Abstract—The soft-switching gallium-nitride (GaN) based critical conduction mode (CRM) totem-pole power factor correction (PFC) converter is a good candidate for the front-end rectifier in wireless power transfer (WPT) applications. In multi-receiver MHz WPT systems, the PFC converter is required to have fast dynamic response and noise immunity. In this work, a GaN-based CRM totem-pole PFC converter is designed for a multi-receiver wireless charging power supply. Digital-based variable on-time control is used to achieve the zero voltage switching (ZVS) within the whole line cycle, and a voltage-loop controller with notch filters is designed to improve the transient response. The impact of high-frequency noise on the sensing signals and ZVS control are analyzed, and implementation methods are proposed to mitigate the disturbances. A GaN-based CRM totem-pole PFC that is demonstrated with 98.5% full-load efficiency is built as the first stage in the transmitter of a 100 W 6.78 MHz multi-receiver WPT system. The noise immunity of the CRM PFC is verified by testing the whole WPT system. Experimental results show that the system end-to-end efficiency at full load is 90.16%, and fast dynamic response is achieved during load variation.

Index Terms—GaN, CRM, totem-pole PFC, ZVS, WPT, fast dynamic response, noise immunity

I. INTRODUCTION

Wireless power transfer (WPT) based on MHz magnetic resonant coupling has gained increasing attention in applications like consumer electronics and electric vehicles, due to the advantages of spatial freedom, long transfer distance, and high efficiency [1], [2]. In order to maximize the end-to-end efficiency and power density, research efforts are made on transmission coil optimization [3], and advanced power conversion techniques with two-stage or single-stage transmitters [4] and receivers [5]. To also satisfy the power factor and harmonic requirements, a high-efficiency power factor correction (PFC) converter is required.

Recently, the gallium-nitride (GaN) based totem-pole PFC converter has been studied in front-end rectifiers, because of the simple topology, low conduction loss, and high utilization of devices and components [6], [7]. With the critical conduction mode (CRM) operation, zero voltage switching (ZVS) or valley switching can be achieved [8]. To further reduce the switching loss for achieving higher efficiency and power density, ZVS control strategies are developed to realize soft switching within the whole line cycle [9]–[11]. The soft-switching CRM totem-pole PFC has been applied on MHz WPT systems and reported with > 98% high efficiency [12].

However, MHz wireless charging brings in new requirements and challenges for the GaN-based CRM totem-pole PFC design. On the one hand, for WPT systems, especially the ones allowing multiple devices to charge simultaneously, fast dynamic response is indispensable. This requirement can be realized by regulating the inverter with constant output current behavior [12], but this technique relies on a constant DC bus voltage at the input of the inverter. Also, the voltage-loop control bandwidth of the single-phase rectifier is usually designed well below the line frequency (typically < 20 Hz), which is limited by the twice-line frequency ripple on the DC bus. To improve the dynamic response during load changing, higher-bandwidth control schemes should be investigated [13], [14], and adopted in the CRM totem-pole PFC converter.

On the other hand, ZVS control of the CRM PFC requires instantaneous voltage sensing and high-frequency current sensing. In the MHz switching system, high dv/dt noise is easily coupled into the sensing signals electrically and magnetically [15], resulting in erroneous switching actions and fragile ZVS control [16]. The noise impact is even more severe in multi-receiver WPT systems with MHz magnetic resonant coupling and large coils. Parasitic capacitance exists between the transmitter coil and receiver coils, also between the WPT coils and the ground, which transfers the noise current from the primary side to the load side, and from the coil to the ground. Involved in the resonant tank, the noise current has 6.78 MHz resonant ripple, and can be propagated to the PFC stage and disrupt the sensing signals. Therefore, a front-end totem-pole PFC with noise immunity is required to maintain reliable ZVS control.

In this paper, a GaN-based CRM totem-pole PFC converter with fast dynamic response and low sensitivity to noise is developed for a 100 W 6.78 MHz multi-receiver wireless charging power supply in consumer electronics. Fig. 1(a)
presents the example application of the WPT power supply, where computer mouse, cellphone, display, and laptop are powered simultaneously on one transmission desk. Fig. 1(b) shows the circuit topology of the WPT system, which includes a two-stage transmitter consisting of a CRM totem-pole PFC and a 6.78 MHz half-bridge inverter, WPT coils, and receiverside diode rectifiers.

The paper is organized as follows. Section II introduces the operation principle and ZVS control of the GaN-based CRM totem-pole PFC. Section III provides the digital-based variable on-time control with high bandwidth and notch filters for enhancing the transient response. In section IV, hardware implementation is introduced, high-frequency noise impacts on the sensing signals are analyzed, and solutions to mitigate the noise disturbances are proposed. Section V shows the experimental results, and section VI gives the conclusions.

II. OPERATION PRINCIPLE OF THE CRM TOTEM-POLE PFC WITH FULL-LINE-CYCLE ZVS

As shown in Fig. 1(b), for the totem-pole PFC, GaN devices $S_1$ and $S_2$ operate at high frequency, and $S_3$ and $S_4$ work at AC line frequency. During the positive half line cycle, $S_4$ is always on, $S_3$ is always off, $S_2$ is the active switch (AS), and $S_1$ is the synchronous switch (SS). During the negative half line cycle, $S_3$ is always on, $S_4$ is always off, $S_1$ is the active switch, and $S_2$ is the synchronous switch.

The operation principle of the CRM totem-pole PFC is illustrated with the theoretical operation waveforms shown in Fig. 2 and the state-plane trajectories shown in Fig. 3. Discussion here is based on the operation within the positive half line cycle, with the assumption that $V_{in}$ is nearly-constant in one switching cycle. For the state-plane trajectories, $V_{bound}$ is the boundary input voltage between the natural ZVS region and the non-natural ZVS region. The characteristic impedance $Z_n$ is defined as $Z_n^2 = L_0/(2C_{oss})$, where $C_{oss}$ is the equivalent drain-to-source capacitance of $S_1$ and $S_2$, assuming $C_{oss,S1} = C_{oss,S2} = C_{oss}$. The resonance angular frequency is designated as $\omega_r^2 = 1/(2C_{oss}L_0)$.

As illustrated in Fig. 2, to achieve ZVS within the full line cycle, conduction time of the synchronous switch $S_1$ is purposely extended for $T_{ex,SS}$. In the natural ZVS region when $V_{in} \leq V_{bound}$, $T_{ex,SS}$ is not needed, and $S_1$ is turned off at the positive-to-negative zero current crossing point, $t_0$. When $V_{in} > V_{bound}$, $S_1$ conduction time is prolonged until $t_1$ to increase the energy stored in the inductor, so that the active switch $S_2$ drain-to-source voltage $V_{ds,S2}$ is able to resonate to zero before turning on $S_2$, and ZVS is maintained.

Based on the state plane in Fig. 3, ZVS operation is achieved as long as $r_1$ is larger than $V_{in}$. Hence, the adaptive ZVS margin is defined as

$$k = \frac{r_1}{V_{in}} = \begin{cases} \frac{V_{in} - V_{in}}{k_0}, & V_{in} \leq V_{bound} \\ \frac{V_{in}}{k_0}, & V_{in} > V_{bound} \end{cases}$$

where $V_{bound} = V_r/(k_0 + 1)$, and $k_0$ is the selected ZVS margin which is slightly larger than 1. The value of $k_0$ cannot be too large, because large ZVS margin means more negative inductor
current, hence larger RMS current and conduction loss. Accordingly, the extended conduction time of the synchronous switch is

$$T_{ex\_SS} = \frac{\sqrt{(k^2 - 1)V_{in}^2 - V_o^2 + 2V_oV_{in}}}{w_r(V_o - V_{in})}$$  \hspace{1cm} (2)$$

When $V_{in} \leq V_{bound}$, $T_{ex\_SS} = 0$; when $V_{in} > V_{bound}$, $T_{ex\_SS} > 0$.

Then, during the first resonant time $T_{r1}$, both $S_1$ and $S_2$ are off, $V_{ds\_S1}$ decreases from $V_o$ to zero, and inductor current $i_L$ resonates through the minimum value, $i_{L\_valley}$.

$$i_{L\_valley} = -\frac{kV_{in}}{Z_n}$$  \hspace{1cm} (3)$$

After the resonant time $T_{r1}$, a small ZVS time margin $T_{ZVS}$ is provided. $S_1$ is turned on during $T_{ZVS}$. The inductor is then charged by the input voltage until turning off $S_2$ at $t_4$. During the second resonant time $T_{r2}$, $V_{ds\_S1}$ decreases from $V_o$ to zero, and $i_L$ resonates through the peak value, $i_{L\_pk}$.

$$i_{L\_pk} = \frac{V_{in}}{T_{on\_AS}}$$  \hspace{1cm} (4)$$

With the previous resonance, $S_1$ is turned on with ZVS at $t_5$, and the inductor is discharged by the output voltage. A PWM time-based counter is reset on the positive-to-negative zero current point at $t_6$ and starts a new switching period.

Since the resonant time intervals are very short, the inductor current can be approximated as a triangular waveform with average value $i_{L\_ave} \approx (i_{L\_pk} + i_{L\_valley})/2$. Assuming $v_{in} = \sqrt{2}V_{in\_rms},\sin\omega t$ and $i_{in} = (\sqrt{3}P_o/(\eta V_{in\_rms})) \sin\omega t$, the active switch conduction time $T_{on\_AS}$ is given by letting $i_{in} = i_{L\_ave}$.

$$T_{on\_AS} = \frac{2P_o L_n}{\eta V_{in\_rms}^2} + \frac{k}{w_r}$$  \hspace{1cm} (5)$$

where $\eta$ is the converter efficiency, $T_{on\_c} = \frac{2P_o L_n}{(\eta V_{in\_rms})^2}$ is constant at a steady-state operation, and $T_{on\_v} = \frac{k}{w_r}$ varies simultaneously with the input voltage. More detailed calculations of the time intervals are available in [11].

III. DIGITAL-BASED VARIABLE ON-TIME CONTROL WITH FAST DYNAMIC RESPONSE

A. Digital-Based Variable On-Time Control

To realize the full-line-cycle ZVS, digital-based variable on-time control is implemented. Fig. 4 presents the control structure with a digital signal processor (DSP). The input voltage $v_{in}$, output voltage $V_o$, and the inductor positive-to-negative zero current point are sensed through hardware circuits. In the DSP, a proportional integral (PI) controller regulates $V_o$ in the outer loop, and generates the on time $T_{on\_c}$ to control the peak inductor current. Meanwhile, PWM switching time intervals are calculated in real time based on the converter model and sensed voltage signals. The zero current detection (ZCD) signal is used to reset the PWM timer and limit the minimum current stress. Based on the time reference, devices $S_1$ and $S_2$ are turned off and on after the associated time delays.

$$\begin{align*}
\Delta T_{off\_SS} &= T_{ex\_SS} \\
\Delta T_{on\_AS} &= T_{ex\_SS} + T_{r1} \\
\Delta T_{off\_AS} &= \Delta T_{on\_AS} + T_{ZVS} + T_{on\_AS} \\
\Delta T_{off\_SS} &= \Delta T_{off\_AS} + T_{r2}
\end{align*}$$  \hspace{1cm} (6)$$

In addition, control signals for $S_3$ and $S_4$ are based on a phase locked loop (PLL) and are switched at line frequency.

B. Voltage-Loop Controller Design with Notch Filters

As aforementioned, the peak inductor current is controlled by the on-time $T_{on\_AS}$ which is composed of two parts, $T_{on\_AS} = T_{on\_c} + T_{on\_v}$, where $T_{on\_c}$ is the part generated from the low-bandwidth PI controller, and $T_{on\_v}$ is the variable part based on the real-time calculation. When $V_{in} > V_{bound}$, $T_{on\_v} = 0$ and $T_{on\_AS} = T_{on\_c}$. Hence, the voltage-loop controller is easier to be designed in the non-natural ZVS region with $V_{in} > V_{bound}$. Here, we pick the peak input voltage point to design the voltage controller.

The on-time-to-output open-loop transfer function is required for the controller design. Since $|i_{L\_valley}| \ll |i_{L\_pk}|$, the average inductor current can be simplified as

$$i_{L\_ave} \approx \frac{1}{2}i_{L\_pk} = \frac{1}{2} \frac{V_{in}}{T_{on\_c}}$$  \hspace{1cm} (7)$$

Applying KCL law on the output network, we get

$$C_{dc} \frac{dv_o}{dt} = i_{L\_ave}d' - \frac{v_o}{R_L}$$  \hspace{1cm} (8)$$

where the duty cycle $d' \approx \frac{v_{in}}{V_{o\_ref}}$ for the boost converter. Implementing perturbation and linearization on (7) and (8), the small-signal result is

$$i_{L\_ave} = \frac{V_{in}i_{on\_c} + \dot{v}_{in}T_{on\_c}}{2L_b}$$  \hspace{1cm} (9)$$

Fig. 4. Structure of the digital-based variable on-time control for the CRM totem-pole PFC.


\[ C_{dc}V_o \frac{d\hat{v}_o}{dt} = V_{in}^2 \hat{i}_{L,ave} + \hat{v}_{in}^2 I_{L,ave} - \frac{2V_o \hat{v}_o}{RL} \] (10)

Inserting (9) into (10), and rearranging it in the s domain, we obtain

\[ (sC_{dc}V_o + \frac{2V_o}{RL}) \hat{v}_o = V_{in}^2 \frac{\hat{i}_{on,c}}{2Lb} + (I_{L,ave})^2 \frac{V_{in}T_{on,c}}{2Lb} \hat{v}_{in} \] (11)

Then, the on-time-to-output transfer function is

\[ G_{olt} = \left. \frac{\hat{v}_o}{\hat{v}_{in}} \right|_{\hat{v}_{in}=0} = \frac{V_{in}^2 RL}{4V_o Lb} \frac{1}{1 + \frac{RC_{d}}{2RL}} \] (12)

Since the open-loop transfer function has only one pole, a PI controller is enough to compensate the loop. In order to improve the transient response, higher control bandwidth is desired. To avoid the impact of the double-line frequency ripple, digital notch filters are applied on the output voltage feedback loop. Because the whole circuit operates with 60 Hz AC line input, notch filters with stop bands at 60 Hz are implemented. As depicted in Fig. 5, assuming the analog-to-digital (A/D) conversion gain and the PWM modulation gain are cancelled out, the final loop gain of the voltage control loop is

\[ T_v = G_{sense}G_{notch_d}G_{RC_d}G_{PI}G_{olt} \] (13)

where \( G_{RC_d} \) is the digital low-pass filter for noise suppression. Fig. 6 shows the bode plot of the PFC loop gain at peak input voltage. The PI compensator is designed to have 160 Hz control bandwidth with a phase margin of 65°.

\section*{IV. IMPLEMENTATION CONSIDERATIONS FOR NOISE IMMUNITY}

\subsection*{A. Hardware Implementation}

A prototype of a GaN-based 100 W multi-receiver wireless charging power supply system is developed and shown in Fig. 7. The WPT system contains a two-stage transmitter, a 0.5 × 0.5 m² transmitter desk, receiver coils, and two diode receivers. The two-stage transmitter is composed of EMI filters, a CRM totem-pole PFC converter, and a 6.78 MHz inverter with total volume at 7.4 × 5.3 × 1.7 cm³. Detailed specifications of the transmitter are listed in Table I.

\begin{table}[h]
\centering
\caption{Specifications of the transmitter for the WPT system.}
\begin{tabular}{|c|c|}
\hline
Parameter & Value \\
\hline
Input voltage \( v_{in} \) & 120 V ac, 60 Hz \\
DC bus voltage \( V_{dc} \) & 200 V dc \\
Output power \( P_o \) & 100 W \\
PFC switching frequency \( f_{s,pfc} \) & 187 - 725 kHz \\
Inverter switching frequency \( f_{s,in} \) & 6.78 MHz \\
GaN devices \( S_1 - S_4 \) & GS66508B, 650 V \\
PFC ZVS margin & \( k_o = 1.1 \), \( T_{zvs, min} = 50 \) ns \\
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\end{tabular}
\end{table}

Fig. 8. Experimental waveforms of the CRM PFC at full-load without the inverter and WPT coils.
C. Analysis of the Noise Propagation

To understand the reason for the noise impacts on sensing signals, common-mode (CM) noise propagation paths within the power supply system are analyzed, as illustrated in Fig. 9. High \( \frac{dv}{dt} \) noise is generated by the switching nodes \( v_{sw1}, v_{sw2}, v_{sw3} \), and the resulting noise currents flow to the common reference ground \( (E_{GND}) \) through parasitic capacitance \( C_{p1}, C_{p2}, C_{p4} \). Since the transmitter coil has large area, parasitic capacitance \( C_{p5} \) between the coil and ground cannot be neglected. Parasitic capacitance \( C_{p6} \) between the transmitter coil and receiver coils transfers the noise current from the primary side to the load side, and further injects the noise current into the ground via \( C_{p7} \). The noise currents carried by \( C_{p6} \) are involved in the resonant tank of the WPT coils. Hence, the current injected to the ground through \( C_{p7} \) has 6.78 MHz resonant ripple. In addition, capacitance \( C_{p3} \) between the power ground \( (P_{GND}) \) and \( E_{GND} \) also contributes to the noise propagation.

The noise currents are then transferred to the input side, and flow back forming a complete conduction loop. Ideally, all the noise currents can be picked up by \( C_{CM} \) in the EMI filter, and the sensing circuits will not be influenced. However, constrained by the allowable leakage current [17], the total values of \( C_{CM} \) may not be sufficient to absorb all the noise current, especially the large amount of 6.78 MHz noise. So partial noise is propagated to the sensing circuits. Due to the ground connection shown in [18], sensing circuits of \( v_{ac} \) and \( V_{dc} \) share the same analog ground \( A_{GND} \), and the high-frequency noise from the power stage is easily transferred to the voltage sensing circuits and the digital controller.

A ZCD circuit based on a sensing resistor \( R_{sense} \) is adopted, where \( R_{sense} \) is connected in series with the input power line [16]. The inductor current is detected in each switching cycle, and the ZCD circuit is very sensitive in the high-frequency environment. Although isolation is adopted to seclude the floating ground \( (F_{GND}) \), part of the noise current still flows into \( F_{GND} \) through parasitic capacitance \( C_{p6} \), and the sensing signals are further distorted. Fig. 12 shows the simulated results of current sensing signals in the ZCD circuit. In the ideal case, the amplifier input differential signal \( v_{d,amplifier} \)
is nearly zero, and the amplifier output \( v_{amplifier} \) is a clean triangular waveform following the inductor current. Accordingly, the comparator generates clean ZCD signal with small propagation delay time of 35 ns (Fig. 12(a)). Nevertheless, with the noise impact, \( v_{d,amplifier} \) is coupled with not only the switching noise from the switching nodes, but also the 6.78 MHz resonant ripple from the WPT coils. The CM noise is converted into differential mode (DM) noise, and \( v_{amplifier} \) is distorted with high-frequency resonant ripple and a small DC bias. As a result, the ZCD signal from the comparator is noisy and has much longer time delay of 279 ns. In AC-DC operation, 60 Hz line-cycle ripple is also reflected on the noise current, and further distorts the amplifier output, leading to the line-cycle varying ZCD time delay as shown in Fig. 11(a).

D. Proposed Methods for Noise Immunity

1) Isolated Sensing Circuit with Filter

To suppress the high-frequency noise coupled on \( v_{ac} \) and \( V_{dc} \) sensing, voltage sensing circuits with EMI filters and high common-mode transient immunity (CMTI) isolated amplifiers are used, as depicted in Fig. 13. The bandwidths of the EMI filters should be designed between the voltage control bandwidth and the high switching frequency to avoid the apparent filtering delay on the feedback loop. Since the noise is predominately in the MHz in the WPT system, a filter corner frequency in a few kHz is sufficient to filter out the noise and has no impact on the feedback control. For the differential voltage sensing circuit, the voltage divider resistors should be considered in the EMI filter design. Corner frequencies of the CM and DM filters are determined as

\[
\begin{align*}
    f_{CM} &= \frac{1}{2\pi (R_H || R_L + R_Y) C_Y} \\
    f_{DM} &= \frac{1}{2\pi (R_H || R_L + R_Y) (C_Y + 2C_X)}
\end{align*}
\]

(14)

(15)

In order to avoid the erroneous ZCD signal due to the high-frequency noise, RC filters are placed on the input terminals of the comparator in the ZCD circuit, as shown in Fig. 14. Also, a symmetrical configuration with \( R_5 = R_6 \) and \( C_1 = C_2 \) is employed to balance and improve the CM impedance.

2) Balanced and High CM Impedance

To eliminate the varying ZCD time delay and prevent the CM noise from flowing to the ZCD circuit, a large and balanced CM impedance should be inserted in the middle of the system circuit. Typically, a EMI filter with larger CM choke on the DC bus is effective. However, with the limited converter size for such high-density power supply, an extra EMI filter is not desired. For the WPT application, the impedance matching network (IMN) tank can be adjusted to an L-C-L configuration instead of an L-C structure. As illustrated in Fig. 15, \( L_{IMN} \) is modified from a single inductor to a coupled inductor with the same inductance on the DM loop. To achieve higher CM impedance, a lower coupling coefficient \( k \) is realized by separating the windings to opposite ends of the low-permeability core (Fig. 15(c)). As listed in Table II, for the same core and winding wire, the CM impedance varies widely with different inductor structures.

![Fig. 12. Simulated signals of the amplifier and comparator in the ZCD circuit. (a) Normal signals. (b) Distorted signals with noise impact.](image1)

![Fig. 13. Differential voltage sensing circuit with EMI filters.](image2)

![Fig. 14. ZCD circuit.](image3)

![Fig. 15. IMN inductor. (a) Original single \( L_{IMN} \). (b) Coupled \( L_{IMN} \) with higher CM impedance. (c) Coupled \( L_{IMN} \) with higher CM impedance.](image4)

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<tbody>
<tr>
<td>( L_{DM} )</td>
<td>1.68 uH</td>
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<td>1.68 uH</td>
</tr>
<tr>
<td>( L_{CM1} )</td>
<td>1.68 uH</td>
<td>0.55 uH</td>
<td>0.73 uH</td>
</tr>
<tr>
<td>( L_{CM2} )</td>
<td>0 uH</td>
<td>0.57 uH</td>
<td>0.7 uH</td>
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<td>( k )</td>
<td>0</td>
<td>0.4822</td>
<td>0.1816</td>
</tr>
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V. EXPERIMENTAL VERIFICATION

As previously demonstrated, combined with the 6.78 MHz inverter and WPT coils, the CRM PFC cannot operate normally because of the noise impact (Fig. 11). To enhance the operation performance, the proposed noise immunity approaches are implemented, including the isolated voltage sensing circuits, filters, and the modified IMN inductor.

Both the current and voltage sensing signals are much cleaner, and the ZCD signal varying time delay is eliminated, as shown in Fig. 16. The CRM PFC is able to operate reliably in the 6.78 MHz WPT system. Fig. 17 and Fig. 18 present the tested PFC waveforms in the full-load operation of the WPT system. With 120 V\textsubscript{ac} input voltage, the PFC generates 200 V DC bus voltage, and realizes ZVS within the whole line cycle. Based on the measurement with the power analyzer Yokogawa WT3000E, the tested end-to-end power efficiency of the WPT power supply system is 90.16%.

To validate the capability of fast dynamic response, a load transient is conducted by removing one of the receivers. Fig. 19 and Fig. 20 display the tested waveforms of the CRM PFC during the transition between full load and half load, and Fig. 21 shows the transmitter coil current during the load transient. As can be seen, fast transient response is achieved with very small DC bus voltage fluctuation, and the 6.78 MHz transmitter coil current is not influenced during the process.

Fig. 16. Improved PFC sensing signals with 6.78 MHz inverter and WPT coils. (a) Clean ZCD signal. (b) Clean voltage sensing signals.

Fig. 17. Experimental waveforms of the CRM PFC at full-load operation of the 100 W 6.78 MHz WPT system.

Fig. 18. Experimental switching waveforms of the CRM PFC at full-load operation of the 100 W 6.78 MHz WPT system. (a) When \(V\textsubscript{in} = 130\) V. (b) When \(V\textsubscript{in} = 170\) V.

Fig. 19. Experimental transition waveforms of the CRM PFC in the 6.78 MHz WPT system during load variation between full load and half load. (a) From full load to half load. (b) From half load to full load.

Fig. 20. Experimental transition waveforms of the CRM PFC in the 6.78 MHz WPT system during load variation between full load and half load. (a) From full load to half load. (b) From half load to full load.

Fig. 21. Experimental transmitter coil current during load variation between full load and half load (a) Long-scale waveform. (b) Zoom-in waveform.
VI. CONCLUSIONS

A GaN-based CRM totem-pole PFC converter with ZVS control is developed for a 100 W 6.78 MHz wireless charging power supply system with multiple receivers. To meet the requirement of fast dynamic response, digital-based voltage-loop controller with high bandwidth and notch filters are employed. In order to avoid the noise impact and maintain stable ZVS control in the high-frequency environment, fully isolated sensing circuits with EMI filters are implemented, and balanced CM impedance are achieved by coupled IMN inductor. The proposed methods for fast transient response and noise immunity have the advantages of easy implementation, and do not require additional components and control. A prototype of a 100 W GaN-based 6.78 MHz WPT system is developed, and demonstrated with 90.16% end-to-end full-load efficiency. PFC operates reliably with ZVS control, and fast dynamic response is validated during load changing.

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