






Pulse Magnitude Modulation and Token Rotation-Based Voltage Balancing Method of Multilevel Inverter for Wireless Power Transfer

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Abstract—The multilevel inverter has attracted more and more attention in high-power wireless power transfer (WPT) systems, but its switching loss and capacitor voltage balancing are worrying issues. To solve these problems, this article proposes and implements a sigma-delta (Σ - Δ) pulse magnitude modulation (PMM) of a flying-capacitor multilevel inverter (FCMLI) for WPT systems. It can achieve wide-range output voltage regulation and zero voltage switching simultaneously. Besides, a token rotation-based flying-capacitor voltage balancing method of the FCMLI for WPT is proposed. It can balance the capacitor voltages of the FCMLI with arbitrary voltage levels while maintaining little ripples and resisting the step dynamic. The corresponding parameter selection of the flying capacitors is discussed. Theoretical analysis, computer simulation, and an 850 W seven-level flying capacitor inverter-based hardware experimentation are given to verify the effectiveness of the proposed Σ - Δ PMM FCMLI-based WPT system.

Index Terms—Capacitor voltage balance, multilevel inverter, pulse magnitude modulation (PMM), wireless power transfer (WPT), zero voltage switching.

I. INTRODUCTION

WIRELESS power transfer (WPT) technology can be traced back over one century to the work of Nikola Tesla and has been applied to a wide range of fields with power levels ranging from milliwatts to megawatts [1], [2], [3], helping to charge micro-robots, electric toothbrushes, electric vehicles [4], and high-speed train [5], due to electrical isolation, the safety, convenience, and better user experience. Identified as an epoch-making technology, the WPT has been attracting

a growing number of academic researchers and engineers to participate in the great explorations.

In recent years, WPT systems based on multilevel inverters have attracted more and more attention due to the advantage of realizing a large operating range [6], [7], breaking voltage limitations [8], [9], [11], [12]. On the one hand, a multilevel inverter-based WPT system has been studied for low-voltage high-power EV charging whose loading conditions can vary up to 450% [6], [7]. On the other hand, a medium-voltage (1.5–50 kV) dc distribution network is often equipped in high-power applications [10] to decrease the current in the dc bus, such as the submarine data center and fast charging of EVs. Multilevel inverters-based WPT systems can be connected to the medium-voltage dc-power system to reduce the voltage stress for each semiconductor component [11], [12]. Moreover, multilevel inverters have been applied for decoupled multichannel WPT systems to eliminate the transformers and simplify the system configuration [13]. It is noticed that the modulation method and capacitor voltage balancing of multilevel inverter-based WPT systems are two key problems to be further investigated.

Recently, many significant modulation schemes of two-level inverter-based WPT systems have been fully investigated, including pulsewidth modulation (PWM), ON-OFF keying modulation, hybrid modulation, pulse density modulation (PDM), pulse frequency modulation (PFM), and step density modulation (SDM). As a kind of PWM with a controllable angle, phase-shift control [14], [15] is generally utilized in the converters of the WPT system, but it suffers from hard switching. ON-OFF modulation uses the low-frequency ON-OFF duty ratios to control the converter [16], but causes large output ripples. Hybrid modulation combining the PFM and ON-OFF modulation suffers from the same issues [17]. By comparison, PDM [18], [19], [20] and PFM are more suitable for WPT systems due to breaking through the above limitations. A low-subharmonic, full-range, and rapid PDM strategy is proposed in [20] for the full-bridge converters-based WPT system. The PFM for WPT systems was proposed in [21], [22], [23], and it regulates the output power by using high-order harmonics of the square wave at a specific frequency. The studies of the Σ - Δ PFM are also presented in [22] to decrease the output ripple, but the realization is a little complicated. Recently, SDM, which uses the density of steps

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to regulate the output power, was proposed in [24], combining both the advantages of PDM and PFM.

However, the above methods are unsuitable for the multilevel inverter-based WPT system. Liu et al. [13] proposes the PWM T-type multilevel inverter-based multichannel WPT system, and switching signals are exchanged to balance neutral voltages, but there is a hard switching problem. The zero voltage switching (ZVS) modulation method of the modular multilevel converter (MMC)-based WPT system is proposed [7], but the output voltage of the MMC-based WPT system is not continuous, and the capacitor voltage balancing method of multilevel inverter with arbitrary voltage levels remains unclear. It is unsuitable for other multilevel inverters such as T-type or flying-capacitor multilevel inverters (FCMLIs) [25].

Our previous work proposes the segmented vector PFM method of the flying-capacitor three-level inverters for the WPT system [26], but the total harmonic distortion (THD) of the transmit current is high especially when under the light load, which could lead to EMI noise that will influence the system stability. Besides, due to the capacitor voltage balancing, the precious method can only be used for the three-level inverter-based WPT system. Moreover, a ZVS modulation method with low THD and dc-side voltage ripple for the FCMLI needs to be further investigated.

The voltage balancing of the FCMLI-based WPT is a key issue to be solved. Many flying capacitor voltage balance methods have been proposed for 50 Hz/60 Hz FCMLIs [27], [28], [29], [30], which are based on the low-frequency phase currents and voltage deviation of flying capacitors [31], [32]. Different from the 50 Hz/60 Hz FCMLI, the capacitor voltage balancing method of the FCMLI for the high-frequency WPT system has more challenges. First of all, the algorithm needs to be completed in a more high-speed control cycle, which is half the resonant cycle of the WPT system. Second, the high-frequency transmitter current in the WPT system will not be measured. Besides, the capacitor voltage balancing method of the FCMLI needs to match the ZVS modulation method. Moreover, the voltage balancing approach needs to be easy to implement and easily portable to the FCMLIs with arbitrary voltage levels.

To realize the soft-switching modulation, balance capacitor voltages of the FCMLI in a high-speed control cycle, and reduce the dc-side voltage ripple and THD of the transmitter current of the FCMLIs-based WPT systems, this article proposed an Σ - Δ pulse magnitude modulation (PMM) method of the multilevel inverter and a token rotation-based capacitor voltage balancing method, the main contribution of this article is listed as follows.

- 1) The Σ - Δ PMM of the FCMLI for WPT systems is proposed to realize the wide-range continuous DC-side output voltage regulation and ZVS simultaneously, reduce the THD of the transmitter current, and decrease the dc-side voltage ripple.
- 2) A token rotation-based capacitor voltage balancing method of the FCMLI for the WPT system is proposed.
- 3) The parameter design method of the flying capacitor of the FCMLI for the WPT system is given.

The rest of the article is organized as follows. Section II presents the proposed Σ - Δ PMM method of the FCMLI-based

WPT system. Section III presents the proposed token rotation-based capacitor voltage balancing method of the FCMLI for the WPT system. Section IV presents the simulation and experimental verification. Finally, Section V concludes this article.

II. DELTA-SIGMA PULSE MAGNITUDE MODULATED MULTILEVEL INVERTER-BASED WPT SYSTEM

A. Description of the FCMLI-Based WPT System

Fig. 1 shows a typical WPT system based on the FCMLI, where C_{dc1} and C_{dc2} are the dc-side capacitors; u_1, u_2, u_3, \dots , and u_{n-2} are the voltages of the flying capacitors $C_{f1}, C_{f2}, C_{f3}, \dots$, and C_{fn-2} ; L_t and L_r are the inductances of the transmitter and receiver coils; R_t and R_r are the equivalent series resistances of the transmitter and receiver parts; M denotes the mutual inductance between the primary coil and the secondary coil; C_t and C_r are the capacitances of series compensated capacitors, respectively; R_{dc} and C_{dc} are the dc-link load and capacitor of the diode rectifier. The FCMLI is fed by a dc input voltage V_{dc} and drives a series-series resonant tank. The proposed control of the multilevel inverter consists of two parts: the Σ - Δ -based PMM and the flying-capacitor voltage balancing method. The voltages of the flying capacitors u_1, u_2, u_3, \dots , and u_{n-2} are sampled for capacitor voltage balancing. The proposed control of the FCMLI-based WPT system will be discussed in Sections II-B and III in detail. The resonant frequencies of the resonators on the two sides are given as

$$f_r = \frac{1}{2\pi\sqrt{L_t C_t}} = \frac{1}{2\pi\sqrt{L_r C_r}}. \quad (1)$$

The input impedance on the transmitter side of the WPT system can be derived as follows:

$$Z_{in} = j\omega L_t + \frac{1}{j\omega C_t} + R_t + \frac{\omega^2 M^2}{j\omega L_r + \frac{1}{j\omega C_r} + R_r + R_{eq}} \quad (2)$$

where R_{eq} is the equivalent ac load of the diode load.

To equalize the voltage of each switch, the voltages of the m_{th} flying capacitor C_{fm} of the FCMLI should be controlled as

$$V_c(m) = \frac{(n-1-m)V_{dc}}{n-1}, m = 1, 2, \dots, n-2. \quad (3)$$

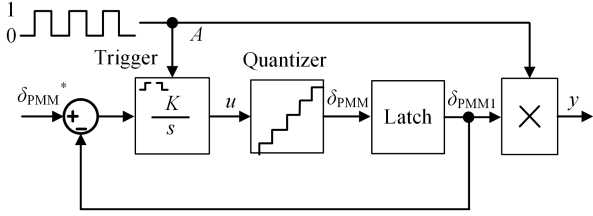
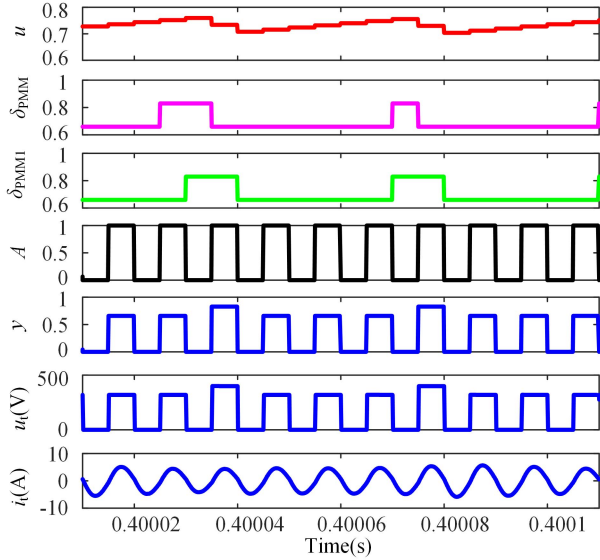
The output voltage of the n -level flying-capacitor inverter is equal to the sum of the MOSFET voltages of all lower side arms, which can be derived as

$$u_t = \frac{V_{dc}}{n-1} \sum_{m=1}^{n-1} S_m. \quad (4)$$

Except for 0 and V_{dc} , the other voltage levels have redundant switching states. As shown in Fig. 2, according to (4), it can be concluded that there are Cm $n-1$ kinds of switching states to produce the output voltage of $mV_{dc}/(n-1)$ for the n -level flying capacitor inverter.

B. Proposed PMM Method

The series-series compensated WPT system exhibits the bandpass characteristics around the resonant frequency. It can


 Fig. 4. Proposed Σ - Δ PMM for the multilevel inverter.

 Fig. 5. Waveforms of Σ - Δ modulator with $\delta^* = 0.7$ of seven-level FCMLI.

edges of the input pulses trigger the integrator, and the input of the integrator is the difference between the given δ_{PMM}^* and the output of the quantizer. It is noticed that the output of the integrator needs to be capped in $[0, 1]$. According to the output u of the integrator, the quantizer will output the magnitude δ_{PMM} . The quantizer of the Σ - Δ PMM can be given as follows:

$$\begin{cases} \delta_{\text{PMM}} = 1, u \geq \frac{2n-3}{2(n-1)} \\ \delta_{\text{PMM}} = \frac{m}{n-1}, \frac{2m+1}{2(n-1)} \geq u \geq \frac{2m-1}{2(n-1)}, m = 1, 2, \dots, n-2 \\ \delta_{\text{PMM}} = 0, u < \frac{1}{2(n-1)} \end{cases} \quad (8)$$

where n is the voltage level number of the multilevel inverter.

Then, the latch is used to keep δ_{PMM} for at least two control cycles, because the whole-cycle voltage pulse is the minimum unit of the output voltage of the half-bridge multilevel inverter for the WPT system. The modulation wave is produced by multiplying the output of the quantizer and the input pulse A . Finally, the modulation wave will be used as the input of the capacitor voltage balance method in the next chapter. Fig. 5 shows the waveforms of the proposed Σ - Δ PMM modulator and the output voltage and current of the multilevel inverter in detail. The relationship between the output of the Σ - Δ PMM modulator and the output voltage of the multilevel inverter can

be given as follows:

$$u_t(t) = y \frac{V_{\text{dc}}}{2}. \quad (9)$$

It can also be found that the current direction can be judged by the given input pulse (A). The transmitter current is positive when $A = 1$, and is negative when $A = 0$.

D. ZVS Analysis

To enable the robust ZVS of the multilevel inverter-based WPT system, the input impedance of the series-series compensated WPT system needs to be slightly inductive at the switching frequency, which can be realized by slightly increasing the transmitter capacitor. Then, the magnitude and phase of the input impedance Z_{in} at the switching frequency can be approximately derived as [26]

$$|Z_{\text{in}}| \approx R_t + \frac{\omega_r^2 M^2}{R_r + R_{\text{eq}}} \quad (10)$$

$$\tan \alpha \approx \frac{(R_{\text{eq}} + R_r) \Delta C}{(\omega_r^2 M^2 + R_t R_r + R_t R_{\text{eq}}) \omega_r C_r^2} = k \frac{\Delta C}{C_r}. \quad (11)$$

The ZVS of the multilevel inverter can be realized when Q_{zvs} , defined as the integral of i_t during the dead time, can charge/discharge all the drain-source capacitors of the MOSFETs from zero to $V_{\text{dc}}/(n-1)$, or vice versa. Taking a seven-level inverter as an example, while “ $S_1 S_2 S_3 S_4 S_5 S_6$ ” is changed from “111111” to “000000,” six upper switches need to be charged from zero to $V_{\text{dc}}/6$, and six lower switches need to be charged from $V_{\text{dc}}/6$ to zero during the dead time. Thus, in order to realize the ZVS of n -level inverter, it can be obtained as follows:

$$Q_{\text{zvs}} = \int_0^{T_d} i_t dt \geq \int_0^{V_{\text{dc}}/(n-1)} (C_{\text{oss}1} + C_{\text{oss}2} + \dots + C_{\text{oss}m} \dots + C_{\text{oss}2n-2}) dv \quad (12)$$

where $C_{\text{oss}1}, C_{\text{oss}2}, \dots,$ and $C_{\text{oss}2n-2}$ represent the drain-source capacitors of $2n-2$ MOSFETs, respectively.

Assuming that the transmitter current i_t during the dead time is part of the sine waves and all the C_{oss} are the same, (12) can be derived as follows:

$$\int_0^{T_d} I_t \sin(\omega_r t) dt \geq \int_0^{V_{\text{dc}}/(n-1)} 2(n-1) C_{\text{oss}} dv. \quad (13)$$

Then, according to (13), the minimum phase difference α_{min} and dead time T_{dmin} can be derived as

$$\cos(\alpha_{\text{min}}) = 1 - \frac{2\omega_r C_{\text{oss}} V_{\text{dc}}}{I_t} \quad (14)$$

$$T_{\text{dmin}} = \frac{\alpha_{\text{min}}}{\omega_r}. \quad (15)$$

According to (6), ignoring the influence of the harmonic current, I_t can be calculated as

$$I_t = \frac{2\delta_{\text{PMM}} V_{\text{dc}}}{\pi |Z_{\text{in}}|}. \quad (16)$$

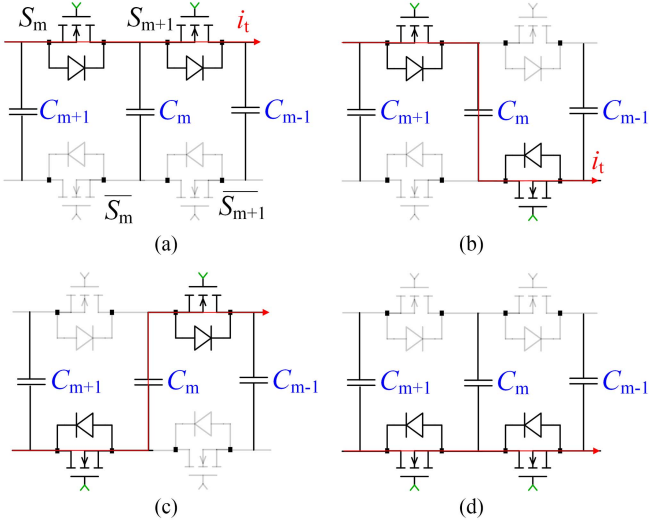


Fig. 6. Current flow of sub-circuit of the FCMLI in different switch states of “ $S_m S_{m+1}$ ”. (a) 11 (b) 10 (c) 01 (d) 00.

Besides, it can be obtained as

$$(\tan \alpha)^2 + 1 = \frac{1}{(\cos \alpha)^2}. \quad (17)$$

Based on (10)–(17), $(\Delta C)/C_r$ can be derived as follows:

$$\frac{\Delta C}{C_r} \geq \frac{1}{k} \sqrt{\frac{\delta_{\text{PMM}}^2}{(\delta_{\text{PMM}} - \pi \omega_r C_{\text{oss}} |Z_{\text{in}}|)^2} - 1}. \quad (18)$$

Although (18) contains no n and V_{dc} , the dc-side voltage of the MOSFET $V_{\text{dc}}/(n-1)$ will influence C_{oss} . A bigger C_{oss} and a bigger δ_{PMM} lead to a bigger ΔC . Limited by the transfer power of the WPT system, $\Delta C/C_r$ is suggested to be less than 5% to realize wide-range ZVS during power control [26].

III. PROPOSED VOLTAGE BALANCING METHOD OF MULTILEVEL INVERTER FOR WPT SYSTEM

A. Switching State Analysis

Whether a flying capacitor is charging or discharging depends on the transmitter current (i_t) and the way the flying capacitor is connected to the circuit. The charge or discharge of C_m is determined by the switching state of adjacent switch sets “ $S_m S_{m+1}$ ”. As shown in Fig. 6(a) and (d), while “ $S_m S_{m+1}$ ” is selected as “11” and “00,” C_m is bypassed, and the transmitter current flows on the upper and lower branches, respectively. As shown in Fig. 6(b) and (c), while “ $S_m S_{m+1}$ ” is selected as “10” and “01,” C_m is charged and discharged by the transmitter current, respectively. Moreover, while i_t is the opposite and “ $S_m S_{m+1}$ ” is selected as “10” and “01,” C_m is discharged and charged by the transmitter current, respectively. The key information of the FCMLI can be summarized as follows.

- 1) The output voltage of the multilevel inverter depends on the number of one in “ $S_1 S_2 S_3 \dots S_{n-1}$.”
- 2) The charging or discharging pattern of the flying capacitors $C_{f1}, C_{f2}, C_{f3}, \dots,$ and C_{fn-2} depends on the switching

TABLE I
SWITCHING STATES OF SEVEN-LEVEL INVERTER ($U_T = 0.5V_{\text{dc}}, I_T > 0$)

$S_1 S_2 S_3 S_4 S_5 S_6$	u_t	C_{f1}	C_{f2}	C_{f3}	C_{f4}	C_{f5}
000111	$0.5V_{\text{dc}}$	—	—	↓	—	—
001011	$0.5V_{\text{dc}}$	—	↓	↑	↑	—
001101	$0.5V_{\text{dc}}$	—	↓	—	↓	↓
001110	$0.5V_{\text{dc}}$	—	—	↓	↑	—
010011	$0.5V_{\text{dc}}$	—	↓	↑	—	—
010101	$0.5V_{\text{dc}}$	↓	↑	—	—	—
011001	$0.5V_{\text{dc}}$	↑	—	—	—	—
011010	$0.5V_{\text{dc}}$	—	—	—	↓	—
010110	$0.5V_{\text{dc}}$	—	—	—	—	↑
011100	$0.5V_{\text{dc}}$	—	—	—	—	—
100011	$0.5V_{\text{dc}}$	—	↓	—	↑	—
100101	$0.5V_{\text{dc}}$	↓	—	↑	—	—
100110	$0.5V_{\text{dc}}$	—	↑	—	—	—
101001	$0.5V_{\text{dc}}$	↑	—	—	—	↓
101010	$0.5V_{\text{dc}}$	—	—	↑	—	—
101100	$0.5V_{\text{dc}}$	—	—	↓	—	—
110001	$0.5V_{\text{dc}}$	↓	—	—	—	↑
110010	$0.5V_{\text{dc}}$	—	↓	—	—	—
110100	$0.5V_{\text{dc}}$	↑	—	—	—	—
111000	$0.5V_{\text{dc}}$	—	—	↑	—	—

states of “ $S_1 S_2$,” “ $S_2 S_3$,” “ $S_3 S_4$,” \dots , and $S_{n-2} S_{n-1}$, and the direction of transmitter current.

According to the above analysis, the charge/discharge states of every flying capacitor can be easily judged by the switching states of $S_1 S_2 \dots S_{n-1}$. For example, Table I presents all the charge/discharge states of the seven-level flying-capacitor inverter when the output voltage is $0.5V_{\text{dc}}$ and i_t is positive, where “↓,” “↑,” and “—” mean the discharging, charging, and keeping of the flying capacitors, respectively. Furthermore, one special case where the switching state ($S_1 S_2 \dots S_6$) is “010101” will be explained. Considering the switching states of $S_1 S_2, S_2 S_3,$ and $S_3 S_4,$ are “01,” “10,” “01,” “10,” and “01,” thus, when the transmitter current is positive, C_{f1} is discharging, C_{f2} is charging, C_{f3} is discharging, C_{f4} is charging, and C_{f5} is discharging based on the above analysis.

B. Token Rotation-Based Flying-Capacitor Voltage Balancing Method

As is required for the FCMLI, the voltage stress of each MOSFET is supposed to be the same or similar value. Thus, the voltages of flying capacitors need to be controlled at the reference value. Fig. 7 shows the token rotation-based charging/discharging sequence of the flying capacitors, where the inner ring represents the charging priority of the capacitors $C_{f1}, C_{f2}, C_{f3}, \dots,$ and C_{fn-2} , and the outer ring represents the position of the token. The position of the token is cyclic along the pointer and it represents the highest priority of charging and discharging of the capacitor. For example, when the token reaches the position of C_{f1} , the priority for the charging or discharging of the flying capacitors is “ $C_{f1} > C_{f2} > C_{f3} > \dots > C_{fn-2}$.” Then, the token reaches the position of C_{fn-2} , and the priority for the charging/discharging of the flying capacitors becomes “ $C_{fn-2} > C_{f1} > C_{f2} > C_{f3} > \dots > C_{fn-3}$.” After each charging or discharge process is completed, the capacitor charge or discharge priority is changed in turn to ensure that each capacitor voltage can be balanced at the reference value.

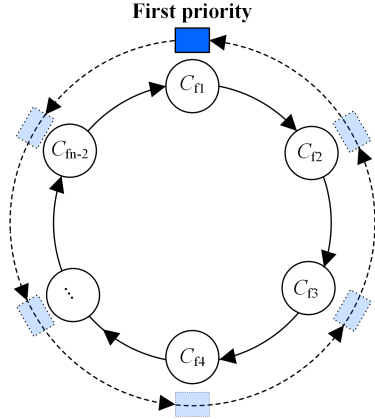


Fig. 7. Token rotation-based charging/discharging sequence of the flying capacitors.

First of all, the voltage across each flying capacitor is sensed by an isolated voltage-sensing circuit, and the ADC values $u_1, u_2, \dots, u_m, \dots, u_{n-2}$ are passed to the microcontroller. The deviations of the flying capacitor voltages from their nominal values are given as follows:

$$\begin{cases} a_1 = u_1 - \frac{n-2}{n-1} V_{dc} \\ a_2 = u_2 - \frac{n-3}{n-1} V_{dc} \\ \vdots \\ a_m = u_m - \frac{n-1-m}{n-1} V_{dc} \\ \vdots \\ a_{n-2} = u_{n-2} - \frac{1}{n-1} V_{dc} \end{cases} \quad (19)$$

Based on the signs of $A_n = [a_1, a_2, \dots, a_m, \dots, a_{n-2}]$, the optimal switching state will be selected to balance the capacitor voltage. Fig. 8 shows the process of switching state selection based on A_n , which is divided into two parts. First of all, according to A_n , the optimal switching states ($B_n = [b_1, b_2, \dots, b_m, \dots, b_{n-1}]$) can be selected. Then, according to modulation wave y , the number of one in B_n can be obtained, further determining the final switching states ($S_{nf} = [S_1, S_2, \dots, S_m, \dots, S_{n-1}]$).

Fig. 9 shows the flow chart of the proposed token rotation-based flying capacitor voltage balance method over one control cycle. The steps of the proposed token rotation-based flying capacitor voltage balance method of the n -level FCMLI can be summarized as follows.

Step 1: The first priority is rotated from $C_{fn-2}-C_{fn-3}-\dots-C_{fm}-\dots-C_{f3}-C_{f2}-C_{f1}-C_{fn-2}$. It can be realized by a priority sign j in Fig. 9. For example, when $j = 1$, the priority of capacitor charge and discharge is $C_{f1} > C_{f2} > \dots > C_{fm} > \dots > C_{fn-2}$. Only when $0 < y < 1$, does j need to subtract 1 to change the order of priority.

The output of the wave generator (y) shown in Fig. 5 consists of “1,” “ $(n-2/n-1)$,” \dots , “0.” When $y = 1$ and 0, the voltage levels of the multilevel inverter are V_{dc} and 0, respectively. Thus, the switching state is “111...111” and “000...000,” and

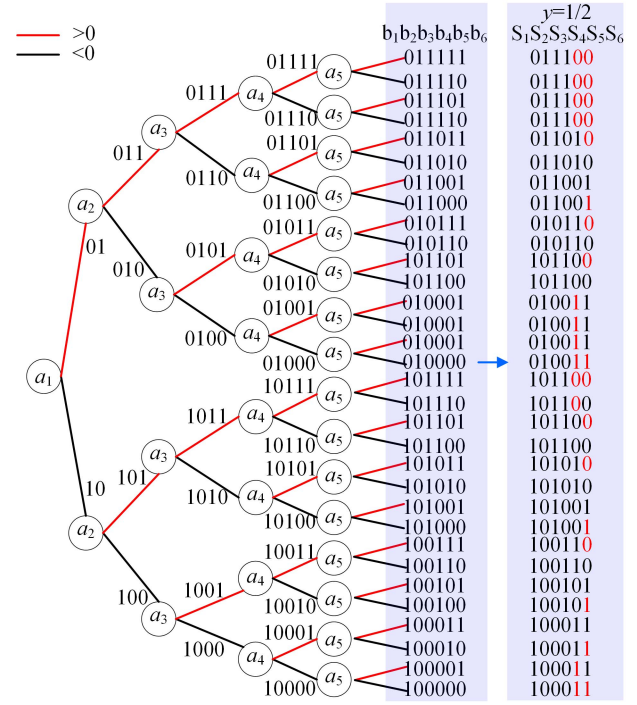


Fig. 8. Tree-map-based ideal switching state for different voltage samples of the seven-level FCMLI when C_{f1} is the first priority and $i_t > 0$.

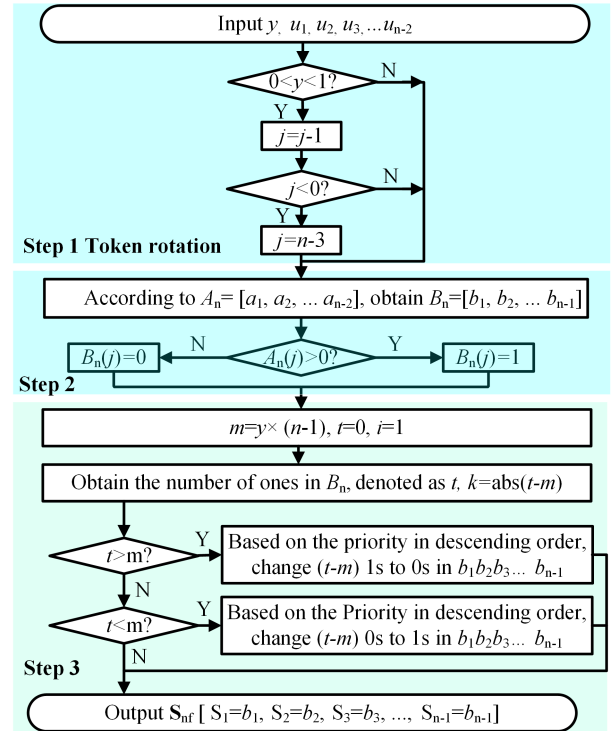


Fig. 9. Flow chart of the token rotation-based flying capacitor voltage balance method for n -level inverter-based WPT system.

no capacitor is charging/discharging. Therefore, the order of priority is kept when $y = 1$ or 0.

Step 2: Sampling the voltages ($u_1, u_2, u_3, \dots, u_{n-2}$) of $n-2$ flying capacitors, then, obtain $a_1 a_2 a_3 \dots a_{n-2}$ according to (20). According to the priority sign, obtain the optimal $b_1 b_2 b_3 \dots b_{n-1}$. First of all, if $a_1 > 0$, " $b_1 b_2$ " is selected as "01"; otherwise, " $b_1 b_2$ " is selected as "10." Then, if $a_m > 0$, $b_{m+1} = 1$; otherwise, $b_{m+1} = 0$, and $m = 2, 3, 4, \dots, n-2$. Furthermore, if $a_j > 0$, $b_j = 0$; otherwise, $b_j = 1$.

Step 3: According to the output of the Σ - Δ PMM for the multilevel inverter (y), calculate the limited number of ones in the desired switching states, defined as i . Based on (4), the calculation method is given as follows:

$$i = y(n-1). \quad (20)$$

Then, calculate the number of ones in $b_1 b_2 b_3 \dots b_{n-1}$, which is calculated as follows:

$$j = b_1 + b_2 + b_3 + \dots + b_{n-1}. \quad (21)$$

If i equals to j , $b_1 b_2 b_3 \dots b_{n-1}$ keeps unchanged; If $i > j$, in descending order of the priority sequences, change ($i-j$) 1s to 0s in $b_1 b_2 b_3 \dots b_{n-1}$; If $i < j$, in descending order of the priority sequences, change ($i-j$) 0 to 1s in $b_1 b_2 b_3 \dots b_{n-1}$. Then, $S_{nf} = [b_1; b_2; b_3 \dots b_{n-1}]$.

C. Capacitor Voltage Balance Analysis

To further explain the charging/discharging process of the proposed token rotation-based capacitor voltage balancing method, Fig. 10 shows an example of the flying capacitor charging/discharging of the seven-level inverter-based WPT system when $\delta_{\text{PMM}}^* = 0.7$. As shown in Fig. 10, the half-wave transmitter current is used for charge/discharge. The charging and discharging of the flying capacitor of the multilevel inverter is proportional to the integral of the positive half-wave transmitter current of the WPT system. The balancing algorithm execution frequency is twice the switching frequency. During a control period, the voltage variation of m_{th} flying capacitor can be expressed as follows:

$$\Delta u_{\text{cm}} = \frac{1}{C_{fm}} \int_0^{T_s/2} i_t(t) dt = \frac{I_t T_s}{\pi C_{fm}} \quad (22)$$

where I_t is the peak value of the transmitter current.

A heavier load and a larger δ_{PMM}^* can lead to a bigger transmitter current. Ignoring the influence of harmonic current, according to (16), δ_{PMM}^* is proportional to the transmitter current. Thus, the transmitter current can be further derived as follows:

$$I_t = \delta_{\text{PMM}}^* I_{tr} = \frac{\pi \delta_{\text{PMM}}^* P_r}{V_{\text{dc}}} \quad (23)$$

where I_{tr} and P_r are the rate transmitter current and input rate power of the WPT system, respectively.

According to (22) and (23), the voltage ripple of the flying capacitor can be derived as follows:

$$\Delta u_{mn} = N \Delta u_m = \frac{\delta_{\text{PMM}}^* N P_r T_s}{C_{fm} V_{\text{dc}}} \leq \frac{(n-2) P_r T_s}{C_{fm} V_{\text{dc}}} \quad (24)$$

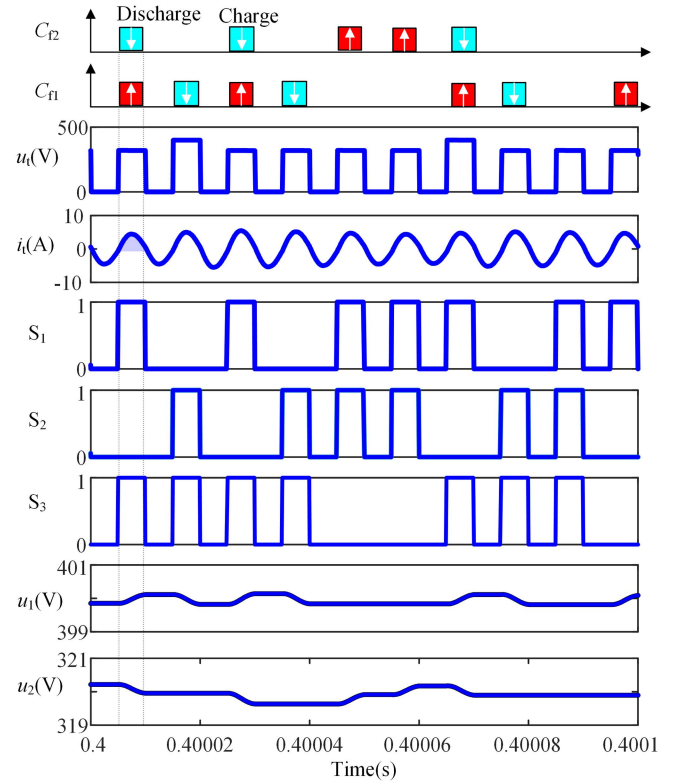


Fig. 10. Simulation results of the flying capacitor charging/discharging of the 7-level inverter when $\delta_{\text{PMM}}^* = 0.7$ and $C_{f1} = C_{f2} = C_{f3} = C_{f4} = C_{f5} = 55 \mu\text{F}$.

where $1 \leq N < n-2$.

As shown in (24), the voltage ripple of every flying capacitor is the same when the capacitor values are the same. The voltage ripple rate of the m_{th} flying capacitors can be defined as follows:

$$\varepsilon_m = \frac{\Delta u_{mn}}{V_c(m)} = \frac{(n-1) \Delta u_{mn}}{(n-1-m) V_{\text{dc}}}. \quad (25)$$

As shown in Fig. 10, the voltage ripple (Δu_1) in one control period is 0.36 V, which is consistent with (24). Based on (24) and (25), the $n-2$ capacitors can be selected as follows:

$$C_{fm} \geq \frac{2(n-2)(n-1) P_r T_s}{(n-m-1) \varepsilon_m V_{\text{dc}}^2}. \quad (26)$$

Furthermore, based on (26), it can be found that while all the flying capacitors are the same, the voltage ripple of m_{th} capacitor can be derived as follows:

$$\varepsilon_m = \frac{n-1-m}{n-2} \varepsilon_1. \quad (27)$$

Moreover, a higher switching frequency leads to a smaller capacitor value required.

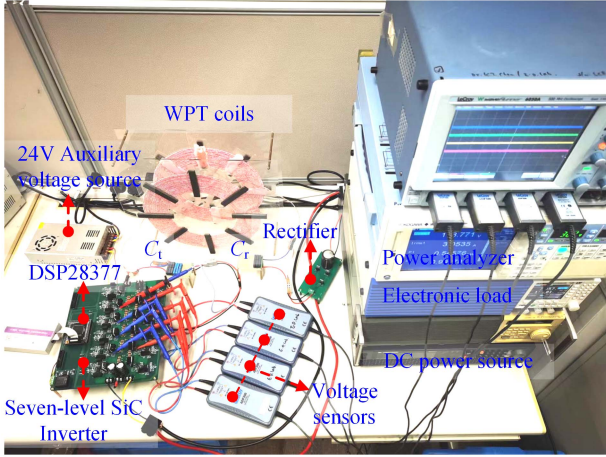


Fig. 11. Experimental platform.

IV. SIMULATION AND EXPERIMENTAL VERIFICATION

A. Hardware Implementation

Fig. 11 shows the 850 W prototype of the seven-level flying capacitor inverter-based WPT system, where the system consists of a dc power source (MR50040), a power analyzer (WT3000), a silicon-carbide-based seven-level flying-capacitor inverter with drivers, a WPT circuit, a bridge rectifier, electronic load (PLZ1003WH), a 24 V auxiliary voltage source, and a TMS320FDSP28377 board. The transmitter and receiver coils are copper Litz wire. The 24 V voltage source is used to feed the auxiliary circuit. The power analyzer with only one channel measures the input power. All flying capacitors have the same capacitance and rated voltage for convenience and lower voltage ripple, consisting of ten parallel MLCCs (450 V/2.2 μ F) and one electrolytic capacitor (450 V/33 μ F). However, it is noticed that the flying capacitors with different capacitance and rated voltage can be used to reduce the cost in the actual application. The system parameters are given in Table II.

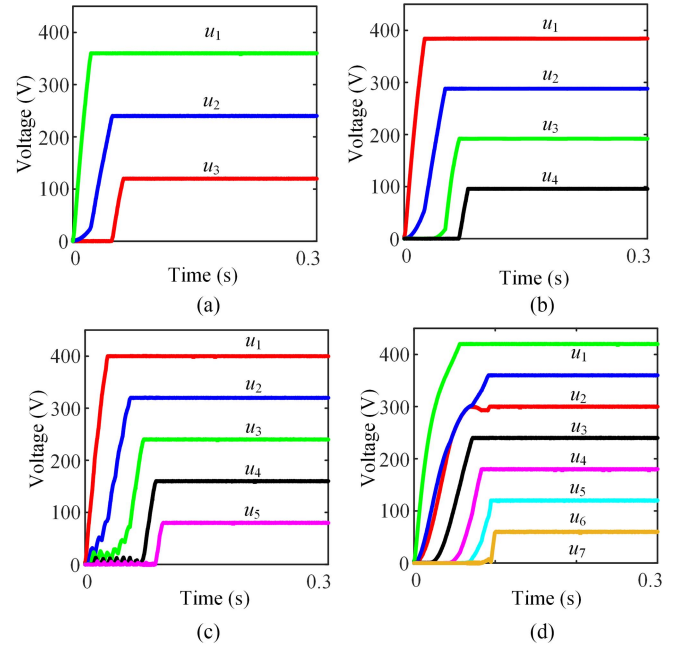
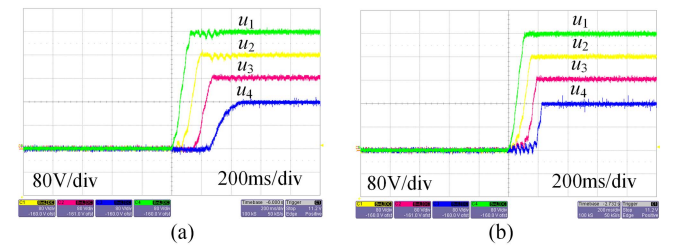
B. Flying-Capacitor Voltage Balance Analysis

Fig. 12 shows the simulation results of the capacitor voltages of the n -level inverter-based WPT system when $\delta_{PMM}^* = 0.7$. All the capacitor voltages can be balanced at the reference values at a quick rate, and the start-up times of all the capacitors are less than 0.3 s. Besides, comparing Fig. 12(a), (b), (c), and (d), it can be found that a bigger n leads to a longer start time of the capacitor voltage. Fig. 12 verifies that the proposed token rotation-based capacitor voltage balance method is effective for multilevel inverter-based WPT systems with arbitrary voltage levels.

Fig. 13 shows the experimental results of the capacitor voltage of the seven-level inverter for the WPT system with different δ_{PMM}^* . All the capacitor voltages can be balanced at the reference values within 0.42 s while $\delta_{PMM}^* = 0.2$, and within 0.24 s while $\delta_{PMM}^* = 0.8$. Comparing Fig. 13(a) and (d),

 TABLE II
SYSTEM PARAMETERS OF THE WPT SYSTEM

Items	Value
DC-link voltage (U_{dc})	480 V
Primary (Secondary) capacitance (C_i, C_r)	(8.7, 8.44) nF
Primary (Secondary) coil inductance (L_i, L_r)	(304.63, 300.15) μ H
Primary (Secondary) series resistance (R_i, R_r)	(0.3, 0.3) Ω
Coils diameter (d_1, d_2)	(30, 30) cm
Distance between two coils (d)	10 cm
Mutual inductance (M)	72.9625 μ H
Resonant frequency (f_r)	100 kHz
Output capacitance (C_o)	220 μ F
DC capacitors (C_{dc1}, C_{dc2})	(100, 100) μ F
Load resistance (R_{dc})	57.86 Ω
Dead time (T_d)	100 ns
Integration coefficient (K)	0.2
Digital signal processor	TMS320F28377
SiC MOSFET	SCT3080
Diode	CVFD20065A
Driver	UCC21710
Voltage sample	AMC1311DWVR


 Fig. 12. Simulation results of the capacitor voltages for n -level inverter-based WPT system when $\delta_{PMM}^* = 0.7$. (a) $n = 3$. (b) $n = 4$. (c) $n = 5$. (d) $n = 7$.

 Fig. 13. Experimental results of the flying-capacitor voltages for seven-level inverter-based WPT system (a) $\delta_{PMM}^* = 0.2$; (b) $\delta_{PMM}^* = 0.8$.

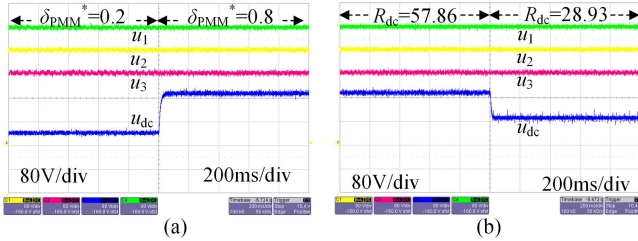


Fig. 14. Output voltage and capacitor voltages of the FCMLI-based WPT system. (a) Increasing δ_{PMM}^* . (b) Decreasing R_{dc} .

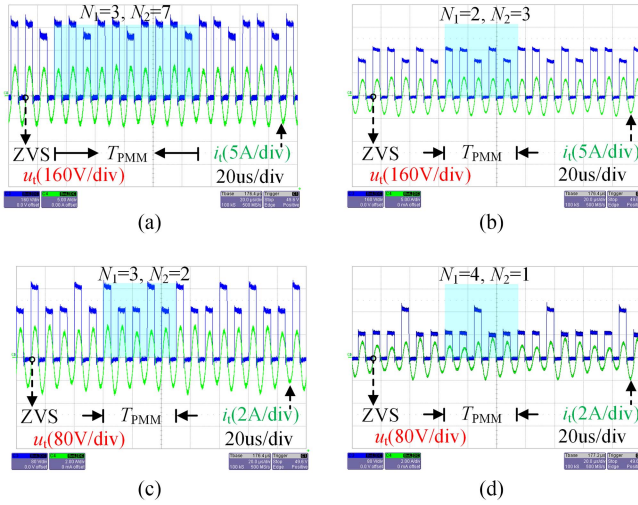


Fig. 15. Inverter output voltages and currents with different δ_{PMM}^* . (a) $\delta_{PMM}^* = 0.95$. (b) $\delta_{PMM}^* = 0.6$. (c) $\delta_{PMM}^* = 0.4$. (d) $\delta_{PMM}^* = 0.2$.

it can be found that a bigger δ_{PMM}^* leads to a shorter start time of the capacitor voltages. This is because a bigger δ_{PMM}^* leads to a bigger transmitter current. Fig. 13 further verifies the effectiveness of the proposed token rotation-based capacitor voltage balancing method of the multilevel inverter-based WPT systems.

Fig. 14 shows the flying capacitor voltages and output voltage of the WPT system with step responses. As shown in Fig. 14(a), while δ^* steps from 0.2 to 0.8, the output voltage steps from 44.1 to 176.0 V within 0.04 s, but the flying capacitor voltages (u_1 , u_2 , and u_3) have insignificant changes. Similarly, as shown in Fig. 14(b), while R_{dc} steps from 57.86 to 28.93 Ω , the output voltage steps from 176.0 to 88.0 V within 0.02 s, the flying capacitor voltages (u_1 , u_2 , and u_3) have insignificant changes. The experimental results in Fig. 14 show the stability of the proposed voltage balance method of the FCMLI-based WPT system under step perturbations.

C. ZVS and Transfer Efficiency Analyses

The proposed PMM FCMLI-based WPT system features ZVS. With different δ_{PMM}^* in the PMM, Fig. 15 shows the measured waveforms of the inverter output voltage (u_t) and the transmitter current (i_t). The system optimization is realized by slightly increasing the transmitter capacitor to enable the robust

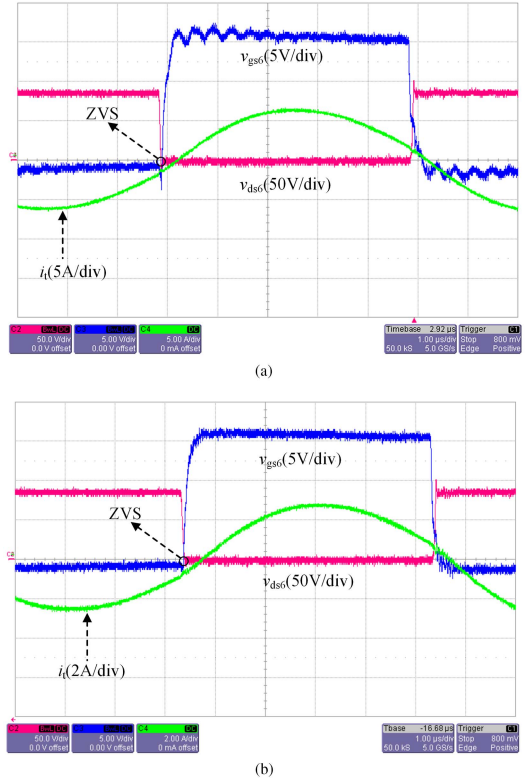


Fig. 16. V_{gs} and V_{ds} and of Q_6 and the transmitter current i_t with different δ_{PMM}^* . (a) $\delta_{PMM}^* = 0.95$. (b) $\delta_{PMM}^* = 0.4$.

ZVS reliably. Under the PMM with a variable δ_{PMM}^* , Fig. 15(a), (b), (c), and (d) shows that the PMM input voltage can energize the transmitter current slightly before the zero-crossing points with a robust ZVS during power control. Furthermore, Fig. 16 shows the gate-source voltage (v_{gs}) and drain-source voltage (v_{ds}) of Q_5 and Q_6 when δ_{PMM}^* equals 0.95 and 0.4, respectively. It can be observed that v_{gs5} and v_{gs6} become high-level voltage after v_{gs5} becomes zero, indicating that S_5 and S_6 can realize reliable ZVS. As the operating characteristics of the 12 switches are the same, all of them can realize the desired ZVS. Figs. 15 and 16 show that the proposed PMM can realize the wide-range output voltage regulation and ZVS simultaneously. It is noticed that the hard switching appears during the switch-off period, but the SiC MOSFETs turn OFF near the zero transmitter current. As a result, the power loss caused by the hard switching off is minimal.

Fig. 15 shows that the minimal cycle of u_t is equal to $10T_s$, $5T_s$, $5T_s$, and $5T_s$ when $\delta_{PMM}^* = 0.95$, 0.6, 0.4, and 0.2. It can be found that the modulated numbers of whole-cycles N_1 and N_2 are the minimum solution, which matches the calculated result. For example, in Fig. 15(a), while $\delta_{PMM}^* = 0.95$, δ_1 and δ_2 in (8) are $5/6$ and 1, respectively. Furthermore, according to (8), it can be obtained that $7N_1 = 3N_2$. Considering N_1 and N_2 are positive integers, that “ $N_1 = 3$ and $N_2 = 7$ ” are the minimum solution, which is consistent with the experimental result in Fig. 15(a).

Besides, Fig. 17 shows the relationship among the system efficiency, output power, output voltage, and δ_{PMM}^* , where

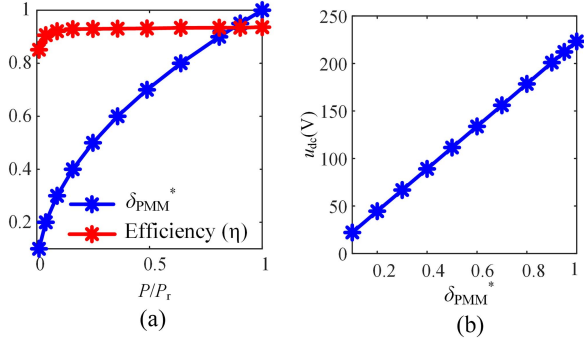


Fig. 17. The relation between the system efficiency, output power, and output voltage by changing δ_{PMM}^* . (a) The system efficiency and output power by changing δ_{PMM}^* . (b) The output voltage by changing δ_{PMM}^* .

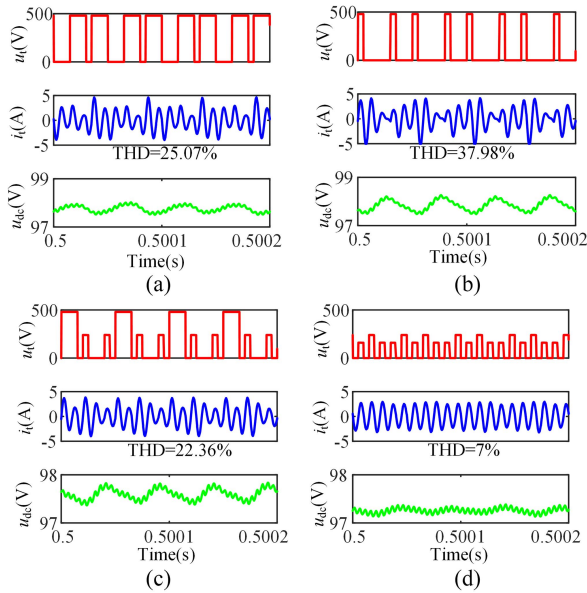


Fig. 18. Waveforms of PFM, PDM, SVPFM, and PMM for $\delta = 0.4$. (a) PFM [23]. (b) PDM. (c) SVPFM [26]. (d) the proposed PMM.

δ_{PMM}^* is selected as [0.1, 0.2, 0.3, 0.4, 0.5, 0.6, 0.7, 0.8, 0.9, 0.95, 1]. As shown in Fig. 17(a), the system efficiency is kept above 92% in a wide output power range (9%–100% P_r , and P_r is the rated power of the WPT system). Besides, Fig. 17(a) shows that the output power is proportional to the square of δ_{PMM}^* , and Fig. 17(b) indicates that the dc-side output voltage is proportional to δ_{PMM}^* . Due to the realization of the ZVS, the system efficiency is high during a wide-range output voltage and output power.

D. Comparison of the Proposed Method and Conventional Methods

Fig. 18 compares the waveforms of the proposed PMM and different modulation methods by using the half-bridge inverter. Both of them have the same circuit and parameters, except that Fig. 18(a) and (b) use the two-level inverter, Fig. 18(c) uses the

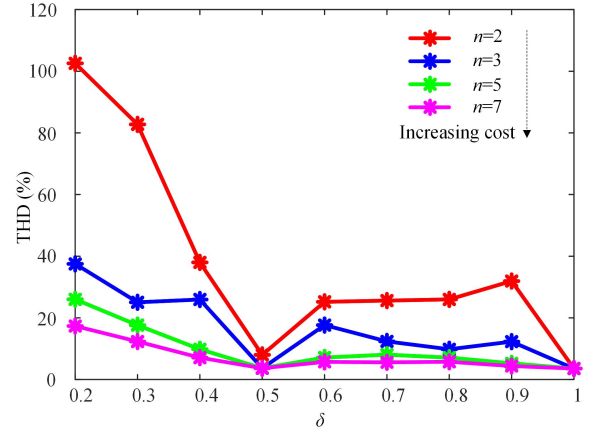


Fig. 19. THD of the transmitter current of the PMM n -level inverter-based WPT system.

three-level inverter, and Fig. 18(d) uses the seven-level inverter. It can be shown that the proposed PMM method has the lowest harmonic current on the transmitter side and the lowest dc-side voltage ripple, compared with SVPFM, PFM, and PDM.

Reducing THD in the WPT system can lead to increased reliability, compliance, performance, power quality, and lower power loss. The THD of the transmitter current of the WPT system is related to factors such as power level, the normalized voltage ratio (δ), modulation method, the compensated circuit, etc. Based on the MATLAB/Simulink, Fig. 19 further shows the THD of the transmitter current of the PMM n -level inverter-based WPT system under different normalized voltage ratios. It can be observed that a larger n leads to a smaller THD of the PMM multilevel inverter-based WPT system over a wide-range normalized voltage ratio. But when n increases by 1, two additional MOSFETs, an additional set of flying capacitors, and an additional set of voltage sampling circuits are required. The cost of the system increases linearly with an increase in n . As shown in Fig. 19, by comparison, the PMM seven-level inverter-based WPT system has the lowest THD and can be connected to the highest dc voltage bus, but it has the highest cost.

V. CONCLUSION

The PMM and token rotation-based capacitor voltage balance methods of the multilevel inverter for WPT systems have been proposed and implemented. Some conclusive remarks are summarized as follows.

- 1) This article proposes the Σ - Δ PMM method of multilevel inverters for WPT systems. The proposed PMM has the significant advantage of realizing easily and achieving wide-range voltage regulation and ZVS simultaneously.
- 2) Besides, the token rotation-based capacitor voltage balancing method of the fly-capacitor multilevel inverter for the WPT system has been proposed. The technique can make the flying-capacitor voltages of the multilevel inverter with arbitrary voltage levels balanced as the reference voltage.

- 3) The parameter design of the flying capacitor for the proposed PMM multilevel inverter-based WPT system has been given.

Finally, experimental verification is given to verify the feasibility of the proposed modulation method and token rotation-based capacitor voltage balancing method.

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