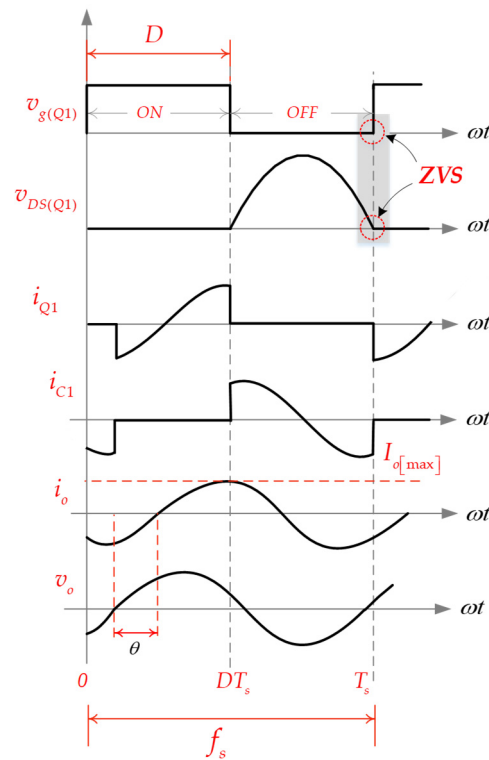


Figure 6. The key waveforms for a Class E ZVS resonant inverter.

- $v_{g(Q1)}$: gate drive signal of the MOSFET.
- $v_{DS(Q1)}$: drain–source voltage of Q_1 .
- i_{Q1} : current through switch Q_1 .
- i_{C1} : current through the shunt capacitor C_1 .
- i_o : output current delivered to the load.
- v_o : voltage across the load.

The output current i_o of the resonant circuit is assumed to follow a sinusoidal form:

$$i_o = I_{o[\max]} \sin(\omega t + \theta) \quad (5)$$

where $I_{o[\max]}$ is the peak output current, $\omega = 2\pi f_s$ is the angular frequency, and θ is the phase angle. The switching behavior of Q_1 is characterized as follows:

- Q_1 is turned ON when the capacitor current i_{C1} becomes zero at $\omega t = 0$, ensuring ZVS.
- During the ON interval $0 < \omega t < DT_s$, the current through Q_1 , i_{Q1} , is

$$i_{Q1} = \begin{cases} I_{dc} - I_{o[\max]} \sin(\omega t + \theta), & \text{for } 0 < \omega t < DT_s \\ 0, & \text{for } DT_s < \omega t < T_s. \end{cases} \quad (6)$$

During the OFF interval $DT_s < \omega t < T_s$, the current flows through capacitor C_1 is

$$i_{C1} = \begin{cases} 0, & \text{for } 0 < \omega t < DT_s, \\ I_{dc} - I_{o[\max]} \sin(\omega t + \theta), & \text{for } DT_s < \omega t < T_s. \end{cases} \quad (7)$$

Applying Kirchhoff's Current Law at node a in Figure 5, the total output current is

$$i_o = I_{dc} - (i_{C1} + i_{Q1}) \quad (8)$$

which can be rearranged to

$$i_{C1} + i_{Q1} = I_{dc} - i_o \quad (9)$$

Substituting Equation (5) into Equation (9) and analyzing at the boundary where $i_{C1} = 0$, the peak output current becomes

$$I_{o[\max]} = \frac{I_{dc}(T_s(1-D))}{\cos(DT_s + \theta) - \cos \theta}. \quad (10)$$

The output power of the resonant circuit is given by

$$P_o = R_x \frac{(I_{o[\max]})^2}{2}. \quad (11)$$

Finally, substituting Equation (10) into Equation (11), the output power of the Class E ZVS resonant inverter becomes

$$P_o = \frac{R_x}{2} \left(\frac{I_{dc}(T_s(1-D))}{\cos(DT_s + \theta) - \cos \theta} \right)^2. \quad (12)$$

This analytical model enables the evaluation of inverter performance, such as current stress, switching conditions, and output power, under varying duty cycles and frequency shifts.

The analysis of power losses of the proposed inverter for the ultrasonic cleaning application, based on the Class E ZVS resonant inverter shown in Figure 5, can be represented by the equivalent circuit of a Class E resonant inverter with parasitic resistances, as shown in Figure 7. In the operation of a Class E ZVS resonant inverter, the parasitic resistances of circuit components play a decisive role in determining overall efficiency. In particular, the power switch exhibits intrinsic on-resistance, which leads to conduction losses. These losses not only degrade the efficiency of the system but also cause thermal accumulation and may compromise the ZVS condition. Moreover, the parasitic resistances of the inductor and capacitor, represented by their equivalent series resistances (ESRs), contribute significantly to power dissipation. Since large resonant currents flow through these passive elements, the ESRs of both the inductor and the capacitors become the dominant factor responsible for the noticeable reduction in inverter efficiency.

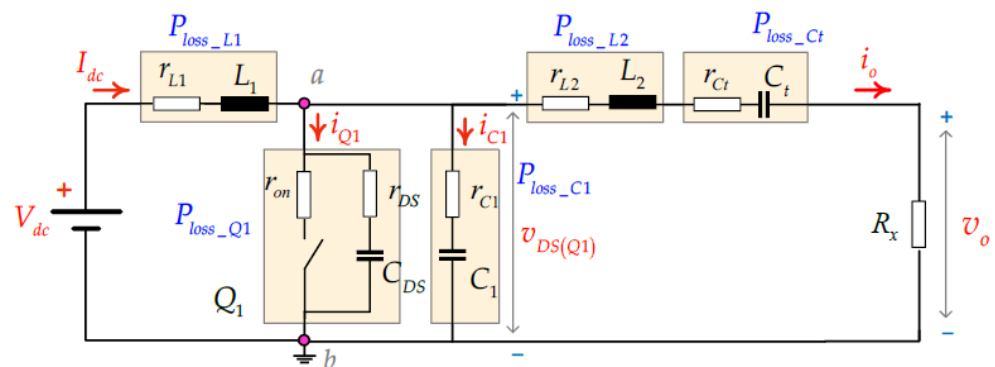


Figure 7. The equivalent circuit of a Class E resonant inverter with parasitic resistances.

The power losses of a Class E ZVS resonant inverter for the ultrasonic cleaning application consist of three components: (1) power loss due to the parasitic resistance of the power MOSFET, (2) switching loss during the turn-on of the power MOSFET, and (3) power losses in the inductor and the capacitor filter.

- (1) The power losses due to the parasitic resistance of the power MOSFET Q_1 occur in two stages: the power loss when the power MOSFET Q_1 conducts current depends on the MOSFET on-resistance, r_{on} , and the power loss due to the internal resistance (ESR or equivalent series resistance) of the capacitance between the drain and source in the MOSFET Q_1 , r_{DS} . The power loss equations for both cases can be expressed as follows:

$$P_{loss_ron} = \frac{1}{T_s} \int_0^{T_s} r_{on} i_{Q1}^2 d\omega t = r_{on} i_{Q1_rms}^2 \quad (13)$$

$$P_{loss_rDS} = \frac{1}{T_s} \int_0^{T_s} r_{DS} i_{Q1}^2 d\omega t = r_{DS} i_{Q1_rms}^2. \quad (14)$$

- (2) The switching loss is caused by energy stored in the parasitic capacitance during the transition of the switch from the off state to the on state. The total loss is calculated as the average of the energy stored during the switching cycle, which affects the circuit's efficiency, especially at high switching frequencies. This loss can be reduced by improving the control of the switching process. Therefore, the power loss during the turn-on of the power MOSFET Q_1 is

$$P_{loss_SW} = \frac{W_{turn-on}}{T_s} \quad (15)$$

where the energy stored during the switching cycle is

$$W_{turn-on} = \int_0^{v_{DS_on}} C_1 v_{DS} d\omega t = C_1 V_{DS_rms}.$$

- (3) The power losses caused by the parasitic resistances of the inductor and capacitors include the power loss of inductor L_1 , P_{loss_L1} ; power loss of capacitor C_1 , P_{loss_C1} ; power loss of inductor L_2 , P_{loss_L2} ; and power loss of capacitor C_t , P_{loss_Ct} . Therefore, the power losses due to the parasitic resistances of the inductor and capacitors are expressed by the following equations:

$$P_{loss_L1} = r_{L1} I_{dc}^2 \quad (16)$$

where r_{L1} is the parasitic resistance of the inductor L_1 .

$$P_{loss_C1} = \frac{1}{T_s} \int_0^{T_s} r_{C1} i_{C1}^2 d\omega t = r_{C1} i_{C1_rms}^2 \quad (17)$$

where r_{C1} is the parasitic resistance of the capacitor C_1 .

$$P_{loss_LC} = \frac{1}{T_s} \int_0^{T_s} (r_{L2} + r_{Ct}) i_o^2 d\omega t = (r_{L2} + r_{Ct}) I_{o_rms}^2 \quad (18)$$

where r_{L2} is the parasitic resistance of the inductor L_2 , and r_{Ct} is the parasitic resistance of the capacitor C_t .

Therefore, the total power losses of the Class E resonant inverter can be calculated as follows:

$$P_{t_loss} = P_{loss_ron} + P_{loss_rDS} + P_{loss_SW} + P_{loss_L1} + P_{loss_C1} + P_{loss_LC}. \quad (19)$$

Consequently, the efficiency of the Class E resonant inverter can be determined as follows:

$$\eta = \frac{P_o}{P_o + P_{t_loss}}. \quad (20)$$

The system efficiency calculation is presented in Appendix A.

4. Implementation and Results

4.1. Implementation

Figure 8 illustrates the control architecture of the proposed ultrasonic cleaning system, in which the Class E ZVS resonant inverter with a CLC filter drives the PZT-based ultrasonic cleaner. A TMS320F28335 digital signal processor (DSP) is used as the main controller and is manufactured by Texas Instruments, which is based in Dallas, Texas, USA. This device operates at a clock speed of 150 MHz with a 32-bit fixed-point architecture and supports a 10 kHz sampling rate, providing real-time control over the inverter's switching operation.

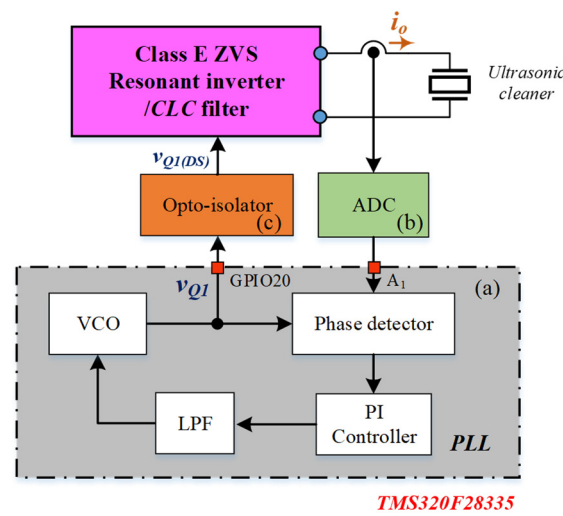


Figure 8. Control diagram of proposed system: (a) resonant frequency tracking controller using PLL control; (b) analog-to-digital converter (ADC) conditioning board; and (c) opto-isolator circuit.

To maintain optimal ZVS conditions under varying load conditions—such as changes in water volume and cleaning load—the system employs a PLL algorithm. The PLL block consists of a phase detector, a PI controller, a low-pass filter (LPF), and a voltage-controlled oscillator (VCO). Together, these modules dynamically track the resonant frequency of the PZT and adjust the inverter switching frequency accordingly.

The zero-crossing detector circuit shown in Figure 9 is used to monitor the output current i_o and detect its phase. This analog signal is conditioned using the TLV7011 comparator (U1) from Texas Instruments, which is based in Dallas, Texas, USA. The circuit detects the zero-crossing point of the sinusoidal i_o waveform and converts it into a digital pulse signal, which is sent to the DSP via its analog input pin A1.

Figure 9 shows the zero-crossing detector circuit, applying an op-amp comparator (TLV7011) to detect the output current of the Class E resonant inverter, characterized by a sinusoidal waveform or AC signal from a current sensor of LAM (LA 100-P) from California, USA, to execute zero-crossing and level-shift the voltage to retain only the positive half. From the circuit, R_1 (100 k Ω) and R_2 (4.7 k Ω) are used as a voltage divider to attenuate the AC signal before it is sent to the non-inverting input of the comparator. D_1 represents a diode that keeps the non-inverting input of the comparator from falling below the minimum allowed value dictated by the negative input common-mode voltage limit. The output of the zero-crossing detection circuit, V_{ref_A} , is a pulse signal that switches

between 0 V and 3.3 V at the frequency f_s . This signal is then sent to the A_1 pin of the TMS320F28335 DSP.

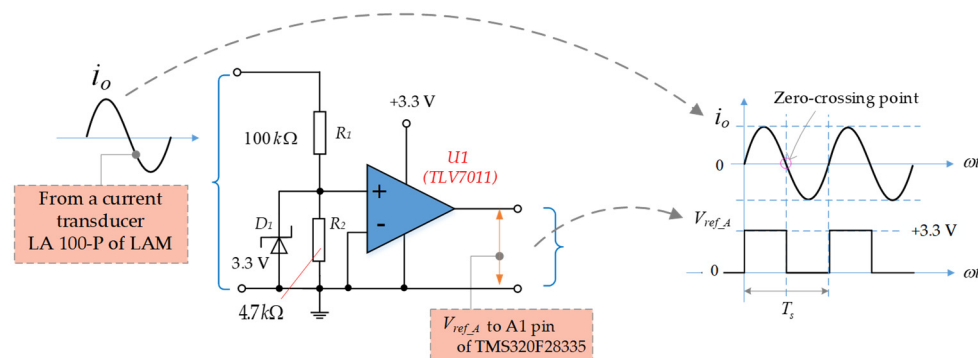


Figure 9. The circuit for the zero-crossing detector.

Figure 10 presents the opto-isolation circuit used to safely transmit gate control signals to the power stage. The opto-isolator TLP250 (U2) from Toshiba (Electronic Source Company, Bangkok, Thailand) is used to provide electrical isolation between the low-voltage control and high-voltage switching sides. The circuit is powered by a +15 V supply and includes appropriate current-limiting resistors for reliable operation.

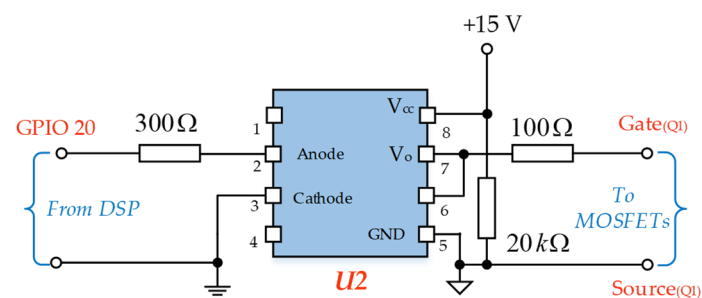


Figure 10. Opto-isolation circuit.

To characterize the PZT used in the system, an LCR meter (HIOKI 3532-50) was employed from Electronic Source Company, Bangkok, Thailand. The key parameters of the transducer, including the mechanical and electrical components, are listed in Table 1. These values were used in both the simulation and the experimental analysis.

Table 1. PZT parameters.

Parameters	Value	Unit
R_m	320.05	Ω
L_m	85.25	mH
C_m	170.15	pF
C_o	2.52	nF

Figure 11 shows a process diagram of the system implemented to maintain the frequency variations for the load in ultrasonic cleaning. The procedure started by acquiring the output current i_o data from the sensor connected to the ADC (analog-to-digital converter) for subsequent processing. Subsequently, the system computes the phase detection or phase angle from the acquired data. The system subsequently verifies if the phase angle (θ) exceeds 0. If the phase angle is not zero, the system adjusts it to zero and reprocesses the data. When the phase angle is zero, the system calculates the PI controller using a

loop filter for enhanced control precision. Afterwards, the system computes the VCO (voltage-controlled oscillator), which controls frequency regulation. Finally, the system produces a PWM (Pulse Width Modulation) signal to regulate its functioning, thereby concluding the process.

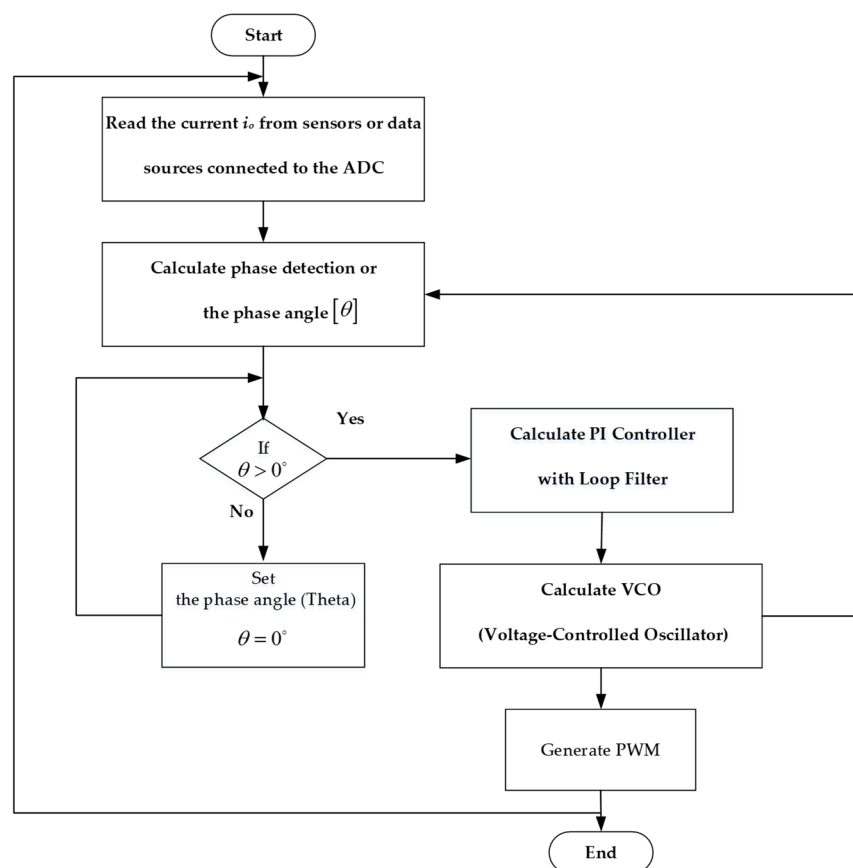


Figure 11. Algorithm resonant frequency tracking flowchart for Class E resonant inverter using phase-locked loop.

To validate the operational principles and theoretical analysis of the proposed Class E ZVS resonant inverter system, a laboratory prototype was developed, as shown in Figure 12. Figure 12a depicts the physical implementation of the inverter circuit. The system includes the power switch Q_1 , implemented using an IRFP460 MOSFET from Electronic Source Company, Bangkok, Thailand, along with the resonant elements and the CLC filter components C_1 , C_2 , C_3 , and L_2 . The circuit is controlled by a TMS320F28335 DSP, with the driver and isolation circuitry (including U1 and U2) positioned on a compact test board. Figure 12b shows the experimental ultrasonic cleaning tank, where a PZT is mounted at the bottom of a stainless steel container (dimensions: 20 cm × 13 cm). The transducer is driven at resonance to generate ultrasonic waves for cleaning applications.

This setup was tested with various cleaning targets, including printed circuit boards, jewelry, stainless steel parts, plastic components, and electrical contacts, demonstrating the practical effectiveness and versatility of the proposed system in real-world ultrasonic cleaning tasks.

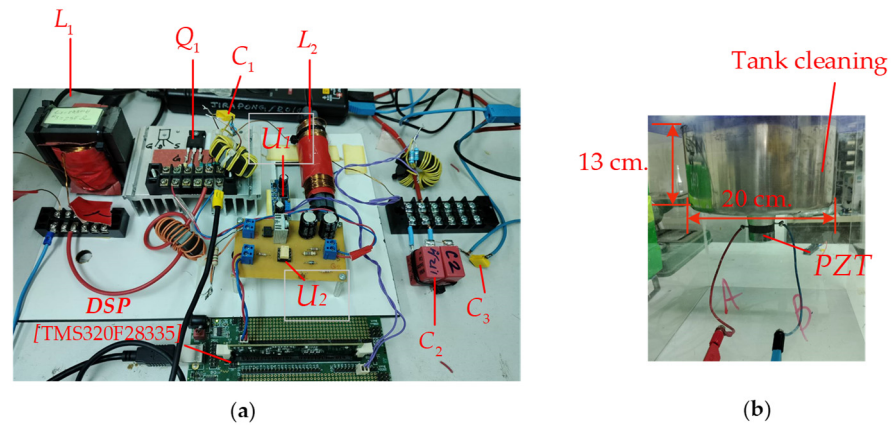


Figure 12. A laboratory prototype of the proposed system. (a) The Class E ZVS resonant inverter with the CLC filter. (b) Using the ultrasonic cleaner on the tank.

4.2. Results

To verify the performance of the proposed Class E ZVS resonant inverter with CLC filtering, both the simulation and experimental results were analyzed under varying load conditions. Figure 13 presents the simulated waveforms at a switching frequency $f_s = 30$ kHz and a duty cycle $D = 0.5$, demonstrating proper ZVS operation. The gate–source voltage $v_{g(Q1)}$ is shown in Figure 13a, while the corresponding drain–source voltage $v_{DS(Q1)}$, peaking at 250 V, is depicted in Figure 13b. The output current i_o and the switch current i_{Q1} both reach a maximum of 0.85 A, as shown in Figure 13c,d, respectively. The output voltage v_o , across the load, reaches 130 V peak as illustrated in Figure 13e. The shunt capacitor current i_{C1} , shown in Figure 13f, peaks at 0.75 A. Under these conditions, the simulated output power P_o is approximately 90 W.

Figure 14 illustrates the experimental waveforms of the inverter operating at $f_s = 32.34$ kHz and a duty cycle $D = 0.5$, without employing the resonant frequency tracking control. In this case, as the liquid level in the tank decreases, the resonant frequency of the PZT increases. Since the inverter is not tracking this shift, it begins to operate under a non-ZVS condition, which is clearly visible in the zoomed region of the waveform. The observed oscillations in the current i_{Q1} lead to increased switching losses and reduced efficiency.

To address this, Figure 15 presents the experimental results with the PLL-based resonant frequency tracking enabled, operating at a frequency $f_s = 30.4$ kHz. Figure 15a shows the 3.3 V gate–source voltage $v_{g(Q1)}$, while Figure 15b displays the output voltage v_o , which peaks at 250 V with an RMS value of 176 V. The output current i_o , shown in Figure 15c, reaches a peak of 0.85 A with an RMS value of 0.601 A. The switch current i_{Q1} also peaks at 0.85 A, as shown in Figure 15d, and the shunt capacitor current i_{C1} peaks at 0.75 A, as shown in Figure 15f. The measured output power in this case is approximately 94.5 W, and the liquid solvent volume in the tank was maintained at 3 L.

These results confirm that the proposed system, when combined with PLL-based resonant frequency tracking, can maintain ZVS operation and deliver stable output power despite variations in the load conditions, thereby improving overall efficiency and system reliability.

The experimental results of the Class E ZVS resonant inverter operating under PLL-based frequency tracking at a switching frequency $f_s = 32.4$ kHz are shown in Figure 16. These results correspond to conditions with a dynamically altered load, where the inverter successfully maintains ZVS operation. The gate–source voltage $v_{g(Q1)}$, as shown in Figure 16a, with a duty cycle $D = 0.5$, reaches a peak of 3.3 V. The drain–source voltage $v_{DS(Q1)}$ peaks at 250 V, as shown in Figure 16b, confirming proper switching. The output

current i_o , as shown in Figure 16c, and the current i_{Q1} , as shown in Figure 16d, both reach a maximum value of 0.85 A.

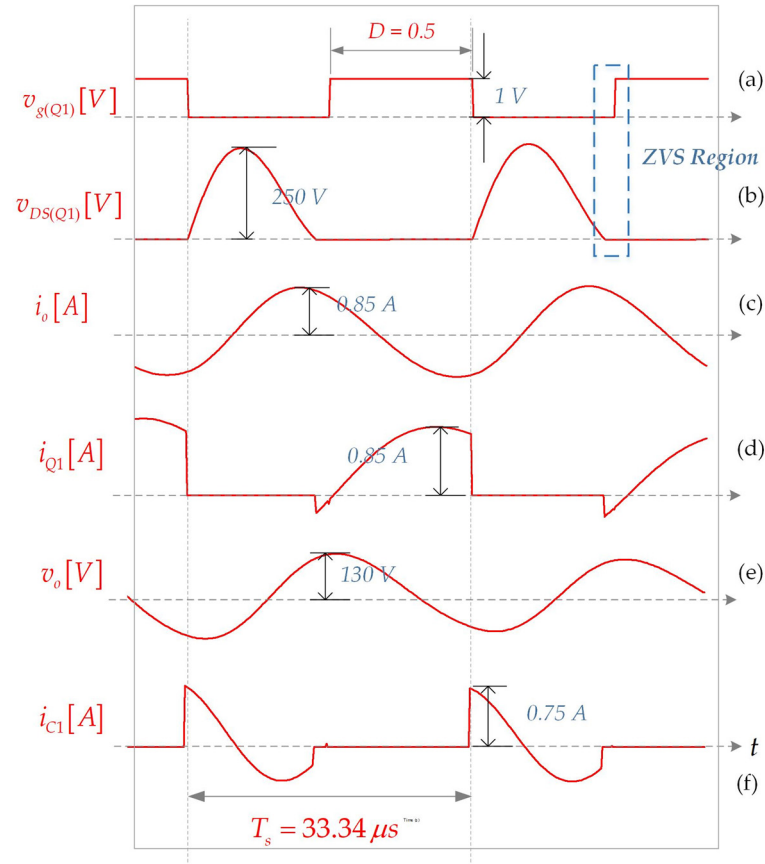


Figure 13. Simulation waveforms of the Class E ZVS resonant inverter at duty $D = 0.5$: (a) gate–source voltage $v_{g(Q1)}$; (b) switch voltage $v_{DS(Q1)}$; (c) output current i_o ; (d) switch current i_{Q1} ; (e) output voltage v_o ; and (f) shunt capacitor current i_{C1} .

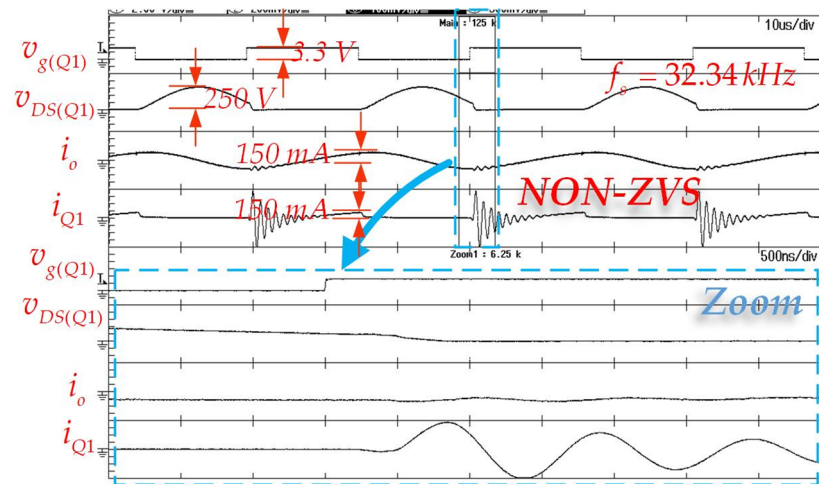


Figure 14. The non-ZVS operation under load variations.

The oscillatory behavior of i_{Q1} , as shown in Figure 16d when the diode conducts, is caused by the excitation of the parasitic L–C network, the reverse recovery charge (Q_{rr}), and the large dv/dt at commutation. Such ringing is typical of high-frequency converters; careful layout and appropriate damping networks can suppress it so that has a negligible impact on performance and EMI.

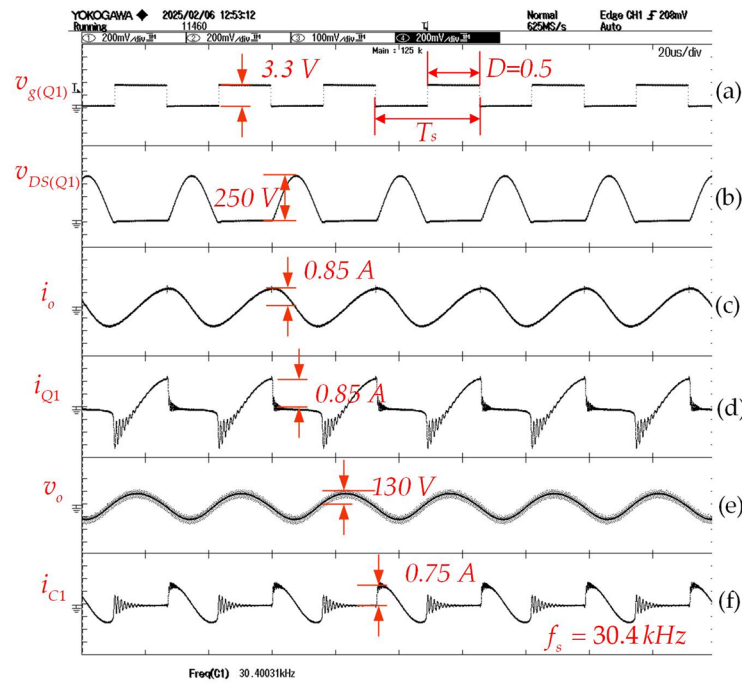


Figure 15. The experimental results of the proposed system at duty $D = 0.5$: (a) gate–source voltage $v_{g(Q1)}$; (b) switch voltage $v_{DS(Q1)}$; (c) output current i_o ; (d) switch current i_{Q1} ; (e) output voltage v_o ; and (f) shunt capacitor current i_{C1} .

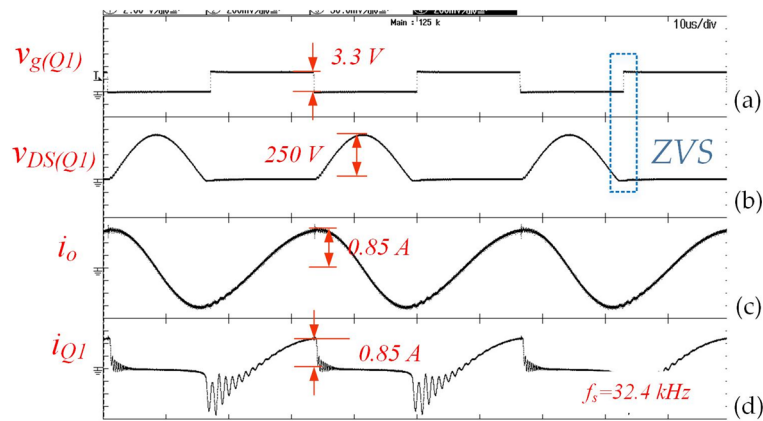


Figure 16. The experimental results of the proposed system at the duty $D = 0.5$ under load variations: (a) gate–source voltage $v_{g(Q1)}$; (b) switch voltage $v_{DS(Q1)}$; (c) output current i_o ; and (d) switch current.

At a liquid solvent volume of 2 L, the system delivers a measured output power value P_o of approximately 93.25 W. These results validate the effectiveness of the proposed control strategy in maintaining ZVS and stable power delivery under varying load conditions.

Figure 17 presents the experimental results of the Class E ZVS resonant inverter operating with PLL-based frequency tracking at a switching frequency f_s of 32.98 kHz under light-load conditions. These results demonstrate the inverter's ability to maintain ZVS operation despite dynamic load variations. With a duty cycle D of 0.2, the gate–source voltage $v_{g(Q1)}$ reaches a peak of 3.3 V, while the drain–source voltage $v_{DS(Q1)}$ peaks at 100 V, confirming proper switching performance. The output current i_o reaches a maximum of 0.38 A. Under a liquid solvent volume of 1 L, the system delivers an output power P_o of approximately 40.5 W. These findings validate the effectiveness of the proposed control strategy in sustaining ZVS operation and ensuring stable power delivery under fluctuating load conditions. The zoom area for the time per division on the horizontal axis, set at 1 μ s

as shown in Figure 17, can maintain the ZVS region, while the water level in the ultrasonic cleaning tank decreases.

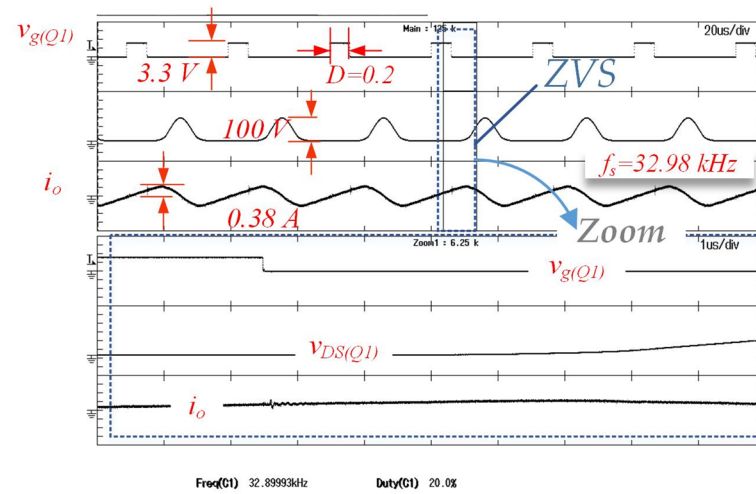


Figure 17. The experimental results of the proposed system at duty $D = 0.2$ under load variations.

Figure 18a shows the power adjustment at 40% of full load, which is approximately 40 W, with the liquid solvent level in the cleaning tank set to 1 L. The power then adjusts according to the water level increase. When the level reaches 3 L, the power reaches full load, which is 94.5 W. To confirm the operation under full load conditions, the experimental results of the $v_{DS(Q1)}$ voltage and i_o current are shown in Figure 18b. The experimental results of the $v_{DS(Q1)}$ voltage and i_o current under 40% load are also shown in Figure 18c. The results demonstrate that the proposed inverter operates under ZVS conditions and can effectively track frequency f_r according to load changes.

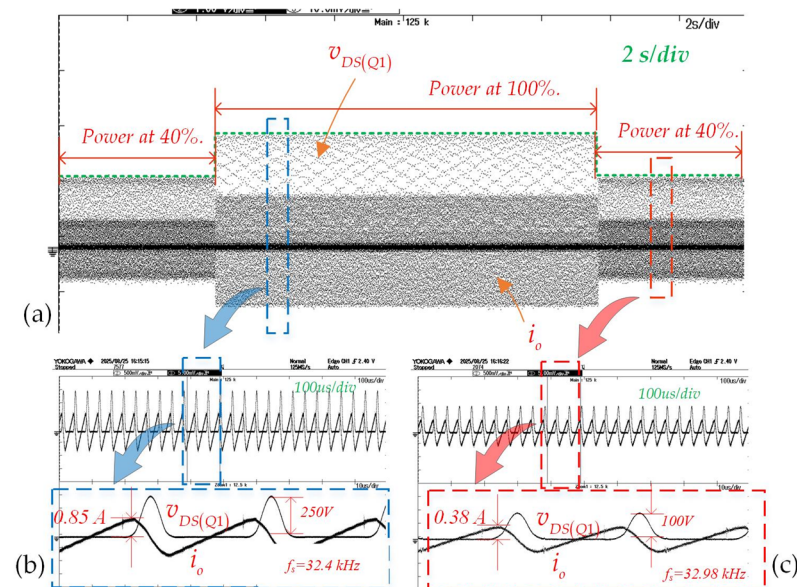


Figure 18. Experimental results of power adjustment according to load change in ultrasonic cleaning applications for Class E resonant inverter: (a) voltage waveform $v_{DS(Q1)}$ and current i_o at time scale of 2 s/div; (b) voltage waveform $v_{DS(Q1)}$ and current i_o for power at 100% at time scale of 10 μ s/div; and (c) voltage waveform $v_{DS(Q1)}$ and current i_o for power at 40% at time scale of 10 μ s/div.

5. Conclusions

This study presents an enhanced Class E ZVS resonant inverter integrated with a CLC filter for ultrasonic cleaning applications. The proposed system incorporates a DSP-based

gate drive controller, enabling fast and precise control of the switching operation. To ensure optimal performance under varying load conditions, a phase-locked loop (PLL) algorithm is implemented to dynamically track the resonant frequency of the piezoelectric transducer.

Both the simulation and experimental results confirm the effectiveness of the proposed approach in maintaining high efficiency and reliable operation. The key advantages of the system are summarized as follows:

1. The Class E ZVS resonant inverter successfully maintains zero-voltage switching (ZVS) across a range of resonant frequencies and varying load conditions.
2. The CLC filter effectively smooths the output voltage and current waveforms, improving waveform quality and reducing harmonic distortion.
3. The inverter achieves lower switching losses and offers a compact design, making it suitable for space-constrained applications.

In conclusion, the proposed inverter system demonstrates strong potential for use in ultrasonic cleaners and other applications involving resonant loads, where efficiency, adaptability, and compactness are critical.

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Appendix A

The design specifications for the Class E resonant inverter used in the ultrasonic cleaning system are based on the following operating conditions: the voltage V_{dc} is 250 V; the load quality Q_L is defined as 10; the switching frequency is set at 30 kHz; and the maximum output power of the PZT is 100 W.

The shunt capacitor C_1 is calculated using

$$C_1 = \frac{8}{\pi(4 + \pi^2)\omega R_x} = 8.45 \text{ nF}.$$

The combined parallel capacitor C_a , derived from the transformation of the PZT equivalent model, is

$$C_a = \sqrt{\frac{\left(\frac{R_m}{R_x} - 1\right)}{\omega_r^2 R_m^2}} = 8.7 \text{ nF}.$$

From Equation (1), the capacitor C_3 is obtained as

$$C_3 = C_a - C_0 = 6.17 \text{ nF}.$$

The series capacitor C_t in the resonant tank is calculated using

$$C_t = \frac{1}{L_2 \cdot \omega^2} = 36.8 \text{ nF}.$$

The parasitic resistance of capacitor C_t is calculated using

$$r_{Ct} = r_{C2} + r_{C3}.$$

The equivalent series capacitor C_x , from Equation (3), is

$$C_x = \frac{1 + \omega^2 R_m^2 C_a^2}{\omega^2 R_m^2 C_a} = 40.3 \text{ nF}.$$

Then, using the series capacitor transformation, C_2 is computed as

$$C_2 = \frac{C_x C_t}{C_x - C_t} = 424 \text{ nF}.$$

Using Equation (2), the equivalent resistive component R_x is derived as

$$R_x = \frac{8}{4 + \pi^2} \left(\frac{V_{dc}^2}{P_o} \right) = 14.42 \Omega.$$

Overall, the L_1 in a Class E resonant inverter tends to be fairly large because its primary role is to act as a near-constant-current source (an RF choke) that preserves the switch voltage and current waveforms required for reliable ZVS. At the chosen switching frequency, the reactance of L_1 must be much greater than the load's equivalent resistance, which drives a high inductance value. Since the circuit carries high peak and RMS currents, L_1 must store magnetic energy without saturating; this calls for a ferrite core with a sufficient cross-section and air gap, together with a large litz (stranded) wire to limit heating from skin and proximity effects. These requirements increase the physical size and weight of L_1 . A larger L_1 also helps keep ripple from flowing back into the DC bus and improves EMI behavior. If L_1 is too small, current ripple increases, the waveforms deviate, the ZVS window narrows or is lost, switch voltage overshoot grows, efficiency drops, and EMI worsens. Therefore, L_1 must be “large enough” to avoid these issues.

While increasing the switching frequency reduces the required inductance, and thus the size of L_1 , in ultrasonic applications, the output power is regulated by sweeping the drive frequency around the transducer's resonance. Therefore, L_1 must be sized to maintain RF-choke behavior across the entire control band, particularly at the lowest operating frequency where the inductive reactance, X_L is minimal.

As a result, the achievable size reduction is bounded by the control bandwidth, and higher f_s also entails increased magnetic core and switching losses, as well as greater EMI, which must be accounted for in the design.

Additionally, when increasing f_s , several key trade-offs must be considered:

- Core loss and AC copper loss in L_1 and the transformer rise with frequency, increasing the thermal burden.
- Switching loss, dv/dt stress, and EMI all increase, potentially requiring added snubbers/filters.

Thus, the choke inductor L_1 is

$$L_1 = 2 \left(1 + \frac{\pi^2}{4} \right) \left(\frac{R_x}{f_s} \right) = 3.34 \text{ mH}.$$

The series inductor L_2 for the CLC filter is

$$L_2 = \frac{Q_L R_x}{\omega_s} = 765 \mu\text{H}.$$

Table A1. Summary of proposed system parameters.

Name	Theory	Simulation	Experiment
L_1	3.34 mH	3.34 mH	3.5 mH
L_2	765 μH	765 μH	768 μH
C_1	8.45 nF	8.45 nF	8.5 nF
C_2	424 nF	424 nF	450 nF
C_3	6.17 nF	6.17 nF	6.5 nF

The parasitic resistances of the proposed system parameters, measured using an LCR analyzer (HIOKI3525 form Electronic Source Company, Bangkok, Thailand) at $f = 30$ kHz, are $r_{L1} = 2.75 \Omega$, $r_{L2} = 1.85 \Omega$, $r_{on} = 0.27 \Omega$, $r_{DS} = 0.22 \Omega$, $r_{C1} = 0.228 \Omega$, $r_{C2} = 0.0054 \Omega$, and $r_{C3} = 0.05 \Omega$.

Example of Power Losses Calculation and Efficiency Determination of a Class-E Resonant Inverter at $D = 0.5$

- The power loss of Q_1 due to conduction can be determined from Equation (13).

$$P_{loss_ron} = r_{on} i_{Q1_rms}^2 = 0.27 \left(0.85 / \sqrt{2}\right)^2 = 0.0975 \text{ W}.$$

- The power loss of the MOSFET drain–source capacitance ESR can be calculated using Equation (14).

$$P_{loss_rDS} = r_{DS} i_{Q1_rms}^2 = 0.22 \left(0.85 / \sqrt{2}\right)^2 = 0.0792 \text{ W}.$$

- The turn-on loss of the power MOSFET can be calculated using Equation (15).

$$P_{loss_SW} = \frac{W_{turn-on}}{T_s} = \frac{C_1 V_{DS_rms}}{T_s} = \frac{8.5 \text{ nF} \left(250 / \sqrt{2}\right)}{30 \times 10^{-6}} = 0.05 \text{ W}$$

where $T_s = \frac{1}{f_s} = \frac{1}{32.4 \text{ kHz}} = 30 \times 10^{-6} \text{ s}$.

- The power loss of inductor L_1 can be calculated from Equation (16) as follows:

$$P_{loss_L1} = r_{L1} I_{dc}^2 = 2.75 (1.05)^2 = 3.03 \text{ W}$$

where the measured current I_{dc} is 1.05 A.

- The power loss of capacitor C_1 can be calculated from Equation (17) as follows:

$$P_{loss_C1} = r_{C1} I_{C1_rms}^2 = 0.228 \left(0.75 / \sqrt{2}\right)^2 = 0.064 \text{ W}.$$

- Equation (18) provides the calculation of losses arising from the parasitic resistances of inductor L_2 and capacitor C_t , as shown below:

$$P_{loss_LC} = (r_{L2} + r_{Ct}) I_{o_rms}^2 = (1.65 + 0.0554) \left(0.85 / \sqrt{2}\right)^2 = 1.025 \text{ W}.$$

- Thus, the total power loss of the proposed system can be calculated from Equation (19) as follows:

$$P_{t_loss} = 0.0975 + 0.0792 + 0.05 + 3.03 + 0.064 + 1.025 = 4.3457 \text{ W.}$$

- Consequently, the efficiency of the Class E resonant inverter can be calculated from Equation (20) as follows:

$$\eta = \frac{P_o}{P_o + P_{t_loss}} = \frac{93.25}{93.25 + 4.3457} = 95.55\%.$$

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