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Harmonic Analysis and Attenuation Strategy for a Two-Stage Matrix Converter Fed by Dual-Inverter Based on Pulse Barycenter Method

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Abstract: Since the DC-link of the dual-inverter two-stage matrix converter (TSMC) has no energy storage element, its grid-side current waveform is closely related to the load-side condition. The conventional modulation strategy of the dual-inverter TSMC is only the stack of the double inverters, without considering the cooperation between both modulation strategies. Thus, the fluctuations in the grid-side currents synthesized by both load-side currents will be obvious. To solve this problem, this paper analyzes the effect of the current pulse position on the grid-side current waveform quality, and relevant relationship between the pulse position and current harmonics is obtained according to Fourier transform. On this basis, the harmonic attenuation strategy for the dual-inverter TSMC is proposed. In the proposed strategy, the grid-side current harmonic is reduced based on the optimal duty-cycle combination, obtained from the harmonic optimization problem. The simulation and experimental results are provided to verify that the theoretical analysis of grid-side current harmonic content is consistent with the actual condition, and the proposed modulation strategy can effectively reduce the grid-side current harmonic.



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Keywords: two-stage matrix converter (TSMC); pulse width modulation; harmonic analysis; pulse barycenter distribution

1. Introduction

Two-stage matrix converters (TSMCs) have wide application prospects thanks to the advantages of good input and output performance, invertible energy transmission, compact structure, flexible and adjustable grid-side power factor, and so on. The topology of TSMCs can be divided into the rectifier stage and inverter stage. There is no energy storage element in TSMCs, compared with AC-DC-AC converters, which reduce the volume and weight, improve the reliability, and prolong the life of the system [1–3]. TSMCs have great superiority in the field of aerospace and military equipment, where the requirement of the converter volume is strict.

Since the DC-link has no energy storage element, the inverter modulation strategies will affect the current waveform in the inverter stage, and then have an effect on the rectifier stage through the DC-link, which means the grid-side currents will be influenced by the load directly. Aiming at the harmonic problem of TSMC, the method of Fourier transform analysis is usually utilized. Specifically, the mathematical model about load-side voltage is established by combining the pulse width modulation (PWM) principle with the transfer function firstly, then analytical expression of each harmonic component is obtained by use of mathematic derivation and Fourier transform [4–8]. For example, reference [9] constructs the constraint equation according to the conduction angle in each modulation period and obtains the formula of the amplitude of harmonics by use of Fourier transform. Reference [10] proposes a harmonic calculation method with triple Fourier series and applies this method to calculate the output voltage harmonic of the Direct Matrix Converter

(DMC) when the space vector PWM (SVPWM) strategy is used. Taking the “RMS current ripple” as the evaluating indicator, another common harmonic analysis method takes the integral of the errors between the expected current and actual current, applying this for inverter output waveform analysis [11,12]. For example, reference [12] proposes a current optimization strategy for the three-level inverter, in which the “RMS current ripple” is calculated under different modulation strategies, so the switching sequences according to the minimum “RMS current ripple” principle can be determined.

At present, the random PWM technology is usually adopted to reduce the modulation harmonics, i.e., for references [13,14], the grid-side current harmonics are suppressed by changing the carrier frequency or setting the distribution of the PWM voltage pulse within a certain range. Similarly, the SVPWM technology can also be improved to suppress the harmonics by correcting the duty cycles, such as adjusting the synthesizing sequence of voltage vectors, redistributing the duty cycles of active vectors, and so on [15,16]. Moreover, scholars also utilize different filters to suppress the grid-side current harmonics of MC, such as parallel active filter, hybrid filter, etc. [17–23].

As for the existence of the DC-link structure in TSMCs, the rectifier stage can connect two inverter stages; thus, TSMCs are suitable for operating conditions with a high requirement of power density, such as aircraft power systems, etc. [24,25]. When a DC-bus is linked with two parallel inverters, different modulation strategies need to be employed in double inverters. The currents of dual-inverters will be stacked and affect the grid-side current through the DC-link together, which makes the harmonic analysis of the grid-side current more complex. The harmonics of the grid-side current in dual-inverter TSMCs are related to the input frequency, the output frequency of double inverters, and the carrier frequency, etc. Therefore, the multiple-Fourier transform is needed to explain the coupling relationship in TSMCs, leading to a complex calculation process. In addition, the analysis method based on “RMS current ripple” needs to assume that the DC-bus voltage is constant, hence, it is not suitable for the harmonic analysis in topologies, such as TSMC. Besides, since the asymmetric modulation is normally utilized in TSMCs, there are nonlinear factors in this modulation method and it will lead to the distortion of the grid-side current waveform. Thus, both Fourier transform and the “RMS current ripple”-based method cannot be applied for harmonic analysis in dual-inverter TSMCs.

The method of Fourier transform occupies a dominant position in static harmonic analysis. Generally, fast Fourier transform (FFT) [26] or multiple-Fourier transform [27] are used to obtain the magnitude of fundamental and harmonic components of output voltage. However, the dynamic waveform quality evaluation methods have also been widely studied in recent years. For example, a dynamic harmonic evaluation method based on “effective value of output current fluctuation” is proposed in [28]. Firstly, the geometric relationship between the “effective value of output current fluctuation” and the triangle center of gravity of the output error current vector trajectory of the converter is established, and then the conclusion is derived that if the distance from the center of gravity to the origin is the shortest, the harmonic distortion rate of the converter is the smallest. Reference [29] proposes a dynamic harmonic evaluation method for a converter based on “effective value of stator flux fluctuation”. Firstly, the flux error vector is decomposed in the d-q coordinate system, and the d-axis component and q-axis component of the “stator flux fluctuation vector” are solved, respectively. Then, the two components are combined to obtain the “effective value of stator flux fluctuation” generated in the unit sampling period, which is used to dynamically evaluate the harmonics in the converter. However, the existing harmonic analysis methods are limited by the heavy online calculation and cannot achieve dynamic harmonic evaluation. The pulse barycenter method (PBM) comprehensively considers multiple factors influencing the current harmonics, such as the pulse width, amplitude, and distribution, etc. [30–33]. On this basis, PBM can be used to analyze the grid-side current harmonics of TSMCs according to the pulse barycenter movement [34–36]. Specifically, the grid-side harmonic-equivalent-value-based PBM contains the information on the pulse location and the relationship between the grid-side current harmonics and the

pulse barycenter can be effectively reflected by this value [36–39]. For a dual-inverter TSMC system, this paper employs PBM to analyze the effect of the current pulse distribution on the grid-side current harmonics in detail, then uses Fourier transformation to obtain the change rule of the pulse pattern with current harmonics; thus, a modulation strategy for dual-inverter TSMCs based on zero-vector duty-cycle correction (DCC) is proposed. The simulation and experimental results verify that the proposed modulation strategy according to PBM can effectively reduce the grid-side current harmonics and improve the quality of the grid-side current waveform.

2. Modulation Strategy of Dual-Inverter TSMCs

The topology of the double-inverter TSMC system is shown in Figure 1. From Figure 1, it can be seen that the DC-link has no energy storage element, and the DC-bus voltage is not the constant. Therefore, the operating of the rectifier stage needs to cooperate with the inverter stage, so as to implement the modulation of the whole system.

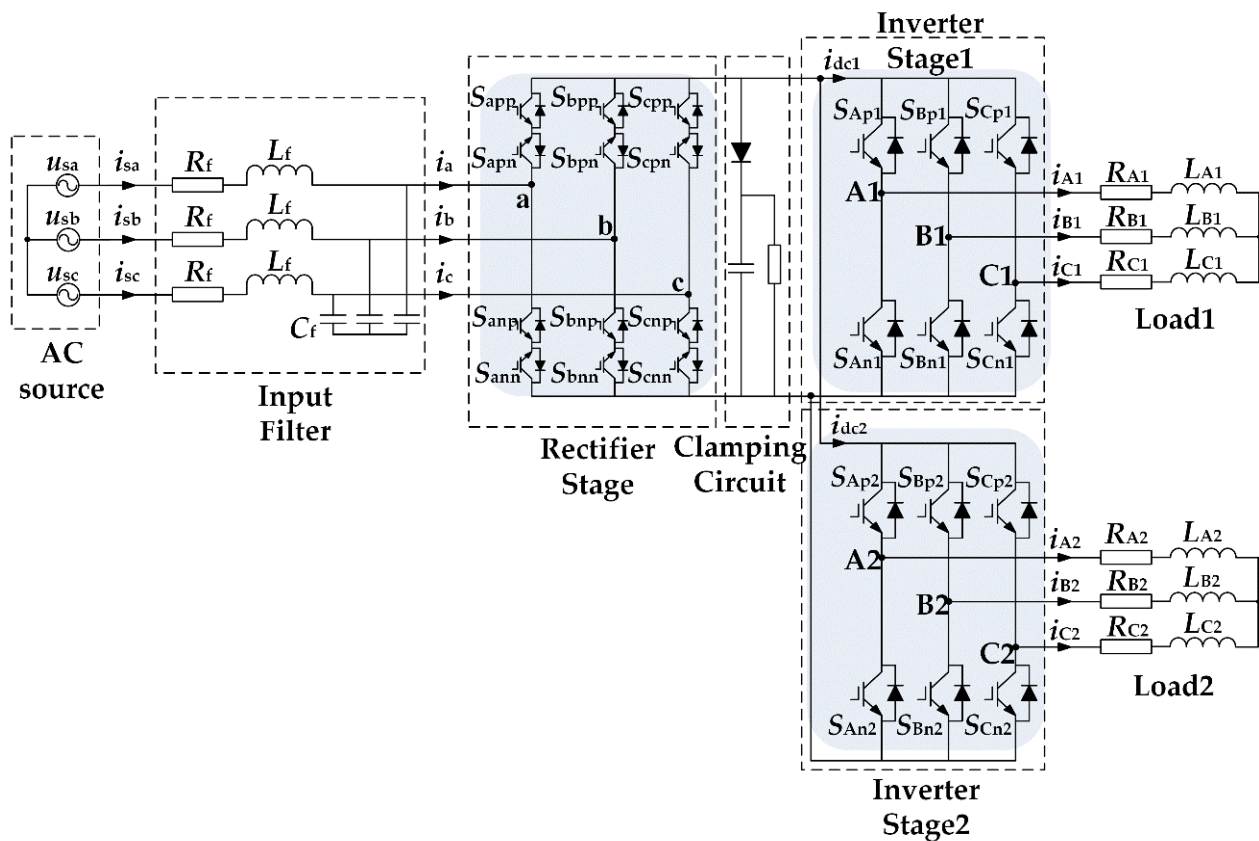


Figure 1. Topological structure diagram of dual-inverter TSMC.

2.1. The Modulation Strategy of Rectifier Stage

The grid-side voltage of the TSMC can be expressed as

$$\begin{cases} u_{sa}(t) = U_{im} \cos(\omega_r t) \\ u_{sb}(t) = U_{im} \cos(\omega_r t - \frac{2\pi}{3}) \\ u_{sc}(t) = U_{im} \cos(\omega_r t + \frac{2\pi}{3}) \end{cases} \quad (1)$$

where, U_{im} and ω_r represent the amplitude of the grid-side phase voltage and angular frequency, respectively.

There are nine switching states in the rectifier stage, and the corresponding grid-side current can be converted to the space vector form. Then, six active vectors and three zero vectors are obtained, where the six active vectors are marked as $i_1 \sim i_6$. The current

complex plane is divided into six sectors, as shown in Figure 2a. There are eight switching states in the inverter stage, and the corresponding output voltage can also be converted to the space vector form. Then, six active vectors and two zero vectors are obtained, where the six active vectors are marked as $u_1 \sim u_6$. The voltage complex plane is divided into six sectors, as shown in Figure 2a.

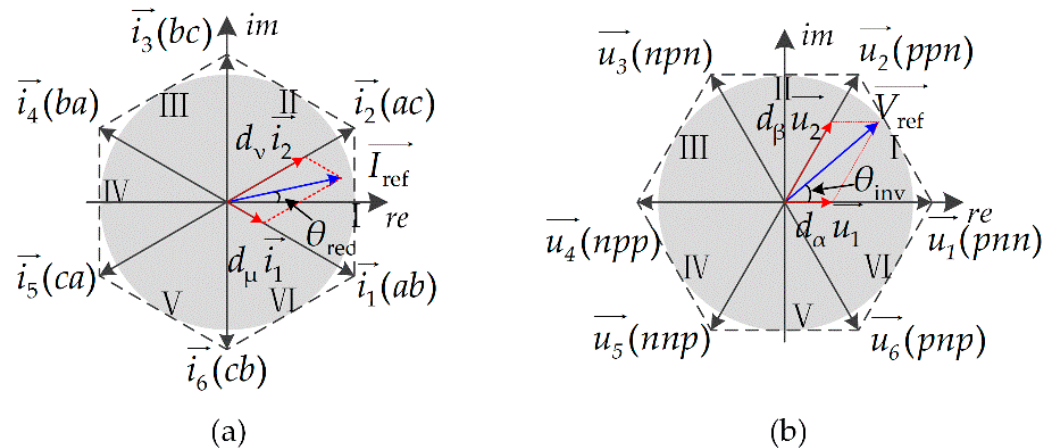


Figure 2. Rectifier and inverter space vector diagrams. (a) Rectifier; (b) inverter.

The grid-side reference current vector can be expressed as

$$\vec{I}_{\text{ref}} = \frac{2}{3}(i_a + i_b e^{j2\pi/3} + i_c e^{j4\pi/3}) = I_r e^{j\theta_{\text{rec}}} \quad (2)$$

where, I_r and θ_{rec} represent the amplitude of the grid-side reference current vector and sector angle, respectively.

For the rectifier stage, the prior applied active vector in each control period is set to ' $\vec{I}_{\mu}'$ ', and its duty cycle is d_{μ} ; the second applied active vector is set to ' $\vec{I}_{\nu}'$ ', and its duty cycle is d_{ν} . Taking sector I for example, the reference vector $\vec{I}_{\text{ref}}$ is synthesized by the active vector $\vec{i}_1$ and $\vec{i}_2$. The formula of the reference vector in each PWM period is

$$\vec{I}_{\text{ref}} = d_{\mu} \vec{i}_1 + d_{\nu} \vec{i}_2 \quad (3)$$

The modulation index of the rectifier is defined as

$$m_{\text{rec}} = \frac{I_r}{\overline{I_{dc}}} \quad (4)$$

where, $\overline{I_{dc}}$ is the average value of the DC-link current. In order to achieve the maximal voltage utilization ratio and reduce the times of commutation, zero vectors are generally not applied in the modulation of the rectifier, and the modulation index m_{rec} is kept constant. Therefore, the duty ratios of the two active current vectors $\vec{i}_1$ and $\vec{i}_2$ can be expressed as [40]

$$\begin{cases} d_{\mu} = \frac{-\cos(\theta_{\text{rec}} - 2\pi/3)}{\cos(\theta_{\text{rec}})} = \frac{-u_b}{u_a} \\ d_{\nu} = \frac{-\cos(\theta_{\text{rec}} + 2\pi/3)}{\cos(\theta_{\text{rec}})} = \frac{-u_c}{u_a} \end{cases} \quad (5)$$

2.2. The Modulation Strategy of Inverter Stage

From the above analysis, it can be seen that the output current vector in the rectifier stage is synthesized by two active vectors. Thus, the DC voltage in the rectifier stage during each modulation period will be separated into two parts and then the modulation for the inverter stage needs to be implemented twice successively under these two segments of

voltage. Meanwhile, in order to improve the safety and reliability of the system and reduce switching loss, the DC-link current is normally maintained at zero at the commutation instant in the rectifier stage, i.e., the voltage vectors of double inverters are zero.

The voltage vectors of the single inverter are shown in Figure 2b. The output voltage vector is set to $\vec{V}_{ref}$, which is synthesized by two adjacent active vectors in each sector. Taking sector I, for example, $\vec{V}_{ref}$ is assumed to be synthesized by $\vec{u}_1$ and $\vec{u}_2$. The duty ratios of $\vec{u}_1$ and $\vec{u}_2$ are set to d_α and d_β , and their formulas are

$$\begin{cases} d_\alpha = m_{inv} \sin(\pi/3 - \theta_{inv}) = m_{inv} \sin(\pi/3 - \omega_c t) \\ d_\beta = m_{inv} \sin(\theta_{inv}) = m_{inv} \sin(\omega_c t) \\ d_0 = 1 - d_\alpha - d_\beta \end{cases} \quad (6)$$

where, θ_{inv} is the sector angle of the output voltage vector; ω_c is the output frequency of the inverter.

The modulation index of the inverter is defined as

$$m_{inv} = \frac{U_{lm}}{u_{dc}} = \frac{2U_{lm} \cos(\theta_{in})}{3U_{lm}} \quad (7)$$

where, U_{lm} represents the amplitude of the load-side line voltage; θ_{in} is the angle of grid-side voltage.

The distribution of the duty cycles of the dual-inverter TSMC system is shown in Figure 3.

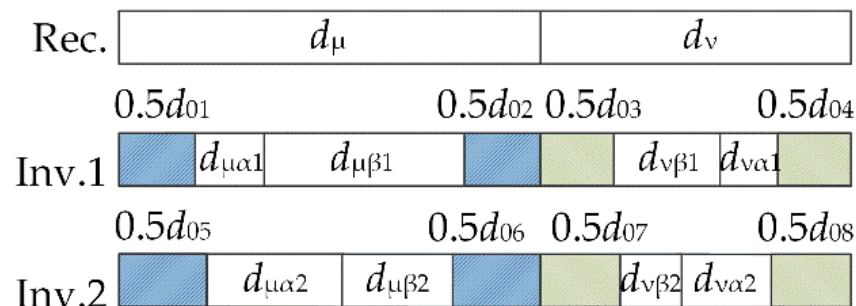


Figure 3. Duty-cycle distribution of dual-inverter TSMC system.

According to the distribution pattern shown in Figure 3, the duty cycles of the voltage vectors for the dual inverter are determined according to the duty cycles in the rectifier stage. In other words, the proportion of the inverter duty cycle is d_μ when $\vec{I}_\mu$ is applied; the proportion of the inverter duty cycle is d_v when $\vec{I}_v$ is applied.

The duty cycles for the active vector $\vec{U}_{\mu\alpha1}$, $\vec{U}_{\mu\beta1}$, $\vec{U}_{v\alpha1}$, and $\vec{U}_{v\beta1}$ in inverter 1 are set to $d_{\mu\alpha1}$, $d_{\mu\beta1}$, $d_{v\alpha1}$, and $d_{v\beta1}$, respectively; the duty cycles for the zero vector $\vec{U}_{01}$, $\vec{U}_{02}$, $\vec{U}_{03}$, and $\vec{U}_{04}$ are set to d_{01} , d_{02} , d_{03} , and d_{04} , respectively. They are expressed as [41]

$$\begin{cases} d_{\mu\alpha1} = d_\mu d_{\alpha1} \\ d_{\mu\beta1} = d_\mu d_{\beta1} \\ d_{01} + d_{02} = d_\mu - d_{\mu\alpha1} - d_{\mu\beta1} \\ d_{v\alpha1} = d_v d_{\alpha1} \\ d_{v\beta1} = d_v d_{\beta1} \\ d_{03} + d_{04} = d_v - d_{v\alpha1} - d_{v\beta1} \end{cases} \quad (8)$$

The duty cycles for the active vector $\vec{U}_{\mu\alpha 2}$, $\vec{U}_{\mu\beta 2}$, $\vec{U}_{v\alpha 2}$, and $\vec{U}_{v\beta 2}$ in inverter 2 are set to $d_{\mu\alpha 2}$, $d_{\mu\beta 2}$, $d_{v\alpha 2}$, and $d_{v\beta 2}$, respectively; the duty cycles for the zero vector $\vec{U}_{05}$, $\vec{U}_{06}$, $\vec{U}_{07}$, and $\vec{U}_{08}$ are set to d_{05} , d_{06} , d_{07} , and d_{08} , respectively. They are expressed as

$$\begin{cases} d_{\mu\alpha 2} = d_{\mu}d_{\alpha 2} \\ d_{\mu\beta 2} = d_{\mu}d_{\beta 2} \\ d_{05} + d_{06} = d_{\mu} - d_{\mu\alpha 2} - d_{\mu\beta 2} \\ d_{v\alpha 2} = d_{v}d_{\alpha 2} \\ d_{v\beta 2} = d_{v}d_{\beta 2} \\ d_{07} + d_{08} = d_{v} - d_{v\alpha 2} - d_{v\beta 2} \end{cases} \quad (9)$$

Since the inverters in the system are not coupled, the modulation strategies of inverters are independent. Hence, the duty cycles for inverters can be calculated independently based on (8) and (9), and multiple combination patterns for the modulations can be constructed.

3. Analysis of the Grid-Side Current Harmonics

3.1. The Relationship between Current Pulse Barycenter and Load-Side Current

The current pulse barycenter (PB) of the single inverter could represent the distribution of the active vector pulse during each modulation period. The current waveform of Inverter 1 in each modulation period is shown in Figure 4.

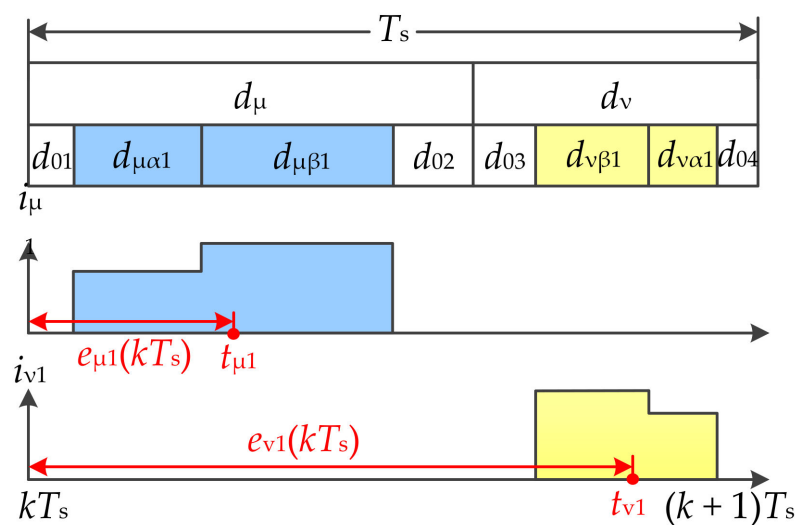


Figure 4. Current waveform of Inverter 1.

In Figure 4, $i_{\mu 1}$ is the current of Inverter 1 when $\vec{I}_{\mu}$ is applied; $i_{v 1}$ is the current of Inverter 1 when $\vec{I}_{v}$ is applied. kT_s and $(k+1)T_s$ are the beginning and ending instant of the k th period, respectively.

The time difference between the instant $t_{\mu 1}$ and the beginning instant of the modulation period kT_s is defined as current PB $e_{\mu 1}(kT_s)$, where the instant of the geometric barycenter of $i_{\mu 1}$ is achieved. $e_{\mu 1}(kT_s)$ can be calculated by the duty cycles and control period T_s .

$i(t)$ is set to the average value of the inverter current in each period

$$i(t)|_{t=kT_s} = i(kT_s) = \frac{1}{T_s} \int_{kT_s}^{(k+1)T_s} i_{\mu 1}(t) dt \quad (10)$$

According to the area equivalence principle, the current $i_{\mu 1}$ of Inverter 1 in the k th period can be approximately replaced by the ideal pulse with the amplitude $T_s i(kT_s)$ and delay time $kT_s + e_{\mu 1}(kT_s)$.

$$i_{\mu 1}(t) \approx \sum_{k=1}^{\infty} T_s i(kT_s) \delta(t - (kT_s + e_{\mu 1}(kT_s))) \quad (11)$$

where, $i(kT_s)$ is the average value of the inverter current during the k th period.

Laplace transform is used to transform both sides of (11), and the formula of the current $i_{\mu 1}(t)$ in the complex frequency domain is

$$i_{\mu 1}(s) \approx \sum_{k=1}^{\infty} [i(kT_s) T_s \cdot e^{-s(kT_s + e_{\mu 1}(kT_s))}] \quad (12)$$

The infinite sum formula on the right side of Formula (12) can be further converted into integral form, then the frequency-domain expression of $i_{\mu 1}(t)$ can be obtained as

$$i_{\mu 1}(s) \approx \int_0^{\infty} i(t) \cdot e^{-s(t + e_{\mu 1}(t))} dt \quad (13)$$

Define the variable τ as the function of t

$$\tau = f(t) = t + e_{\mu 1}(t) \quad (14)$$

Substituting (14) into (13), the result is

$$i_{\mu 1}(s) \approx \int_0^{\infty} \frac{d(f^{-1}(\tau))}{d\tau} \cdot i(f^{-1}(\tau)) \cdot e^{-s\tau} d\tau \quad (15)$$

Formula (15) is the definition formula in the form of Laplace transform. Neglecting the high-frequency components, the current $i_{\mu 1}(t)$ in time domain is

$$i_{\mu 1}(t) = \frac{d(f^{-1}(\tau))}{d\tau} \cdot i(f^{-1}(\tau)) \approx i(t) - \frac{d(e_{\mu 1}(t) \cdot i(t))}{dt} \quad (16)$$

From (16), it is known that the harmonic characteristics of $i_{\mu 1}(t)$ and $i(t)$ are similar, when $e_{\mu 1}(t)$ is constant; the harmonics will generate in the differential block of (16), when $e_{\mu 1}(t)$ is not constant. In other words, the PB of $e_{\mu 1}(t)$ leads to the generation of harmonics in the current.

3.2. The Relationship between Current Pulse Barycenter and Grid-Side Current

Since the DC-bus of the dual-inverter TSMCs has no energy storage element, the grid-side current pulse can be synthesized by the composition of two inverter current pulses. Moreover, the change rule of the grid-side current PB is complex for the asymmetric modulation, so the duty cycles might be calculated firstly to obtain the PB of double inverters, and then the change rule of the grid-side current PB can be concluded. To sum up, (16) can be used to analyze the effect of the modulation strategy of inverters on the grid-side current harmonics.

Assuming that the reference vector in the rectifier stage is located in Section 2, the reference vector of Inverter 1 is located in Section 2, and the reference vector of Inverter 2 is located in Section 3, the distribution of vectors, grid-side current, and inverter currents are shown in Figure 5.

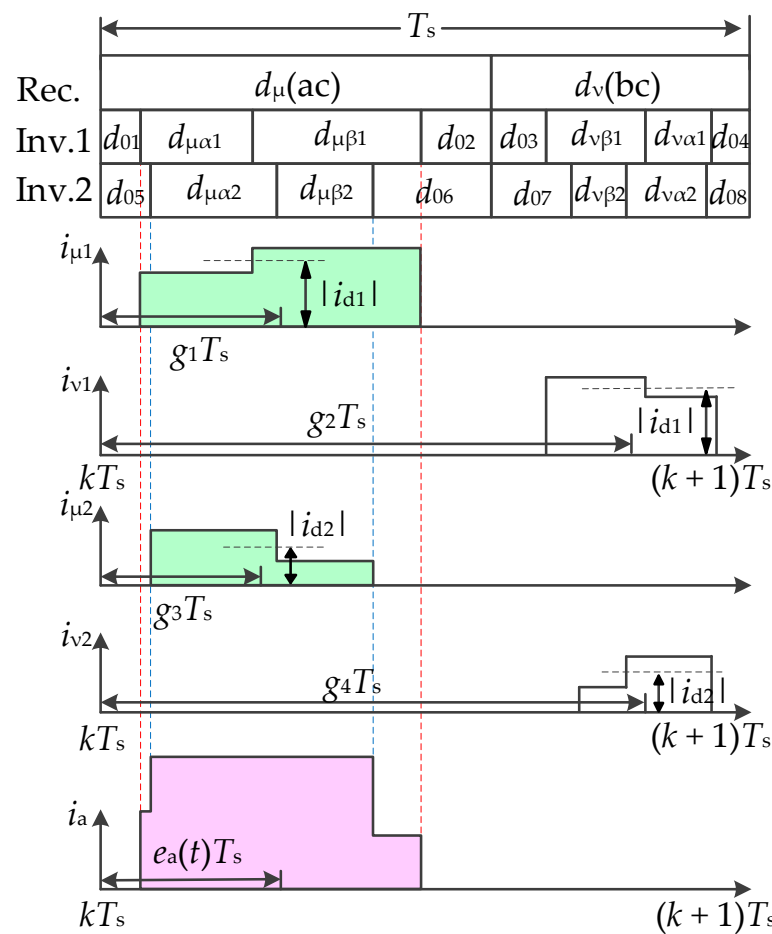


Figure 5. Waveforms of vector distribution, grid-side current, and inverters' currents.

From Figure 2, it can be seen that the output line voltages in the rectifier are u_{ac} and u_{bc} . The DC-bus voltage is u_{ac} when $\vec{i}_2$ is applied; the DC-bus voltage is u_{bc} when $\vec{i}_3$ is applied. Moreover, the a-phase of the rectifier stage is turned on when u_{ac} is applied, then i_a will exist only when u_{ac} is applied (the active vector $\vec{i}_2$ is applied). At this moment, i_a is the sum of the current i_{u1} of Inverter 1 and i_{u2} of Inverter 2.

The distribution coefficient of the zero vector in Inverter 1 is defined as x ; the distribution coefficient of the zero vector in Inverter 2 is defined as y , where x and y are the constant value between 0 and 1. The distribution of the zero vector will change with the variation in x and y , and the barycenter of the active vector pulse will also change accordingly.

In Figure 5, the zero vector of Inverter 1 in each modulation period can be expressed as

$$\begin{cases} d_{01} = x(d_\mu - d_{\mu\alpha1} - d_{\mu\beta1}) \\ d_{02} = (1-x)(d_\mu - d_{\mu\alpha1} - d_{\mu\beta1}) \\ d_{03} = (1-x)(d_v - d_{v\alpha1} - d_{v\beta1}) \\ d_{04} = x(d_v - d_{v\alpha1} - d_{v\beta1}) \end{cases} \quad (17)$$

Similarly, the zero vector of Inverter 2 in each control period can be expressed as

$$\begin{cases} d_{05} = y(d_\mu - d_{\mu\alpha2} - d_{\mu\beta2}) \\ d_{06} = (1-y)(d_\mu - d_{\mu\alpha2} - d_{\mu\beta2}) \\ d_{07} = (1-y)(d_v - d_{v\alpha2} - d_{v\beta2}) \\ d_{08} = y(d_v - d_{v\alpha2} - d_{v\beta2}) \end{cases} \quad (18)$$

From the definition of the current PB in Section 3.1, it is known that the PB of the current $i_{\mu 1}$ and $i_{\mu 2}$ of the inverter can be expressed as

$$\begin{cases} g_1 T_s = [x(d_\mu - d_{\mu\alpha 1} - d_{\mu\beta 1}) + \frac{1}{2} d_\mu (d_{\alpha 1} + d_{\beta 1})] T_s \\ g_3 T_s = [(y(d_\mu - d_{\mu\alpha 2} - d_{\mu\beta 2}) + \frac{1}{2} d_\mu (d_{\alpha 2} + d_{\beta 2}))] T_s \end{cases} \quad (19)$$

As the pulse waveform of i_a is the sum of $i_{\mu 1}$ and $i_{\mu 2}$, the PB of i_a will also be determined by $i_{\mu 1}$ and $i_{\mu 2}$. Considering the amplitude of $i_{\mu 1}$ and $i_{\mu 2}$, the PB of i_a can be expressed as

$$e_a(t) = \frac{g_1 T_s \cdot i_{\mu 1} + g_3 T_s \cdot i_{\mu 2}}{i_{\mu 1} + i_{\mu 2}} \quad (20)$$

The PB of i_a in different sectors are different, and can be expressed as

$$e_a(t) = \begin{cases} (\frac{e_1 d_\mu + e_2 d_\nu}{d_\mu + d_\nu}) T_s & N_r = 1, 4 \\ (e_1 = \frac{g_1 T_s \cdot i_{\mu 1} + g_3 T_s \cdot i_{\mu 2}}{i_{\mu 1} + i_{\mu 2}}) T_s & N_r = 2, 5 \\ (e_2 = \frac{g_2 T_s \cdot i_{\nu 1} + g_4 T_s \cdot i_{\nu 2}}{i_{\nu 1} + i_{\nu 2}}) T_s & N_r = 3, 6 \end{cases} \quad (21)$$

where, N_r is the number of the sectors in the rectifier stage.

From (21), it is known that the PB of the grid-side current changes periodically, and it is related to the duty cycles in the rectifier stage and the inverter stage. Because of the duty cycles are the function of the input frequency ω_r , output frequency ω_{c1} , and ω_{c2} , the harmonics related to the input frequency and output frequency will exist in the grid-side current due to the variation in the current PB.

4. Modulation Strategy for Dual-Inverter TSMCs Based on Duty-Cycle Correction

From the analysis in Section 3.2, it is known that the current PB of the inverter changes with the variation in the value of the distribution coefficients x and y , and thereby, the current harmonic is affected. Therefore, the current harmonics can be suppressed by changing x and y . In order to calculate the grid-side current harmonic content under different values of x and y , the grid-side current harmonic content can be quantitatively calculated by combining with Fourier transformation with (15). The values of x and y , minimizing the grid-side current harmonic content, are obtained through the above procedure, and then the duty cycles of zero vectors for two inverters are corrected.

The Fourier coefficient represents the amplitude function of a certain frequency, and the harmonic content in the grid-side current can be calculated accordingly. The principle of the Fourier transformation is that waveform $f(t)$ can be expressed by the form of infinite series, consisting of the sinusoidal harmonics with different orders, i.e.,

$$f(t) = \frac{a_0}{2} + \sum_{m=1}^{\infty} (a_m \sin m\omega t + b_m \cos m\omega t) \quad (22)$$

The amplitude of each harmonic a_m can be calculated as

$$a_m = \frac{4}{T} \int_0^{T/2} [i_a(t) \sin(m\omega_r t)] dt \quad (23)$$

By use of the transfer function method, the analysis result concluded that the function of the grid-side current is an odd function, so the input current $i_a(t)$ can be expressed as

$$i_a(t) = \sum_{m=1}^{\infty} a_m \sin(m\omega_r t) \quad (24)$$

According to the analysis results in Section 3.2 and (16), the approximate expression of the grid-side current is obtained.

$$i_a(t) \approx i(t) - \frac{d(e_a(t) \cdot i(t))}{dt} \quad (25)$$

The average value in each period can be calculated in real time according to the grid-side current pulse waveform. In actual calculation, the ideal grid-side current can be expressed as

$$i(t) = I_{im} \sin(\omega_r t) \quad (26)$$

Substituting (21), (25), and (26) into (23), the amplitude of the m -th harmonic that changes with x and y can be obtained

$$a_m = \frac{4}{T} \int_0^{T/2} i_a^*(t) \sin(m\omega_r t) dt - \frac{4}{T} \int_0^{T/6} \left[\frac{d}{dt} \left(\frac{e_1 d_u + e_2 d_v}{d_u + d_v} T_s i_a^*(t) \right) \sin(m\omega_r t) \right] dt - \frac{4}{T} \int_{T/6}^{T/3} \left[\frac{d}{dt} (e_1 T_s i_a^*(t)) \sin(m\omega_r t) \right] dt - \frac{4}{T} \int_{T/3}^{T/2} \left[\frac{d}{dt} (e_2 T_s i_a^*(t)) \sin(m\omega_r t) \right] dt \quad (27)$$

From (27), it can be known that the amplitude of the grid-side current harmonic is related to the current PB, and the current PB is related to the input frequency, output frequency, and the distribution coefficients x and y . Hence, the grid-side current harmonic content can be reduced by changing the values of x and y , and then the waveform quality of the grid-side current can be improved.

In order to analyze the changing rule of the grid-side current harmonic content expressed by (27), the following parameters listed in Table 1 are employed.

Table 1. Power, filter, and load parameters.

System Parameters	Values
Grid-side voltage u_s	60 V
Input frequency ω_{in}	50 Hz
Filter capacitor C_f	17 μ F
Filter inductance L_f	0.8 mH
Filter resistance R_f	2.2 Ω
Load resistance R	20 Ω
Load inductance L	18 mH
Output frequency ω_{out}	40 Hz

Considering that the rectifier stage operates under zero-current commutation, the values of x and y vary from 0.2 to 0.8. For ease of calculation, only the percent of the 3rd, 5th, 7th, 9th, 11th, and 13th harmonics are calculated, and the harmonic orders higher than the 15th are neglected. The results of (27) are obtained by use of Maple17, and the results are shown in graphic form, as shown in Figure 6a.

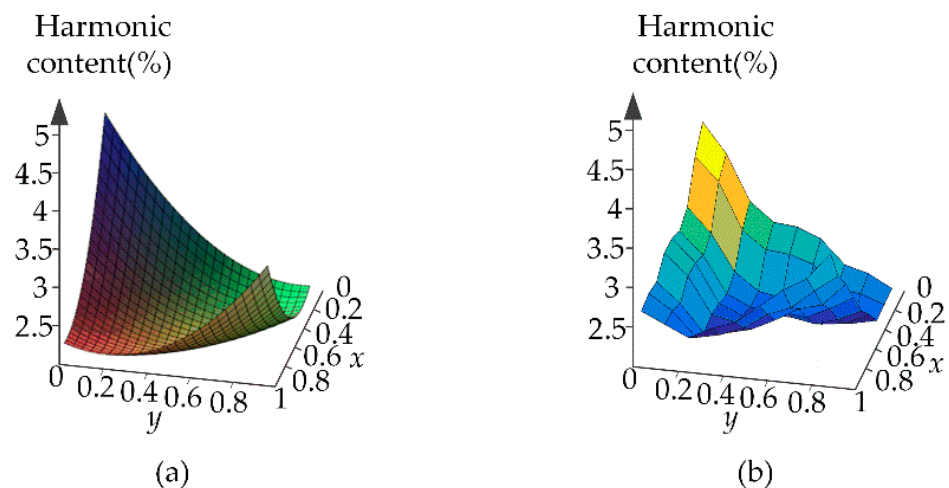


Figure 6. The amplitude of low-frequency harmonic content under different x and y . (a) Equation (27) graphical (b) simulation.

From Figure 6a, it can be seen that the harmonic content changes as the law of paraboloid. Based on this law, the distribution of the PWM pulse for the double inverters can be optimized, and the aim of decreasing the grid-side current harmonic content can be achieved.

In order to verify the results of the theoretical calculation, the values of x and y are selected and the a-phase grid-side current is analyzed by use of fast Fourier analysis. The harmonic content is shown in Figure 6b, and it can be seen that the change trends in Figure 6b are similar to Figure 6a.

In order to quantitatively compare the theoretical result in Figure 6a with the simulation result in Figure 6b, the difference between the theoretical and simulation result is shown in Table 2.

Table 2. Comparison of theoretical calculation results and simulation results for harmonic content.

x								
y	Error/%	0.2	0.3	0.4	0.5	0.6	0.7	0.8
	0.2	8.75	1.3	0.10	6.86	9.40	0.86	1.17
	0.3	0.98	10.4	12.2	8.02	6.67	12.84	8.57
	0.4	0.10	12.2	11.5	8.11	5.14	8.02	6.94
	0.5	6.86	8.01	8.11	1.42	2.31	5.96	1.82
	0.6	9.40	6.67	5.14	2.31	7.49	3.12	6.91
	0.7	0.86	12.8	8.02	5.96	3.13	6.91	3.89
	0.8	1.17	8.57	6.94	1.82	4.96	10.9	6.60

From Table 2, it can be seen that the simulation results are slightly larger than the theoretical results; this is because the high harmonics are neglected in the theoretical calculation, so its harmonic content in the theoretical result is smaller. Moreover, only finite points are used in the simulation, which leads to the discontinuity of the duty cycles. Thus, it is different from the ideal mathematical model obtained by theoretical calculation, so the graphic of the simulation results is not smooth.

From Figure 6, it can be seen that the harmonic content is smaller when x is 0.2 and y is 0.8 (or x is 0.8 and y is 0.2). Therefore, the value of x is set to 0.2 and y is set to 0.8 (or x is set to 0.8 and y is set to 0.2). Substituting the values of x and y into (17) and (18), the optimized duty cycles of zero vectors $xd_{0Inv,1}$ and $yd_{0Inv,2}$ are obtained. The corrected distributions of PWM duty ratios of double inverters are shown in Figure 7.

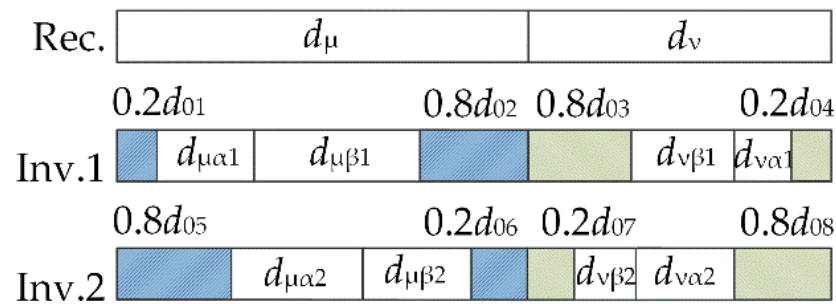


Figure 7. Duty-cycle distribution of dual-inverter TSMC system after duty-cycle correction.

From Figure 7, it can be seen that the width of the pulse is not changed, even if the distribution of the active vector pulse changes. In other words, both the duty-cycles of the active vector and the amplitude of load-side voltage do not change.

The block diagram of the duty-cycle correction strategy is shown in Figure 8. First, the reference voltage of the inverter is obtained, and the angles θ_1^* and θ_2^* of the reference voltage are obtained by PLL. The duty cycles for the inverter $d_{\alpha\text{Inv.1}}$, $d_{\beta\text{Inv.1}}$, $d_{\alpha\text{Inv.2}}$, and $d_{\beta\text{Inv.2}}$, and the duty cycles of zero vectors $d_{0\text{Inv.1}}$ and $d_{0\text{Inv.2}}$ can be calculated according to the angle of reference voltage and (7). After obtaining the duty cycles, the current harmonic contents under different x and y are compared by use of (27) and the optimal pulse is obtained. According to (17) and (18) and the pulse distribution coefficient, the corrected duty cycles of zero vectors can be obtained. Finally, the turn-on times of IGBTs in different sectors are calculated according to the corrected duty cycles and (8) and (9), and the application times $T_{\text{Ap1}} \sim T_{\text{Cn2}}$ are obtained. Comparing the effect time of each IGBT with the triangular carrier, the PWM pulses in Inverter 1 and Inverter 2 can be obtained by logic judgment.

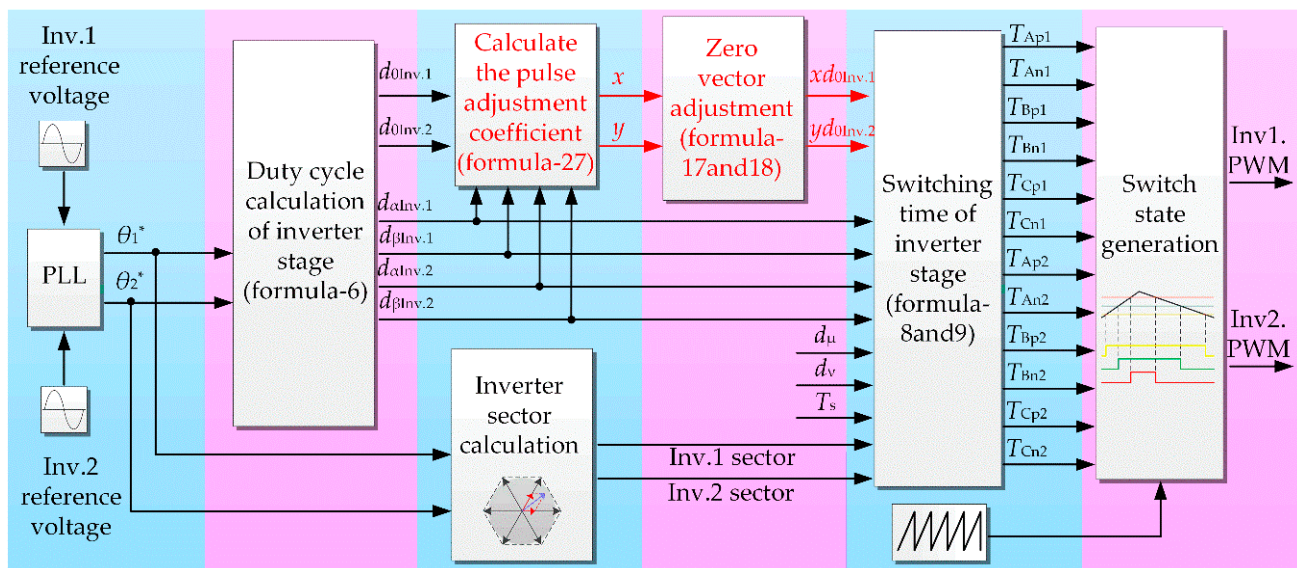


Figure 8. Block diagram of duty-cycle correction.

5. Simulation Results

In order to verify the theoretical results, MATLAB/Simulink was used to establish the simulation model of the dual-system TSMC. The parameters of the system are shown in Table 1.

Employing the modulation strategy without zero vector for the rectifier stage, the waveforms of a-phase voltage u_{sa} and current i_{sa} obtained in simulation are shown in Figure 9, respectively. The output frequency is set to 40 Hz. The load-side line voltage u_{AB}

and phase current i_A are shown in Figure 10. From these figures, it can be seen that the load-side line voltage is an ideal PWM waveform and output line current is near sinusoidal.

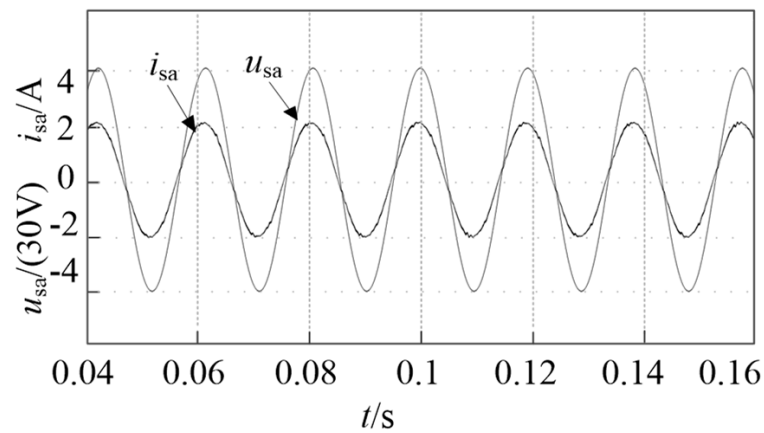


Figure 9. Simulation waveform of grid-side phase voltage u_{sa} and phase current i_{sa} .

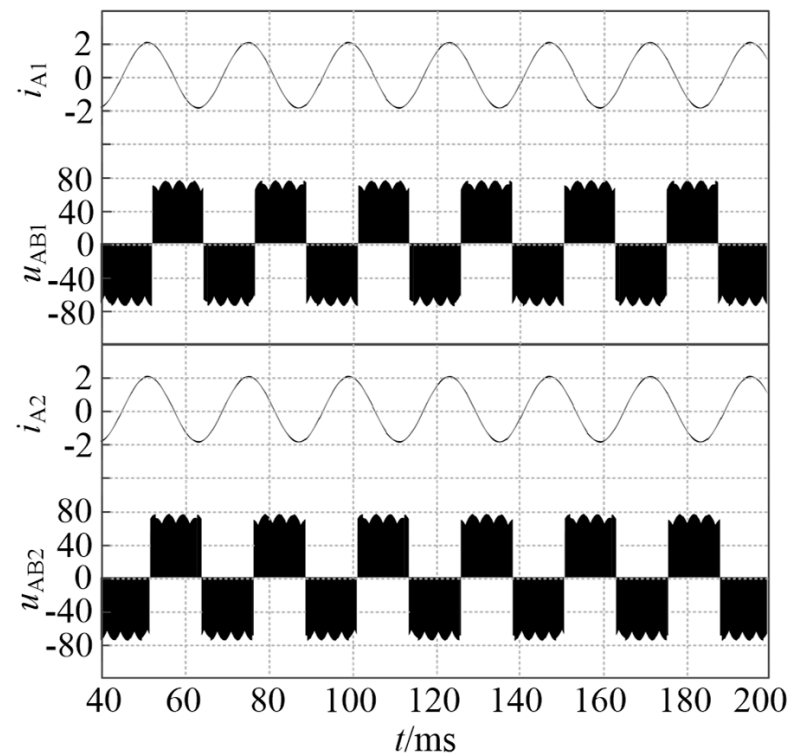


Figure 10. Simulation waveforms of load-side line voltage u_{AB} and current i_A .

Figures 11 and 12 show the simulation waveforms and Fourier analysis results of the grid-side current when the distribution coefficients $x = 0.5$, $y = 0.5$ and $x = 0.2$, $y = 0.8$, respectively. In Figure 11, i_{sa1} is the fundamental waveform of the grid-side current; i_{san} is the n -th harmonic of the grid-side current. From Figure 11, it can be seen that the grid-side current harmonic content decreases with the variation in the values of x and y .

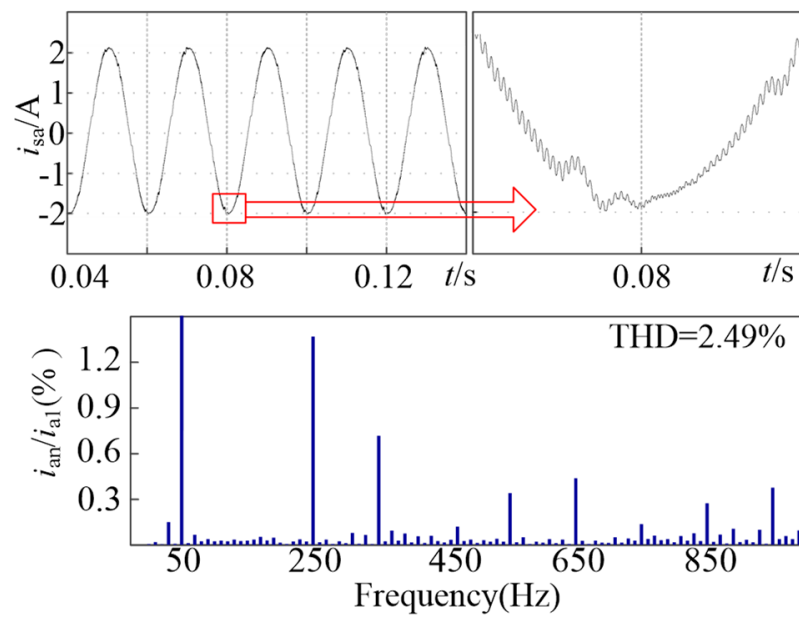


Figure 11. Simulation waveform and Fourier analysis of grid-side current in conventional modulation strategy.

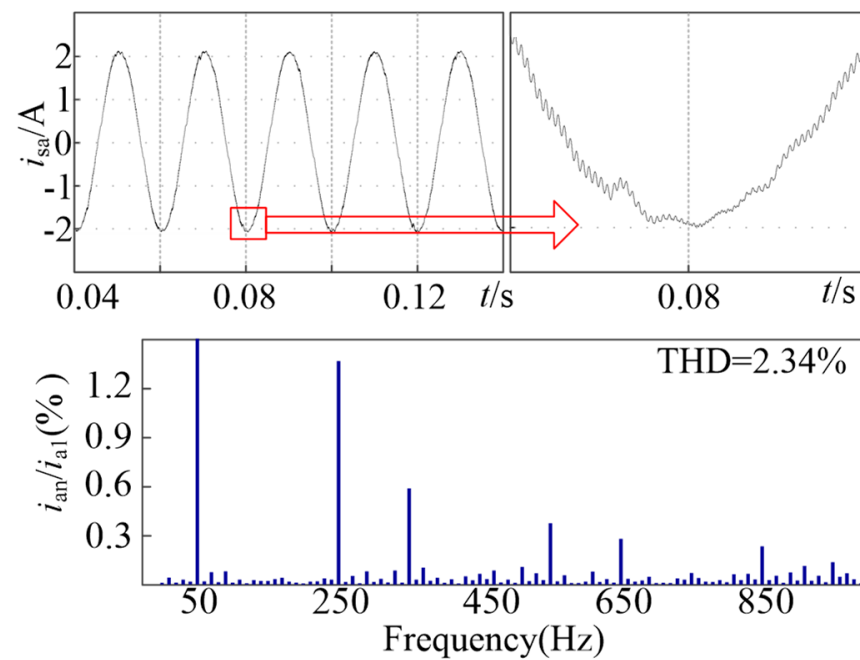


Figure 12. Simulation waveform and Fourier analysis of the grid-side current.

6. Analysis of Experimental Results

The experiments were carried out on the platform (shown in Figure 13), with the parameters shown in Table 1. The sampling frequency is set to 10 kHz; the spectrum analysis of the grid-side current is executed to verify the distribution law of the harmonics.



Figure 13. Figure of Experimental platform.

Figures 14 and 15 show the experimental waveforms of the grid-side phase voltage u_{sa} and current i_{sa} and load-side line voltage u_{AB} and current i_A , when the input frequency is 50 Hz and $x = 0.5$, $y = 0.5$. Figure 14 shows the experimental waveforms of the load-side line voltage u_{AB} and current i_A of double inverters when the input frequency is 40 Hz. From Figure 14, it can be seen that the load-side line voltages of double inverters are an ideal PWM waveform and the output line current is near sinusoidal.

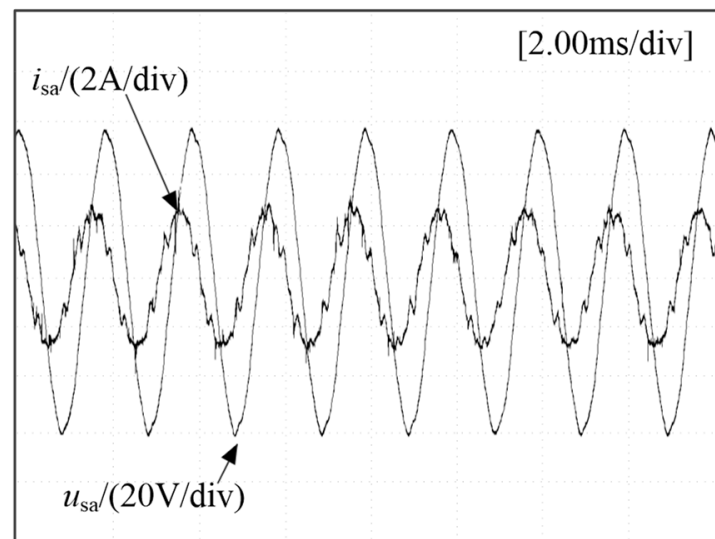
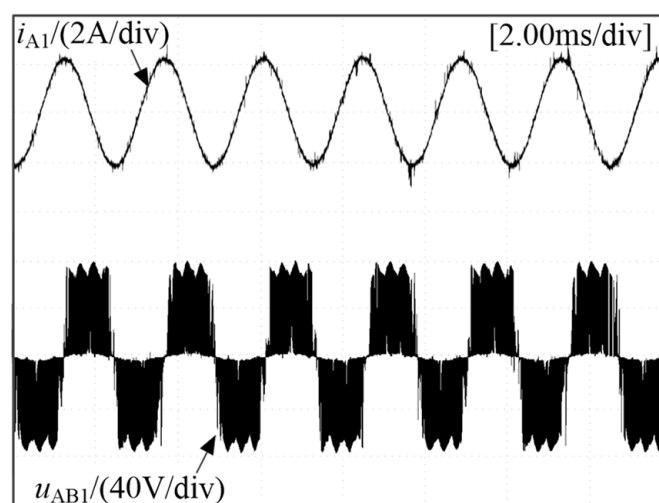
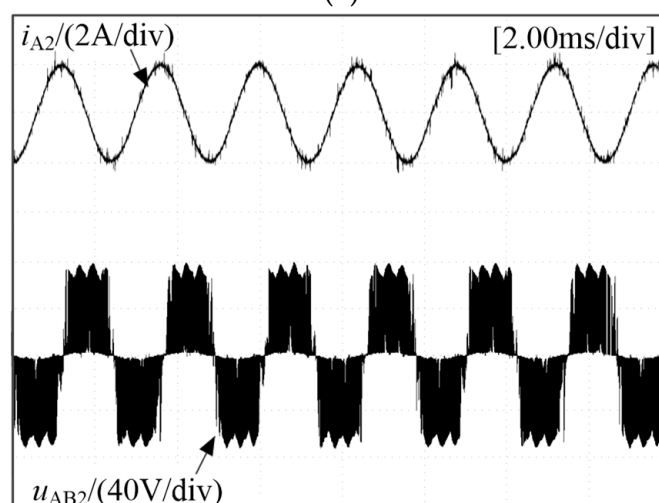


Figure 14. Experimental waveforms of grid-side phase voltage u_{sa} and phase current i_{sa} .

Figures 16 and 17 show the experimental waveforms of the grid-side phase voltage u_{sa} and current i_{sa} and load-side line voltage u_{AB} and current i_A when the input frequency is 50 Hz and $x = 0.2$, $y = 0.8$. Figure 16 shows the experimental waveforms of the load-side line voltage u_{AB} and current i_A of double inverters when the input frequency is 40 Hz.



(a)



(b)

Figure 15. Experimental waveforms of load-side line voltage u_{AB} and current i_A . (a) Inverter 1; (b) Inverter 2.

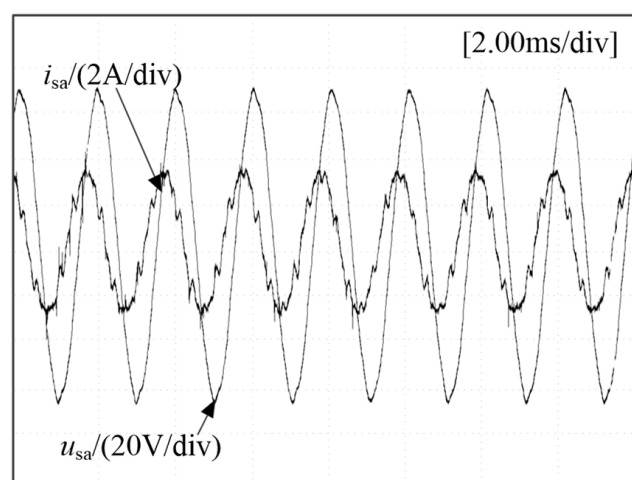


Figure 16. Experimental waveforms of grid-side phase voltage u_{sa} and current i_{sa} after duty-cycle correction.

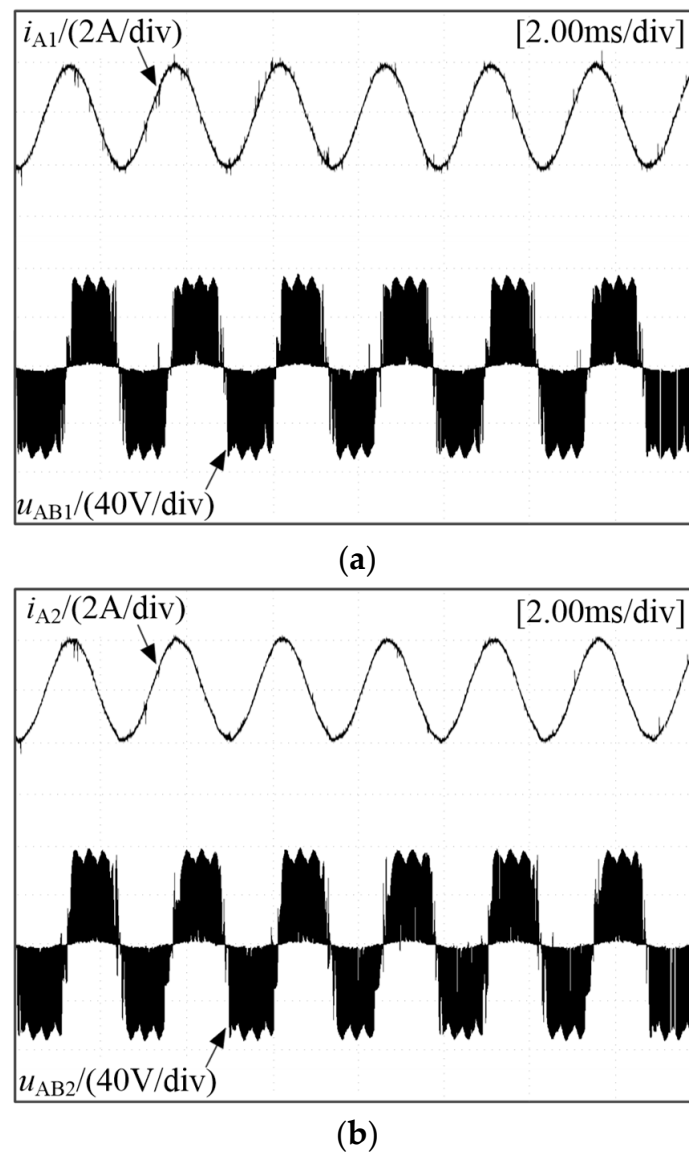


Figure 17. Experimental waveforms of load-side line voltage u_{AB} and current i_A of after duty-cycle correction. (a) Inverter 1; (b) Inverter 2.

Figure 18 shows the experimental waveforms and the Fourier analysis result of the grid-side phase current i_{sa} when $x = 0.5$, $y = 0.5$ and $x = 0.2$, $y = 0.8$. It can be seen that the harmonics of the grid-side current are suppressed by use of the duty-cycle correction strategy, and it is consistent with the theoretical analysis.

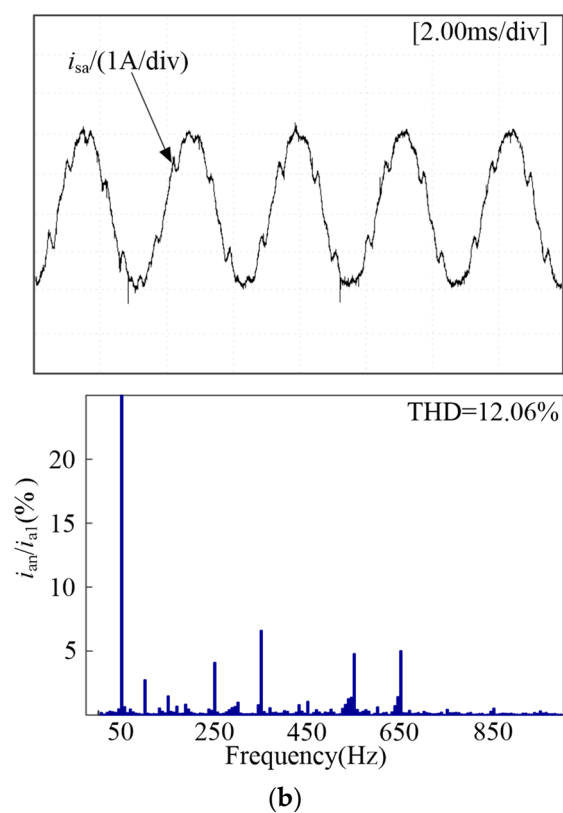
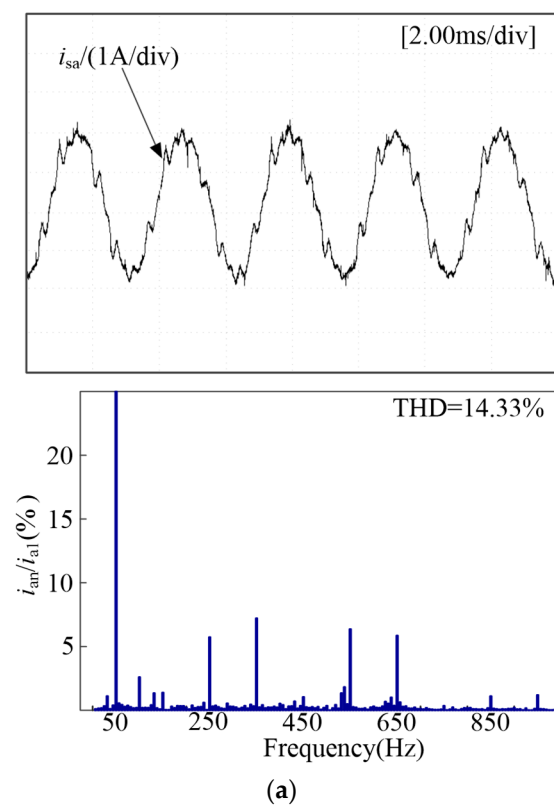


Figure 18. Experimental waveforms and Fourier analysis results of the grid-side current. (a) Conventional modulation strategy; (b) duty-cycle correction strategy.

7. Conclusions

This paper analyzed the harmonic characteristics of the grid-side current of a dual-inverter TSMC system according to PBM, and analyzed the law of the effect from the pulse distribution on the harmonic content of the grid-side current. Further, a modulation strategy to reduce the current harmonics was proposed. The following results were proved by the simulation and experiments.

(1) The grid-side current harmonics in the dual-inverter TSMC are affected by the pulse width and distribution of inverters, and the harmonic content changes with the variation in the distribution coefficients of the pulses in the form of paraboloid.

(2) By optimizing the pulse width and distribution of inverters, the harmonic content of the grid-side currents can be effectively reduced. The grid-side current harmonic content can be effectively reduced when the distribution coefficient x is set to a value near 0.2 (0.8) and y is set to a value near 0.8 (0.2).

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