

# Load Insensitivity Design for a 6.78 MHz kW-Level Inductive Power Transfer Using Bridge Topology

Xinlin Wang , Graduate Student Member, IEEE, Jun Yang , Yu Liu , Senior Member, IEEE, Junrui Liang , Senior Member, IEEE, Haoyu Wang , Senior Member, IEEE, and Minfan Fu , Senior Member, IEEE

**Abstract**—This article presents a design methodology for achieving load-insensitive operation in MHz inductive power transfer (IPT) systems based on bridge inverters. While MHz operation enables reduced coil size and improved efficiency compared with conventional kHz IPT, it also introduces challenges such as severe switching losses, junction capacitances, and harmonic distortion. To address these issues, a wide-band impedance model is established that incorporates device capacitances and transition times, from which a load-insensitive ZVS condition is analytically derived and used to guide the impedance matching network (IMN) design. A 1-kW, 6.78-MHz prototype is constructed to validate the proposed method, demonstrating nearly load-independent ZVS, stable output regulation across a tenfold load variation, and high efficiency with a peak system efficiency of 94.9% and a peak inverter efficiency of 98.8%. The results highlight the effectiveness of the proposed approach in bridging theoretical analysis and practical implementation, offering a systematic pathway toward high-power and high-efficiency MHz IPT systems.

**Index Terms**—GaN devices, impedance matching network (IMN), inductive power transfer (IPT), MHz bridge inverter, load-insensitive zero-voltage switching (ZVS), wide load range.

## I. INTRODUCTION

INDUCTIVE power transfer (IPT) has attracted extensive attention in recent decades owing to its contactless power delivery capability, high safety, and user convenience. It has been widely adopted in consumer electronics, biomedical implants, mobile robotics, and electric vehicle (EV) charging systems [1], [2]. Most commercial IPT products still operate in the kHz frequency range, benefiting from mature semiconductor devices

and well-established control strategies [3], [4], [5]. However, kHz IPT is inherently constrained by bulky passive components, limited power density, and poor scalability, making it unsuitable for emerging high-frequency and high-performance applications [6].

Recent advances in ultra-high- $Q$  self-resonant coils [7] and wide-bandgap (WBG) semiconductor devices [8] have enabled IPT systems to operate in the MHz frequency range. Higher operating frequencies reduce resonant component size and copper loss, thereby increasing efficiency and power density. Several MHz IPT prototypes have achieved power transfer from tens of watts up to the kilowatt level [9], extending applicability from portable devices to industrial and transportation systems [10]. Nevertheless, MHz operation introduces new challenges: device parasitics and finite transition times generate considerable switching losses, while harmonic distortion complicates impedance modeling and soft-switching design [11], rendering conventional kHz design methodologies inadequate.

In terms of circuit topology, single-switch inverters such as Class E and its derivatives have been widely investigated for MHz IPT owing to their intrinsic soft-switching capability and structural simplicity [12], [13], [14]. However, their performance deteriorates rapidly under load variations, and the high voltage/current stresses limit scalability toward higher power levels [15]. Load-tolerant topologies, including Class  $\Phi$  inverters [16] and impedance matching network (IMN)-based approaches [17], have been proposed to improve adaptability, but they often increase circuit complexity and voltage stress.

With the emergence of GaN and SiC devices, bridge inverters have become a promising alternative due to their superior power capability, reduced stress, and better scalability [18]. A 1-kW, 6.78-MHz prototype has achieved over 95% efficiency under a 50  $\Omega$  load [9], demonstrating the potential of bridge-based MHz IPT systems. Nevertheless, maintaining zero-voltage switching (ZVS) over a wide load range remains a major challenge. Auxiliary ZVS branches [19] and modulation-based techniques [20] can extend the ZVS range but inevitably introduce additional losses and reduce power density, while methods such as frequency modulation and dynamic dead-time control [21] require complex sensing and control loops, increasing system cost and design difficulty.

Received 27 December 2025; revised 2 March 2026; accepted 26 March 2026. Date of publication 8 May 2026; date of current version 16 October 2026. This work was supported by the National Natural Science Foundation of China under Grant 52477013 and Lingang Laboratory under Grant LGL-5925-03. (Xinlin Wang and Jun Yang are co-first authors.) (Corresponding author: Minfan Fu.)

The authors are with ShanghaiTech University, Shanghai 201210, China (e-mail: fumf@shanghaitech.edu.cn).

Digital Object Identifier 10.1109/TIE.2026.3682399

1557-9948 © 2026 IEEE. All rights reserved, including rights for text and data mining, and training of artificial intelligence and similar technologies. Personal use is permitted, but republication/redistribution requires IEEE permission. See <https://www.ieee.org/publications/rights/index.html> for more information.

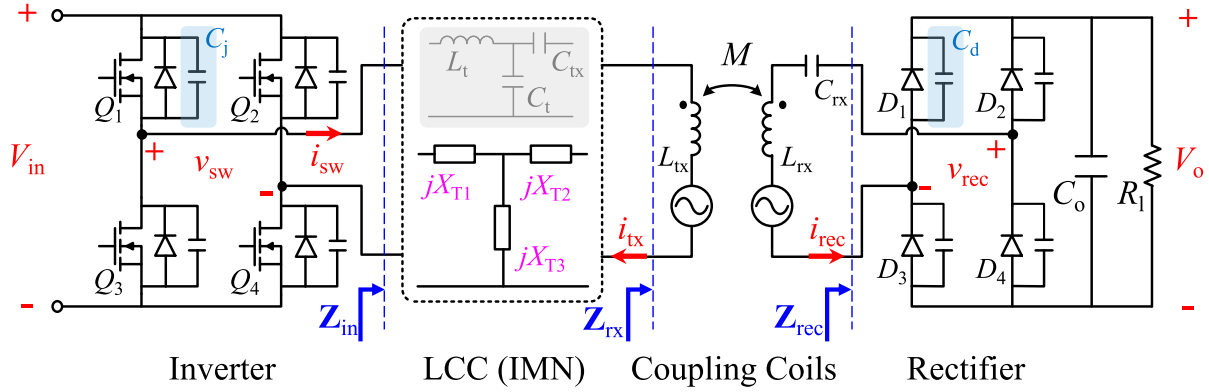


Fig. 1. Typical IPT system.

To overcome these limitations, this work develops an impedance-based modeling framework for bridge-type MHz IPT systems that explicitly incorporates device parasitics and switching transition effects. Based on this model, the wideband impedance characteristics of the inverter, resonant tank, and rectifier are analyzed, and a load-insensitive ZVS condition is derived through current decomposition. The resulting impedance-guided design methodology enables high-efficiency, wide-load-range operation, as validated by a 6.78-MHz, kilowatt-level IPT prototype.

## II. CHALLENGES OF MHz IPT

### A. System Configuration and Classical Design

Fig. 1 illustrates a typical IPT system based on full-bridge converters, which are widely employed in high-power applications. The system consists of an input voltage source  $V_{in}$ , an output voltage  $V_o$ , and two full-bridge converters. The inductive coupler is formed by a pair of coils,  $L_{tx}$  and  $L_{rx}$ , with mutual inductance  $M$ . To establish the resonant tank, an LCC-S compensation network is adopted: the transmitter side comprises  $L_t$ ,  $C_t$ , and  $C_{tx}$ , while the receiver side includes  $L_r$ ,  $C_r$ , and  $C_{rx}$ . In a kilohertz-level IPT system, the classical resonance condition under angular frequency  $\omega_0$  is expressed as

$$\begin{cases} \omega_0 L_t - \frac{1}{\omega_0 C_t} = 0, \\ \omega_0 L_{tx} - \frac{1}{\omega_0 C_{tx}} = \omega_0 L_t, \\ \omega_0 L_{rx} - \frac{1}{\omega_0 C_{rx}} = 0. \end{cases} \quad (1)$$

These resonance conditions are derived under two fundamental assumptions: 1) the switching transition time is negligible compared with the entire resonant period; and 2) the switching device parameters are insignificant and do not affect the resonant behavior.

Based on the perfect resonance defined in (1) and the above assumptions, a kilohertz IPT system can minimize the reactive power in the resonant tank and align the inverter voltage and current, thereby providing a theoretical basis for efficient power transfer. Meanwhile, such resonance inherently ensures a load-independent output voltage. To achieve ZVS, the

resonant tank can be designed to exhibit slightly inductive characteristics at the fundamental frequency, so that the additional circulating energy supports soft switching. This classical design method has been widely adopted in conventional IPT systems. However, this approach becomes ineffective when the operating frequency increases to the megahertz range—particularly around 6.78 MHz, as specified by the AirFuel standard. At such high frequencies, device parasitic capacitances and considerable switching transition durations inevitably distort the resonant behavior, invalidating the assumptions that hold true for kilohertz-level systems.

### B. Impact of Parasitic Effects

At megahertz frequencies, the influence of device parasitic capacitances and switching transitions becomes pronounced. These parasitics introduce additional reactive power and hinder the complete settling of the switch-node voltage during the dead time, making ZVS difficult to sustain and significantly increasing switching losses. Using classical design, Fig. 2 illustrates typical inverter waveforms under different transistor junction capacitances, i.e.,  $C_j$  in Fig. 1. Two major issues can be observed: (1) the transition interval occupies a substantial portion of the dead time, reducing ZVS robustness; and (2) the ZVS condition becomes highly sensitive to parasitic capacitance variations, resulting in efficiency fluctuation across operating points. These phenomena highlight the inherent tradeoff between ZVS robustness and system simplicity in MHz bridge inverters, revealing the necessity of reexamining impedance characteristics and design principles beyond conventional resonance-based approaches.

Building upon this motivation, this work proposes a nonperfect-resonance design framework in which parasitic effects and switching transitions are explicitly incorporated into impedance modeling and resonance design. This approach enables low-complexity, high-efficiency, and load-insensitive ZVS without auxiliary branches or dynamic dead-time tuning, thereby providing a practical pathway for MHz IPT systems.

### C. Impact of Load Variation

In MHz-range inverter circuits, load variation plays a critical role in determining whether soft switching can be effectively

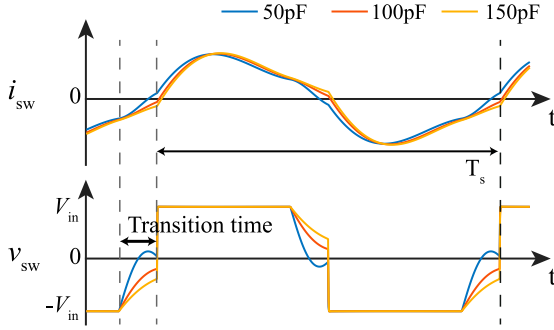
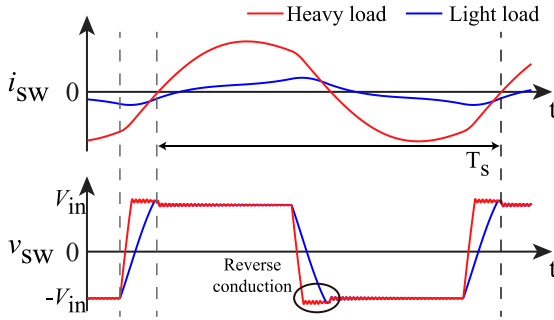
Fig. 2. Waveforms of MHz inverter output for different  $C_j$ .

Fig. 3. Waveforms of MHz inverter output for different loads.

maintained. Although parasitic capacitances of active devices can be included in circuit modeling to achieve ZVS under a specific load, such designs are typically optimized for a single operating point. Techniques such as series inductive compensation or auxiliary soft-switching branches can achieve ZVS under nominal load conditions, yet they fail to ensure robustness when the load deviates from its rated value. As the load changes, the imaginary part of the inverter load impedance varies accordingly, leading to the loss of soft switching. This not only increases switching losses but may even cause inverter malfunction at high frequencies.

In classical kilohertz-range systems, designers often enlarge design margins to guarantee soft switching under worst-case (light-load) conditions, as shown in Fig. 3. However, this approach results in over-ZVS behavior under light loads, where prolonged diode conduction during the dead time and excessive inductive current both deteriorate efficiency. For MHz operations, there remains a lack of systematic theoretical guidance for designing bridge inverters capable of maintaining ZVS across a wide load range. Simply introducing additional inductance to preserve soft switching is neither efficient nor practical. Therefore, it is essential to establish an accurate high-frequency system model and develop a dedicated wide-load-range soft-switching design methodology to meet the stringent requirements of MHz IPT systems.

### III. IMPEDANCE MODEL OF MHz IPT

#### A Input Impedance Model of Bridge Rectifier

In Fig. 1, parasitic effects such as the device junction capacitance  $C_d$  and  $C_j$  not only distort the resonant behavior

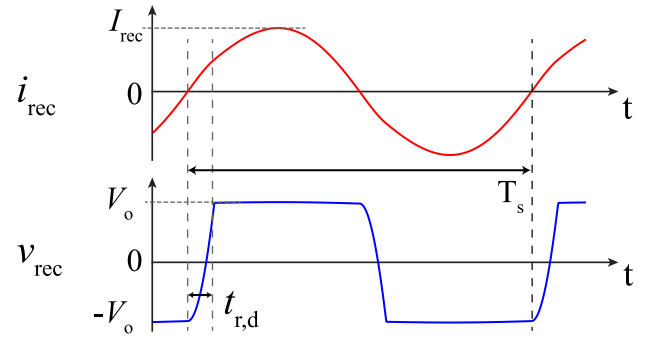


Fig. 4. Input voltage and current of rectifier.

but also modify the overall impedance characteristics, thereby introducing additional switching losses. From the transmitter (TX) perspective, the LCC compensation network can be equivalently represented as a T-type impedance matching network (IMN). To enable systematic parameter design, this article establishes an impedance model that incorporates both fundamental and harmonic components, allowing megahertz IPT systems to preserve performance characteristics comparable to their kilohertz-level counterparts.

The rectifier input impedance is derived from its input waveforms, as illustrated in Fig. 4. Owing to the junction capacitance  $C_d$ , the rectifier input voltage  $v_{rec}$  exhibits a quasi-trapezoidal shape rather than a pure sinusoid. As a result, the corresponding current  $i_{rec}$  contains residual harmonic components in addition to the fundamental. However, the RX resonant tank and the loosely coupled coils present a rapidly increasing impedance at higher harmonic orders, effectively filtering out most of these components. Consequently, the observed  $i_{rec}$  appears nearly sinusoidal across different load conditions, even though it is not strictly so in practice. This filtering effect justifies modeling the rectifier input impedance primarily at the fundamental frequency, while recognizing that the residual harmonics still contribute to switching transitions and are therefore explicitly incorporated in the proposed wideband impedance framework.

Based on the charging behavior of  $C_d$ , the rise time  $t_{r,d}$  is obtained as

$$2C_d V_o = \int_0^{t_{r,d}} i_{rec} dt = \frac{1}{\omega} I_{rec} (1 - \cos \omega_0 t_{r,d}),$$

$$t_{r,d} = \frac{1}{\omega_0} \cos^{-1} \left( 1 - \frac{2\omega_0 C_d V_o}{I_{rec}} \right). \quad (2)$$

At the fundamental frequency, the phasor form of the input voltage is expressed as

$$V_{rec} = \frac{8V_o}{\pi \omega_0 t_{r,d}} \sin \left( \frac{\omega_0 t_{r,d}}{2} \right), \quad \phi_{rec} = \frac{\omega_0 t_{r,d}}{2}. \quad (3)$$

Applying the power conservation principle yields the following:

$$\frac{V_o^2}{R_l} = \frac{1}{T_s} \int_0^{T_s} v_{rec}(t) i_{rec}(t) dt = \frac{2}{\pi} V_o I_{rec} - \frac{4}{T_s} C_d V_o^2 \quad (4)$$

which leads to

$$I_{rec} = \frac{\pi}{2} \left( \frac{1}{R_l} + \frac{4}{T_s} C_d \right) V_o. \quad (5)$$

Finally, the rectifier input impedance is given by

$$\mathbf{Z}_{\text{rec}} = \frac{\mathbf{V}_{\text{rec}}}{\mathbf{I}_{\text{rec}}} = \frac{16 \sin \frac{\omega_0 t_{r,d}}{2}}{\pi^2 \omega_0 t_{r,d} \left( \frac{1}{R_l} + \frac{4}{T_s} C_d \right)} \angle \phi_{\text{rec}}. \quad (6)$$

In a kHz system, it is able to ignore  $t_{r,d}$  and  $C_d$ . The corresponding  $\mathbf{Z}_{\text{rec}}$  becomes

$$\begin{aligned} \lim_{C_d, t_{r,d} \rightarrow 0} \mathbf{Z}_{\text{rec}} &= \lim_{C_d, t_{r,d} \rightarrow 0} \frac{16 \sin \frac{\omega_0 t_{r,d}}{2}}{\pi^2 \omega_0 t_{r,d} \left( \frac{1}{R_l} + \frac{4}{T_s} C_d \right)} \angle \phi_{\text{rec}} \\ &= \lim_{t_{r,d} \rightarrow 0} \frac{16 \sin \frac{\omega_0 t_{r,d}}{2}}{\pi^2 \omega_0 t_{r,d} \frac{1}{R_l}} = \frac{8R_l}{\pi^2} \end{aligned} \quad (7)$$

which is the well-known rectifier model at low-frequency system, it shows the model mathematically degenerates to the classical formulation when  $t_{r,d}$  and  $C_d$  approach zero, MHz design fundamentally differs. At 6.78 MHz, parasitic reactive energy dictates the ZVS boundary, requiring a strict shift from classical perfect-resonance assumptions to parasitic-aware impedance shaping.

### B. Load Impedance Model of Bridge Inverter

Similar to a bridge rectifier, when operating in the megahertz range, the load impedance of the bridge inverter must be accurately modeled to guide the design of the IMN. When  $C_{rx}$  resonates with  $L_{rx}$ , the rectifier impedance is reflected to the TX side as

$$\mathbf{Z}_{rx} = j\omega L_{tx} + \frac{\omega_0^2 M^2}{\mathbf{Z}_{\text{rec}}}. \quad (8)$$

The IMN of Fig. 1 has three reactances  $X_{T1}$ ,  $X_{T2}$ , and  $X_{T3}$  as the design variables, and  $\mathbf{Z}_{rx}$  is transformed into the fundamental input impedance

$$\mathbf{Z}_{in,f} = jX_{T1} + (jX_{T2} + \mathbf{Z}_{rx}) \parallel jX_{T3} \quad (9)$$

where  $X_{T1} = j\omega_0 L_t$ ,  $X_{T2} = 1/(j\omega_0 C_t)$ , and  $X_{T3} = 1/(j\omega_0 C_{tx})$ .

It should be emphasized that the modeling framework explicitly distinguishes between the fundamental and harmonic contributions. The rectifier input impedance is sufficiently represented at the fundamental frequency, whereas the IMN input requires both the fundamental term  $\mathbf{Z}_{in,f}$  and the harmonic term  $\mathbf{Z}_{in,h}$  to capture waveform distortion. Throughout the analysis, the subscripts  $f$  and  $h$  are consistently used to denote the fundamental and harmonic components, respectively, ensuring that the role of higher-order harmonics in ZVS evaluation is explicitly incorporated rather than neglected.

At MHz frequencies, the inverter output voltage  $v_{sw}$  contains substantial higher-order harmonic components. Due to the limited filtering capability of the IMN, these harmonics are not entirely suppressed, resulting in load-dependent distortion in the current waveform. Consequently, a design approach based solely on  $\mathbf{Z}_{in,f}$  is insufficient to guarantee ZVS. To capture this effect, a wideband impedance model is introduced for the IMN input port. Since the RX tank and coupler effectively filter harmonics, the TX-side induced voltage can be approximated as

a short circuit at harmonic frequencies, and the input impedance at harmonic order  $h$  is shown as follows:

$$\mathbf{Z}_{in,h} = j\omega_h L_t + \left( \frac{1}{j\omega_h C_{tx}} + j\omega_h L_{tx} \right) \parallel \frac{1}{j\omega_h C_t} \quad (10)$$

where  $\omega_h$  denotes the harmonic frequency. Unlike  $\mathbf{Z}_{in,f}$ ,  $\mathbf{Z}_{in,h}$  is independent of load and coupling, providing a stable contribution that is essential for wideband ZVS analysis.

### C. Influence of Harmonics and Assumptions of Proposed Model

In practical MHz IPT systems, the passive RX network, consisting of the resonant tank and loosely coupled coils, exhibits a strong low-pass characteristic. The impedance of the RX path increases approximately in proportion to the harmonic order, i.e.,  $|\mathbf{Z}_{rx}(h\omega_0)| \propto h$ , since both the inductive reactance and the resonant capacitive branch deviate sharply from resonance at higher frequencies. Consequently, the higher-order harmonic currents are effectively suppressed and do not significantly propagate back to the TX side. This filtering mechanism ensures that the rectifier input current  $i_{\text{rec}}$  remains quasi-sinusoidal and that the harmonic impedance terms beyond several times the fundamental contribute negligibly to power transfer and ZVS conditions. Therefore, the harmonic impedance of the passive RX network can be safely neglected in the analytical derivation, without loss of generality in predicting steady-state or switching behavior.

Because the harmonic components are strongly attenuated by the RX network, their influence on the TX-side voltage and current phasing is minimal. As a result, the IMN design and ZVS boundary can be primarily determined by the wideband input impedance at the TX port, where only the inverter-side harmonic impedance  $\mathbf{Z}_{in,h}$  needs to be considered. This assumption not only simplifies the analytical framework but also aligns with experimental observations that show negligible load dependence in harmonic distortion once the RX resonant tank is tuned to the operating frequency. Hence, the overall modeling accuracy and ZVS prediction remain valid even when higher-order harmonics of the RX path are omitted.

## IV. LOAD-INSENSITIVE DESIGN

### A. Design Constrains Using Impedance Model

The design objectives of a megahertz IPT system remain consistent with those of its kilohertz counterpart—reducing circulating energy, maintaining load-insensitive ZVS, and ensuring stable output current. However, unlike classical designs, these objectives become closely interdependent once harmonic effects are considered.

The desired ZVS waveforms are depicted in Fig. 5, where both the inverter output voltage  $v_{sw}$  (including a zoomed-in subfigure) and the corresponding current  $i_{sw}$  are illustrated under various load resistances  $R_l$ . Five representative load conditions are considered, ranging from  $50 \Omega$  to open-circuit (infinity). The lower bound of  $50 \Omega$  is selected because high-impedance operation at 6.78 MHz represents the most stringent

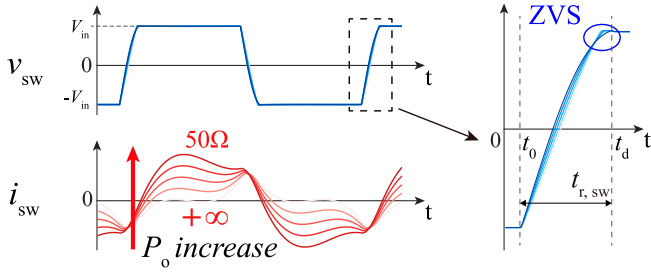


Fig. 5. Waveforms of input of IMN.

condition for achieving ZVS: the reduced current makes it more difficult to fully charge and discharge the device output capacitances within the available dead time. This range therefore provides a more rigorous validation of the proposed method. To ensure that ZVS can always be satisfied under different  $R_l$ , the following constraint must hold:

$$Q_{sw} = \int_{t_0}^{t_d} i_{sw}(t) dt \geq Q_j \quad (11)$$

where  $Q_j$  represents the required charge to completely discharge the switch junction capacitance. Here,  $t_0$  denotes the instant when the switch enters the dead-time interval, and  $t_d$  denotes the end of the dead time, i.e., the latest possible turn-ON instant of the complementary device. Physically, the integral of  $i_{sw}$  over this interval represents the charge available to discharge the device output capacitance. As long as  $Q_{sw}$  remains insensitive to  $R_l$ , ZVS operation can be guaranteed across the entire load range.

To evaluate  $Q_{sw}$ , the impedance model is employed to derive the expression for  $i_{sw}$ . For each frequency component, the current can be expressed as

$$\mathbf{I}_{sw} = \mathbf{V}_{sw} / \mathbf{Z}_{in} = (G_{in} + jB_{in}) \mathbf{V}_{sw} = \mathbf{Y}_{in} \mathbf{V}_{sw} \quad (12)$$

where  $\mathbf{V}_{sw}$  is obtained through the Fourier decomposition of the trapezoidal  $v_{sw}(t)$ . Here,  $G_{in}$ ,  $B_{in}$ , and  $\mathbf{Y}_{in}$  denote the input conductance, susceptance, and admittance, respectively. The inverter output current  $i_{sw}(t)$  can then be decomposed into three components

$$i_{sw}(t) = i_{f,\sin}(t) + i_{f,\cos}(t) + i_h(t) \quad (13)$$

where  $i_{f,\sin}(t)$ ,  $i_{f,\cos}(t)$ , and  $i_h(t)$  represent the resistive and inductive components of the fundamental current, and the overall harmonic current, respectively. Based on (12), these current components can be further expressed as

$$\begin{aligned} i_{f,\sin}(t) &= G_{in,f} |\mathbf{V}_{sw,f}| \sin(\omega_0 t), \\ i_{f,\cos}(t) &= B_{in,f} |\mathbf{V}_{sw,f}| \cos(\omega_0 t), \\ i_h(t) &= \sum_{n=2}^N |\mathbf{I}_{sw,n}| \sin(n\omega_0 t + \angle \mathbf{I}_{sw,n}) \end{aligned} \quad (14)$$

where  $G_{in,f}$ ,  $B_{in,f}$ , and  $\mathbf{V}_{sw,f}$  denote the fundamental components of the input conductance, susceptance, and voltage spectrum, respectively. The notation  $\mathbf{I}_{sw,n}$  refers to the  $n$ th-order harmonic component of the inverter current.

By substituting (13) into (11), the ZVS condition becomes

$$\int_{t_0}^{t_d} [i_{f,\sin}(t) + i_{f,\cos}(t) + i_h(t)] dt = Q_{f,\sin} + Q_{f,\cos} + Q_h \geq Q_j \quad (15)$$

where  $Q_{f,\sin}$ ,  $Q_{f,\cos}$ , and  $Q_h$  represent the integral of corresponding currents. Since  $i_{f,\sin}(t)$  is in phase with  $v_{sw}(t)$ , it does not contribute to reactive power, leading to  $Q_{f,\sin} = 0$ . This indicates that only  $Q_{f,\cos}$  and  $Q_h$  determine whether ZVS can be sustained.

It is worth noting that  $i_h(t)$  is inherently load-insensitive because  $\mathbf{Z}_{in,h}$  is independent of load and coupling [refer to (10)]. Hence,  $Q_h$  is a constant determined solely by the circuit parameters. In contrast, the load-dependent component  $i_{f,\cos}(t)$  directly influences ZVS, and its integral  $Q_{f,\cos}$  must remain approximately constant for different  $R_l$ . This requirement can be summarized as

$$B_{in,f}(R_l) = B^* \quad (16)$$

where  $B^*$  is the designed constant susceptance target. In summary, the IMN is designed in this article to simultaneously guarantee the load-insensitive contribution of  $Q_h$  and to stabilize  $Q_{f,\cos}$ , by satisfying the constraints defined in (15) and (16). While an LCC-S network is used herein as a representative T-type IMN, the constant-susceptance constraint in (16) is inherently topology-agnostic. This load-insensitive framework can be directly extended to other classical networks (e.g., S-S, LCC-LCC) provided they offer sufficient independent tuning reactances.

Although the primary design objective is to guarantee ZVS under load variations, output stability can still be maintained through the redundant degrees of design freedom in the IMN. The system gain, mainly governed by  $G_{in,f}$ , provides flexibility to balance ZVS and voltage regulation. The transmitted power is related to the input conductance as  $P_{in} = |\mathbf{V}_{sw,f}|^2 G_{in,f}$ . To ensure a load-insensitive  $V_o$ , the input power should scale proportionally with the load resistance. For instance, across a  $10\times$  load range (e.g., 50–500  $\Omega$ ), constraining  $G_{in,f}$  to vary over the same  $10\times$  range enables the system to maintain a load-insensitive  $V_o$ .

## B. Understanding the IMN Framework: A Case Study

In a practical MHz IPT design, ensuring robust power delivery and stable voltage gain takes precedence. Therefore, the overall parameter synthesis follows a sequential procedure: the base components, such as the transmitter-side series inductor and the coupling coils, are first determined to satisfy the maximum power capacity and desired voltage gain. Subsequently, the redundant degrees of freedom in the T-type IMN (i.e.,  $X_{T1}$ ,  $X_{T2}$ , and  $X_{T3}$ ) are fine-tuned to achieve ZVS without compromising the primary power delivery objectives.

The system parameters listed in Table I illustrate a specific study case and its corresponding design results. The fine-tuning process for achieving load-insensitive ZVS and  $V_o$  is visualized in Fig. 6. Based on the variation of  $R_l$ , the trajectories of  $\mathbf{Z}_{rec}$  and  $\mathbf{Z}_{rx}$  are determined using (7) and (8), respectively. The next step is to determine a target trajectory for  $\mathbf{Z}_{rx}$  to

TABLE I  
PARAMETERS OF THE VALIDATION SYSTEM

| Symbol            | Value              | Symbol          | Value              | Symbol   | Value                |
|-------------------|--------------------|-----------------|--------------------|----------|----------------------|
| $P_{\text{norm}}$ | 1000 W             | $V_{\text{in}}$ | 225 V              | $R_l$    | $50 \sim 500 \Omega$ |
| $f_o$             | 6.78 MHz           | $C_j$           | 150 pF             | $C_d$    | 50 pF                |
| $L_{tx}$          | 3.40 $\mu\text{H}$ | $L_{rx}$        | 3.48 $\mu\text{H}$ | $k$      | 0.287                |
| $X_{T1}$          | 46.0 $\Omega$      | $X_{T2}$        | -144.0 $\Omega$    | $X_{T3}$ | -42.7 $\Omega$       |
| $L_{T1}$          | 1.08 $\mu\text{H}$ | $C_{T2}$        | 163 pF             | $C_{T3}$ | 550 pF               |

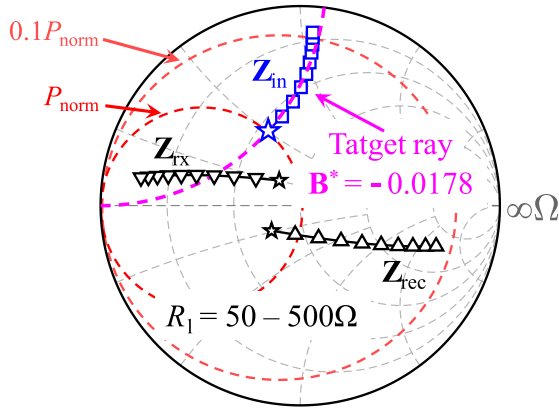


Fig. 6. Impedance matching for load-insensitivity ZVS and  $V_o$ .

match. Specifically,  $Q_h$  is first calculated based on the baseline parameters, which then defines the required  $Q_{f,\text{cos}}$  and subsequently the constant target  $B^*$ . It should be emphasized that the inequality constraint in (15) can in practice be treated as an equality, which not only simplifies the design but also minimizes redundant circulating energy. The resulting constant  $B^*$  defines the desired trajectory, as plotted in Fig. 6.

Two constant power contours,  $P_{\text{norm}}$  and  $0.1P_{\text{norm}}$ , are included in Fig. 6 to illustrate the  $10\times$  power variation. As a result, the variation in  $R_l$  should correspond to a change in  $Z_{\text{in},f}$  that lies between these two contours. The objective of impedance matching, achieved by tuning  $X_{T1}$  to  $X_{T3}$ , is therefore to fulfill the following criteria: 1)  $Z_{\text{in},f}$  lies on the  $P_{\text{norm}}$  contour for the minimum  $R_l$ ; 2)  $Z_{\text{in},f}$  remains close to the constant- $B^*$  curve; 3)  $Z_{\text{in},f}$  lies on the  $0.1P_{\text{norm}}$  contour for the maximum  $R_l$ . The resulting matched  $Z_{\text{in},f}$  is shown in Fig. 6. It should be noted that satisfying all three objectives perfectly is generally not feasible in practice. Therefore, small deviations from the second and third criteria are acceptable, which may slightly introduce load sensitivity into the ZVS condition and output voltage but remain within tolerable limits.

In summary, the design of a MHz-range system should first determine the expected load range, which serves as the basis for parameter optimization. This process helps define two constant-power contours, represented by the red circles in Fig. 6. Based on the proposed impedance model, and by accounting for

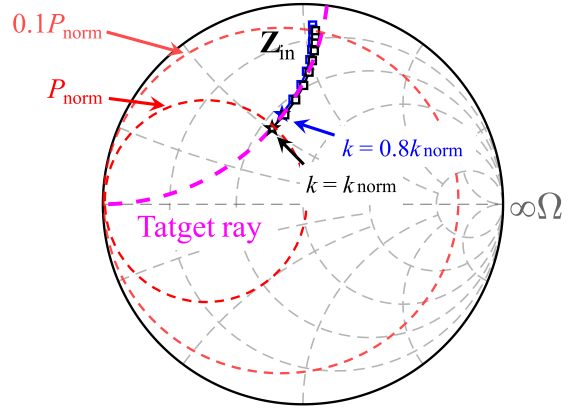


Fig. 7. Trajectory of  $Z_{\text{in}}$  under coupling variation.

both junction capacitance and dead time, the target susceptance  $B^*$  can be derived to guide the tuning of IMN for load-insensitive ZVS. Finally, the IMN can be tuned such that the input impedance  $Z_{\text{in},f}$  aligns with the pink trajectory within the region bounded by the red circles.

### C. Impact of Coupling Coefficient Variation

The classical design in (1) is inherently independent of the coupling coefficient. In contrast, the proposed model and design method are coupling-dependent; however, achieving the design objectives does not require satisfying strict equality constraints. This provides greater design flexibility to accommodate coupling variations. Similar to the kHz LCC-S topology, the output voltage of a MHz system is also proportional to the coupling coefficient. Since  $Q_{f,\text{cos}}$  varies with coupling, maintaining ZVS over a wide coupling range requires that (15) be satisfied within the desired coupling interval. This requirement can be fulfilled by selecting a more conservative value of  $B^*$  during the IMN design stage.

To ensure reliable ZVS across the entire coupling range, the design of the target susceptance  $B^*$  [i.e., meeting (16)] must include sufficient margin. Specifically, rather than selecting  $B^*$  solely based on the nominal coupling coefficient  $k$ , a conservative value is chosen such that condition (11) remains satisfied even under the weakest coupling condition (i.e.,  $0.8k_{\text{norm}}$ ). In Fig. 6, although the trajectory of  $Z_{\text{in}}$  is plotted under the maximum coupling condition, the IMN design summarized in Table I has already accounted for the potential reduction in coupling. When  $k$  decreases by 20%, another trajectory can be obtained, as illustrated in Fig. 7. The new trajectory still satisfies the design constraints on the Smith chart, remaining aligned with the target pink curve and confined within the two constant-power contours.

While the conservative margin safeguards against weak coupling, it is equally critical to define the upper operational boundaries. Analytically, the load-insensitive ZVS breaks down if the system is severely over-coupled or if  $R_l$  drops excessively (thereby crossing the  $P_{\text{norm}}$  contour). Under these extreme conditions, the fundamental current amplitude rises sharply, forcing a premature polarity reversal before the complementary switch

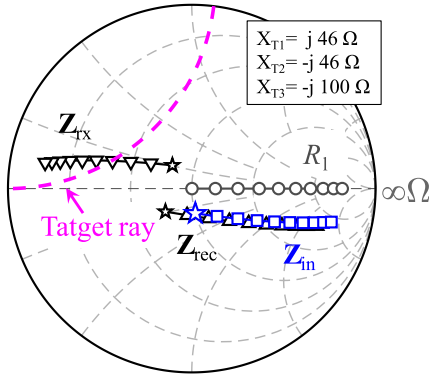


Fig. 8. Impedance trajectory of classical design using different design.

turns ON (prior to  $t_d$ ). This triggers body diode conduction and violates the required charge constraints. Ultimately, ZVS is reliably maintained as long as the concurrent variations of load and coupling confine the fundamental input impedance  $Z_{in,f}$  within the designed constant-power contours.

#### D. Comparison With Classical Design

These results confirm that the proposed impedance-based design framework not only achieves load-insensitive behavior but also demonstrates robustness against practical coupling coefficient variations, which is essential for high-frequency IPT systems.

The proposed impedance model can be significantly simplified at the kHz range, where the resulting IMN design achieves similar objectives to the classical approach. However, due to modeling inaccuracies, the classical design fails to maintain performance at the MHz range. Specifically, Fig. 8 compares the impedance trajectories of a MHz system designed using the proposed method and the classical method. First of all, both design cases operate under the same load range, leading to identical  $R_l$  and  $Z_{rec}$  trajectories. Since both employ the same RX design, the trajectories of  $Z_{rx}$  are also identical. The key difference lies in the IMN design, which results in a pronounced discrepancy in the  $Z_{in}$  trajectories. The classical design notably deviates from the target pink ray, which is essential for ensuring load-insensitive ZVS.

#### V. EXPERIMENT VERIFICATION

Fig. 9 illustrates the constructed 1-kW 6.78-MHz IPT prototype, which consists of a full-bridge GaN inverter, an LCC-S compensation network, and a diode bridge rectifier. The system is powered by a regulated DC source and loaded by a programmable electronic load. All key signals are monitored using a high-bandwidth oscilloscope equipped with a differential voltage probe and a wide-band current probe, while input and output powers are measured with a precision power analyzer to guarantee accurate efficiency evaluation under MHz operation.

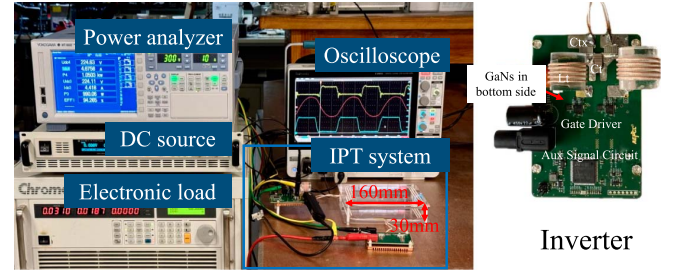


Fig. 9. Experiment setup of the 6.78-MHz 1-kW IPT prototype.

The transmitter and receiver coils are wound with litz wire (0.03 mm  $\times$  1000 strands), and the coil quality factor is measured to be close to 300 at 6.78 MHz using an impedance analyzer. The physical dimensions of the coils are 160 mm  $\times$  30 mm, corresponding to a consistent coupling coefficient of 0.287. Due to the limited core volume and winding turns, the compensation inductor exhibits a relatively lower quality factor of approximately 130, which reflects a common limitation in compact MHz IPT hardware. The inverter stage employs GaN HEMTs (GS66508T), chosen for their low output capacitance and fast switching speed, which are essential for minimizing transition loss at MHz operation. Although the GaN output capacitance ( $C_{oss}$ ) is inherently nonlinear, the switching transition predominantly occurs above 200 V, where capacitance variation is well bounded ( $\pm 10\%$ ). Since (11) evaluates the total required charge, utilizing a linear equivalent capacitance introduces negligible error to the ZVS boundary. The rectifier is realized with SiC diodes (C3D04065A), which provide low junction capacitance and acceptable reverse recovery charge. The circuit parameters used in the prototype are consistent with those listed in Table I.

Fig. 10(a)–(c) display the measured steady-state waveforms under 500  $\Omega$ , 100  $\Omega$ , and 50  $\Omega$  loads, corresponding to light, half, and full load conditions, respectively. In each case, the gate drive signal  $v_{gs}$ , the drain-source voltage  $v_{ds}$ , and the inverter current  $i_{sw}$  are recorded. The results verify that the proposed IMN design successfully enables ZVS across a wide load range. To further examine the switching behavior, Fig. 11 enlarges the device transition interval. It is evident that  $v_{ds}$  falls to zero before turn-ON, confirming nearly ideal ZVS operation. In addition, the negligible reverse conduction of the GaN devices, compared with conventional Si MOSFETs, ensures that redundant circulating energy is minimized, effectively validating that (15) can be treated as an equality constraint in practice.

To further assess the resilience of the proposed design under dynamic operating conditions, Fig. 12 shows the measured switching behavior during a rapid load step from 500  $\Omega$  to 50  $\Omega$ . Despite the abrupt change in resonant current, the drain-source voltage  $v_{ds}$  is still observed to fall to zero prior to device turn-ON, indicating that ZVS is continuously preserved throughout the transient interval. Although the transient waveform does not allow clear observation of cycle-by-cycle switching details, the steady-state results in Fig. 8(a)–(c) demonstrate the same consistent soft-switching behavior across light, medium, and

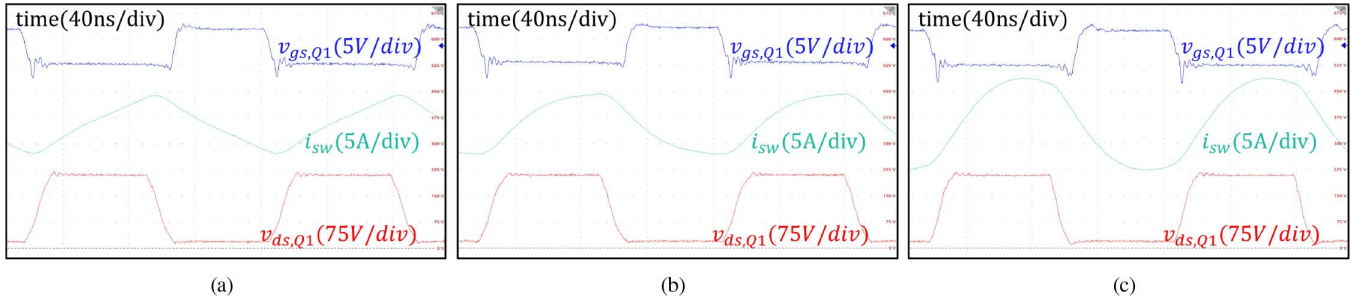


Fig. 10. Steady-state waveforms. (a)  $R_L = 500 \Omega$ . (b)  $R_L = 100 \Omega$ . (c)  $R_L = 50 \Omega$ .

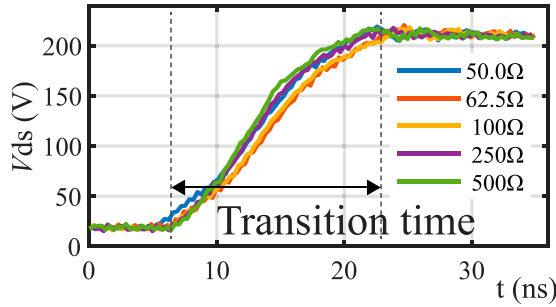


Fig. 11. Measured  $v_{ds}$  during switching transition under different load conditions.

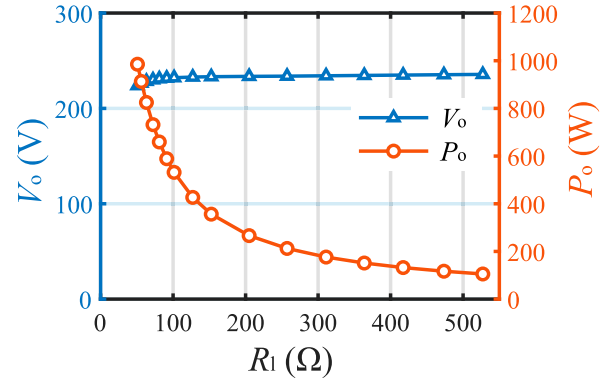


Fig. 13. Measured DC output voltage and power versus load.

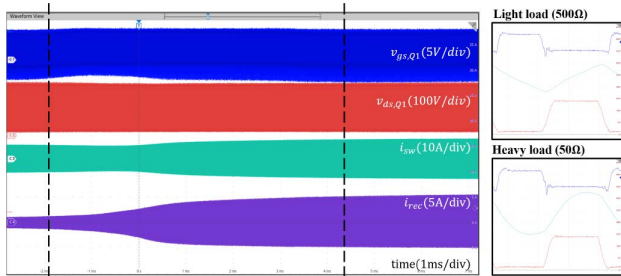


Fig. 12. Dynamic waveform under a load step from light load ( $500 \Omega$ ) to heavy load ( $50 \Omega$ ).

heavy load conditions. Together, these measurements confirm that the proposed impedance matching and control method maintains robust soft-switching performance under both steady-state and dynamic load variations.

The measured DC output voltage  $V_o$  and delivered power  $P_o$  under different loads are summarized in Fig. 13. It is observed that  $V_o$  remains nearly constant across a tenfold load variation ( $50$ – $500 \Omega$ ), demonstrating the load-insensitive regulation capability of the proposed design. A concise loss analysis is performed for the inverter, resonant tank, and rectifier, loss breakdown of the system under full-load condition is shown in Fig. 14. Tank losses are estimated from the measured coil Q-factor and current waveforms, while rectifier losses are obtained from conduction-current evaluation. The overall DC–DC efficiency is derived from the measured input–output power.

The inverter loss, as the main focus of this work, is decomposed into conduction, turn-ON, turn-OFF, and body-diode components. A representative full-load pie chart is provided.

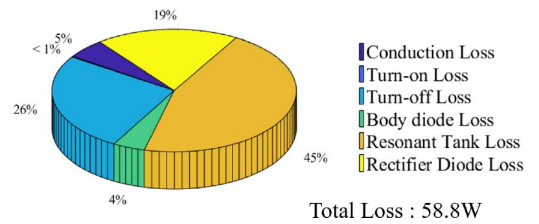


Fig. 14. Loss breakdown of the system under full-load condition.

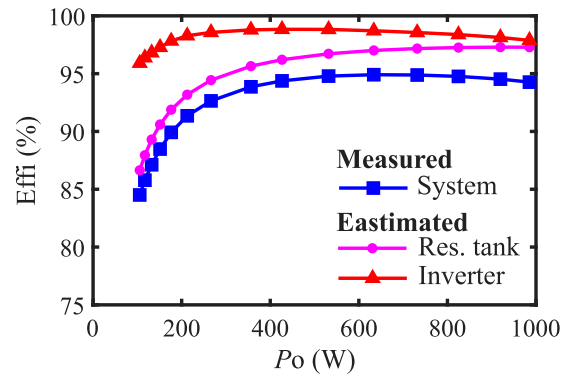


Fig. 15. Measured system efficiency versus output power.

Conduction loss accounts for only about 5% of the total system loss ( $\approx 15\%$  of inverter loss), and the achieved ZVS renders turn-ON loss negligible. Turn-OFF loss remains the major contributor at 6.78 MHz. Furthermore, the 137-ns switching period

TABLE II  
COMPARISON WITH REPRESENTATIVE MHz IPT SYSTEMS

| Ref.      | $f_s$     | Topology       | $V_{in}$ | Comp. Type    | ZVS Method           | $P_o$  | DC-DC Effi.    | Load Range      |
|-----------|-----------|----------------|----------|---------------|----------------------|--------|----------------|-----------------|
| [13]      | 27.12 MHz | Class E        | 65 V     | S/S           | transformed $\Phi^2$ | 25 W   | [56.0%, 90.0%] | 25~250 $\Omega$ |
| [16]      | 10 MHz    | Class E        | 200 V    | S/S           | NA                   | 150 W  | [83.5%, 90.0%] | 50~150 $\Omega$ |
| [14]      | 6.78 MHz  | Class $\Phi^2$ | 110 V    | LCC/S         | NA                   | 320 W  | [72.1%, 91.0%] | 7.2~72 $\Omega$ |
| [20]      | 1 MHz     | Full-bridge    | 40 V     | S/S           | aux. ZVS Branch      | 10 W   | [73.0%, 93.0%] | 25~100 $\Omega$ |
| [19]      | 6.78 MHz  | Full-bridge    | 200 V    | LCC/S         | aux. ZVS branch      | 100 W  | [-, 92.8%]     | 400 $\Omega$    |
| [9]       | 6.78 MHz  | Full-bridge    | 525 V    | Self-resonant | NA                   | 1000 W | [80.0%, 95.3%] | 50 $\Omega$     |
| This work | 6.78 MHz  | Full-bridge    | 225 V    | LCC/S         | proposed             | 1000 W | [84.5%, 94.9%] | 50~500 $\Omega$ |

makes the body-diode loss highly sensitive to dead time: adding only 5 ns increases its share from 4% to 9%. This highlights the importance of achieving load-insensitive ZVS to avoid frequent dead-time retuning and unnecessary diode conduction.

The resulting efficiency is shown in Fig. 15. The prototype achieves a peak DC–DC efficiency of 94.9% at around 700 W, while inverter efficiency reaches 98.8% and remains above 95.5% across the operating range, confirming the effectiveness of the proposed impedance-matching method.

Finally, Table II benchmarks the proposed prototype against representative MHz IPT systems from recent literature. It can be observed that the proposed design simultaneously achieves kilowatt-level power transfer, a wide load range (50–500  $\Omega$ ), and excellent soft-switching capability. This combination of high efficiency, scalability, and robustness distinguishes the proposed system from existing state-of-the-art MHz IPT solutions and highlights its potential for practical deployment.

## VI. CONCLUSION

This article has proposed a systematic design methodology for impedance matching networks to enable load-insensitive operation in MHz IPT systems using a bridge topology. By establishing a wide-band impedance model that incorporates junction capacitances and switching transition times, the conditions for maintaining ZVS were analytically derived. This model provides a clear guideline for parameter selection, ensuring that both ZVS and output voltage regulation can be achieved over a wide load range. A 1-kW, 6.78-MHz prototype was built to validate the proposed method. The experimental results confirmed nearly ideal ZVS transitions, stable output voltage across a tenfold load variation, and high efficiency with a peak system efficiency of 94.9% and a peak inverter efficiency of 98.8%. Compared with existing MHz IPT designs, the proposed approach demonstrates the capability to simultaneously achieve kilowatt-level power transfer, wide load adaptability, and excellent efficiency, highlighting its significance for practical high-frequency wireless power applications.

## REFERENCES

- [1] J. Xu et al., "Automatic time-division multiplexing for inductive power transfer to multiple stainless-steel-enclosed receivers," *IEEE J. Emerg. Sel. Topics Power Electron.*, vol. 12, no. 2, pp. 2333–2344, Apr. 2024.
- [2] E. S. Lee, "Frequency-modulation-based IPT with magnetic communication for EV wireless charging," *IEEE Trans. Ind. Electron.*, vol. 70, no. 2, pp. 1398–1408, Feb. 2023.
- [3] I.-W. Iam, C.-K. Choi, C.-S. Lam, P.-I. Mak, and R. P. Martins, "A constant-power and optimal-transfer-efficiency wireless inductive power transfer converter for battery charger," *IEEE Trans. Ind. Electron.*, vol. 71, no. 1, pp. 450–461, Jan. 2024.
- [4] L. Wang, P. Sun, Y. Liang, L. He, X. Wu, and Q. Deng, "Research on the control strategy of communication-free IPT system based on multi-parameter joint real-time identification," *IEEE Trans. Power Electron.*, vol. 39, no. 1, pp. 1912–1926, Jan. 2024.
- [5] S. Zhao, F. Chen, C. Tang, P. Deng, and C. Duan, "Modeling and control of WPT systems in the presence of load and mutual inductance variations," *IEEE Trans. Power Electron.*, vol. 39, no. 11, pp. 15315–15328, Nov. 2024.
- [6] Y. Wang, Z. Sun, Y. Guan, and D. Xu, "Overview of megahertz wireless power transfer," *Proc. IEEE*, vol. 111, no. 5, pp. 528–554, May 2023.
- [7] R. Qin, J. Li, and D. Costinett, "A 6.6-kW high-frequency wireless power transfer system for electric vehicle charging using multilayer nonuniform self-resonant coil at MHz," *IEEE Trans. Power Electron.*, vol. 37, no. 4, pp. 4842–4856, Apr. 2022.
- [8] Y. Wang, S. Zhao, F. Lu, and H. Zhang, "Challenges in zero-voltage-switched-on multi-MHz multi-kW SiC full-bridge inverter and oriented design for high-power capacitive power transfer system," *IEEE J. Emerg. Sel. Topics Power Electron.*, vol. 13, no. 4, pp. 4256–4277, Aug. 2025.
- [9] L. Gu, G. Zulauf, A. Stein, P. A. Kyaw, T. Chen, and J. M. R. Davila, "6.78-MHz wireless power transfer with self-resonant coils at 95% DC–DC efficiency," *IEEE Trans. Power Electron.*, vol. 36, no. 3, pp. 2456–2460, Mar. 2021.
- [10] B. Regensburger, S. Sinha, A. Kumar, S. Maji, and K. K. Afridi, "High-performance multi-MHz capacitive wireless power transfer system for EV charging utilizing interleaved-foil coupled inductors," *IEEE J. Emerg. Sel. Topics Power Electron.*, vol. 10, no. 1, pp. 35–51, Feb. 2022.
- [11] J. M. Arteaga, S. Aldaher, G. Kkelis, C. Kwan, D. C. Yates, and P. D. Mitcheson, "Dynamic capabilities of multi-MHz inductive power transfer systems demonstrated with batteryless drones," *IEEE Trans. Power Electron.*, vol. 34, no. 6, pp. 5093–5104, Jun. 2019.
- [12] J. Xu, Z. Tong, and J. Rivas-Davila, "1 kW MHz wideband Class E power amplifier," *IEEE Open J. Power Electron.*, vol. 3, pp. 84–92, 2022.
- [13] L. Roslaniec, A. S. Jurkov, A. A. Bastami, and D. J. Perreault, "Design of single-switch inverters for variable resistance/load modulation operation," *IEEE Trans. Power Electron.*, vol. 30, no. 6, pp. 3200–3214, Jun. 2015.
- [14] X. Huang, Y. Lin, Y. Dou, S. Lin, and J. Huang, "Load-independent push-pull Class- $\Phi^2$  inverter with single compact three-winding inductor," *IEEE Trans. Power Electron.*, vol. 38, no. 10, pp. 11916–11927, Oct. 2023.
- [15] S. Aldaher, D. C. Yates, and P. D. Mitcheson, "Modeling and analysis of Class EF and Class E/F inverters with series-tuned resonant networks," *IEEE Trans. Power Electron.*, vol. 31, no. 5, pp. 3415–3430, May 2016.
- [16] S. Aldaher, D. C. Yates, and P. D. Mitcheson, "Load-independent Class E/EF inverters and rectifiers for MHz-switching applications," *IEEE Trans. Power Electron.*, vol. 33, no. 10, pp. 8270–8287, Oct. 2018.
- [17] Y. Shao, H. Zhang, M. Liu, and C. Ma, "Explicit design of impedance matching networks for robust MHz WPT systems with different features," *IEEE Trans. Power Electron.*, vol. 37, no. 9, pp. 11382–11393, Sep. 2022.

- [18] E. Rong, P. Sun, G. Yang, J. Xia, Z. Liu, and S. Li, "5-kW, 96.5% efficiency capacitive power transfer system with a five-plate coupler: Design and optimization," *IEEE Trans. Power Electron.*, vol. 40, no. 1, pp. 2542–2555, Jan. 2025.
- [19] J. Li, R. Qin, J. Sun, and D. Costinett, "Systematic design of a 100-W 6.78-MHz wireless charging station covering multiple devices and a large charging area," *IEEE Trans. Power Electron.*, vol. 37, no. 4, pp. 4877–4889, Apr. 2022.
- [20] H. Li, K. Wang, J. Fang, and Y. Tang, "Pulse density modulated ZVS full-bridge converters for wireless power transfer systems," *IEEE Trans. Power Electron.*, vol. 34, no. 1, pp. 369–377, Jan. 2019.
- [21] H. Tebianian, Y. Salami, B. Jeyasurya, and J. E. Quaicoe, "A 13.56-MHz full-bridge Class-D ZVS inverter with dynamic dead-time control for wireless power transfer systems," *IEEE Trans. Ind. Electron.*, vol. 67, no. 2, pp. 1487–1497, Feb. 2020.



**Xinlin Wang** (Graduate Student Member, IEEE) received the B.S. degree in electrical engineering and automation from Hohai University, Nanjing, China, in 2020. He is currently working toward the Ph.D. degree in power electronics with the School of Information Science and Technology, ShanghaiTech University, Shanghai, China.

His research interests include high frequency resonant converter, multiple-coil, and bidirectional WPT systems.



**Jun Yang** received the B.S. degree in electronic and information engineering in 2025 from ShanghaiTech University, Shanghai, China, where he is currently working toward the Ph.D. degree in power electronics with the School of Information Science and Technology.

His research interests include megahertz wireless power transfer, high-frequency power amplifiers, and medical power supply.



**Yu Liu** (Senior Member, IEEE) received the B.S. and M.S. degrees in electrical power engineering from Shanghai Jiao Tong University, Shanghai, China, in 2011 and 2013, respectively, and the Ph.D. degree in electrical and computer engineering from Georgia Institute of Technology, Atlanta, GA, USA, in 2017.

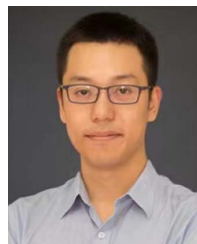
He is currently a Tenured Associate Professor with the School of Information Science and Technology, ShanghaiTech University, Shanghai.

His research interests include modeling, protection, fault location, and state/parameter estimation of power systems and power electronic systems.



**Junrui Liang** (Senior Member, IEEE) received the B.E. and M.E. degrees in instrumentation engineering from Shanghai Jiao Tong University, Shanghai, China, in 2004 and 2007, respectively, and the Ph.D. degree in mechanical and automation engineering from The Chinese University Hong Kong, Hong Kong, China, in 2010.

He is currently an Associate Professor with the School of Information Science and Technology, ShanghaiTech University, Shanghai. His research interests include energy conversion and power conditioning circuits, kinetic energy harvesting and vibration suppression, IoT devices, and mechatronics.



**Haoyu Wang** (Senior Member, IEEE) received the bachelor's (Distinguished Hons.) degree from Zhejiang University, Hangzhou, China, in 2009, and the Ph.D. degree from the University of Maryland, College Park, MD, USA, in 2014, both in electrical engineering.

In 2014, he joined the School of Information Science and Technology, ShanghaiTech University, Shanghai, China, where he is currently a Full Professor with tenure. In 2023, he was a Visiting Academic Fellow with the University of Cambridge, Cambridge, U.K. His research interests include power electronics, electric vehicles, renewable energy systems, and power management integrated circuits.

Dr. Wang is an IET Fellow. He is as an Associate Editor of IEEE TRANSACTIONS ON INDUSTRIAL ELECTRONICS, IEEE TRANSACTIONS ON TRANSPORTATION ELECTRIFICATION, and *CPSS Transactions on Power Electronics and Applications*. He was the Guest Editor of IEEE JOURNAL OF EMERGING AND SELECTED TOPICS IN POWER ELECTRONICS and a Guest Associate Editor for IEEE TRANSACTIONS ON POWER ELECTRONICS.



**Minfan Fu** (Senior Member, IEEE) received the B.S., M.S., and Ph.D. degrees in electrical and computer engineering from the University of Michigan-Shanghai Jiao Tong University Joint Institute, Shanghai Jiao Tong University, Shanghai, China, in 2010, 2013, and 2016, respectively.

He is currently an Associate Professor with the School of Information Science and Technology, ShanghaiTech University, Shanghai. From 2016 to 2018, he held a Postdoctoral Researcher position with the Center for Power Electronics Systems (CPES), Virginia Polytechnic Institute and State University, Blacksburg, VA, USA. His research interests include megahertz wireless power transfer, high-frequency power conversion, high-frequency magnetic design, and the application of widebandgap devices. He holds one U.S. patent and seven Chinese patents, and has authored or coauthored more than 130 papers in prestigious IEEE journals and conferences.