A T-type Switched-Capacitor Multilevel Inverter with Low Voltage Stress and Self-Balancing

Yaoqiang Wang, Member, IEEE, Yisen Yuan, Gen Li, Member, IEEE, Yuanmao Ye, Member, IEEE, Kewen Wang, Member, IEEE, and Jun Liang, Senior Member, IEEE

Abstract—This paper proposes a novel T-type multilevel inverter (MLI) based on the switched-capacitor technique. The proposed inverter not only achieves that the maximum voltage stress of the switches is less than the input voltage but also has a voltage boost capability, which makes it suitable in high voltage applications. It is worth mentioning that the proposed inverter features two topology extension schemes which help it achieve a higher output level and voltage gain. With the merit of low voltage stress and reduced power devices, a seven-level inverter can be achieved using only two capacitors. Moreover, capacitor voltage self-balancing capability can simplify the complexity of the circuit and control. The topology, operating principle, modulation strategy and analysis of the capacitor of the inverter are presented. The superiorities of the proposed inverter are investigated by comparing with recently proposed hybrid MLIs and switched-capacitor MLIs. Finally, a seven-level prototype is constructed to validate the correctness of the theoretical analysis and the feasibility and effectiveness of the proposed inverter.

Index Terms—Multilevel inverter, switched-capacitor, low voltage stress, self-balancing, extension.

NOMENCLATURE

$A_c$ Amplitude of the triangular carriers
$f_c$ Frequency of the triangular carriers
$A_{ref}$ Amplitude of the sinusoidal modulation wave
$f_o$ Frequency of the sinusoidal modulation wave
$V_o$ Output voltage
$V_{dc}$ Voltage of dc source
$V_s$ Voltage stress of the switch
$C_i$ Capacitor number $i$
$C_s$ Parasitic capacitance of the switch
$\Delta V_{C2}$ Voltage ripple of $C_2$
$V_{C2}$ The voltage of $C_2$
$\Delta Q_1$ Discharge amount of $C_2$ during the period of $0-t_1$
$\Delta Q_2$ Discharge amount of $C_2$ during the period of $t_2-t_3$
$\Delta Q_{C2}$ The maximum discharge amount of $C_2$
$Q_o$ The discharge amount of the capacitor when it works at $V_{dc}/2$

$Q_m$ The maximum discharge amount of the capacitor
$K$ The factor of the maximum acceptable voltage ripple
$P_{rip}$ Ripple losses
$P_{con}$ Conduction losses
$P_{sw}$ Switching losses
$P_o$ Output power
$N$ Output levels
$x$ Number of the connected inverters
$n$ Number of the extended H-bridges
$y$ Number of the switches
$N_s$ Switching transitions in one period
$M$ Fundamental wave modulation index
$R_s$ Load resistance
$ESR_C$ Equivalent series resistance of each capacitor
$r_s$ Equivalent resistance of the switch
$f_s$ Switching frequency of the switches
$I_o$ Output current
$r_{eq}$ Equivalent parasitic resistance
$\eta$ Efficiency of the nine-level inverter
$t_s$ Working time of the switches in one period
$T_s$ Time of one cycle
$t_i$ Intersection time instant of the $i$th sine and triangle waves
$m$ Step number of half-cycle
TSV Total standing voltage
THD Total harmonic distortion

I. INTRODUCTION

Multilevel inverters (MLIs) play an important role in power electronics systems thanks to their excellent features in high power capacity, low switching voltage stress, low harmonics, modular and scalable design [1]-[5].

The development of MLIs can be traced back to the late 1960s [6]. Traditional MLIs can be divided into three types: neutral point clamp (NPC), flying capacitor (FC) and cascade H-bridge (CHB). These inverters have a wide range of applications due to the advantages of low voltage stress, low switching frequency and capability for high voltage applications [7]-[9].

1) Motivation and incitement

Based on the three types of inverters, numerous new inverters have been designed and proposed to enhance their performance in the past decades. To achieve a higher output level and improve extensibility, the topology combines the NPC and FC is one of the attractive alternatives to [10]-[12]. The CHB inverter has also been modified to reduce the number of switching devices [13]. These improvements enrich the variety

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of MLIs. They have a common advantage of limiting the voltage stress of all semiconductor devices to the dc input voltage. However, these inverters do not have voltage boosting capability. In addition, the capacitor voltage balance is still a challenge for NPC inverters [14]. The existence of flying capacitors requires additional complex control circuits [15]. CHB circuits require multiple independent power sources [16]. These shortcomings limit their widespread applications. The inverters in [17] and [18] use different switch combinations to connect multiple dc power sources in series to supply the load and output multiple voltage levels. Different dc voltage ratios can be achieved, which increases their flexibility. However, the MVS of their switches is at least the voltage of two dc sources. Moreover, they do not have voltage boosting capability and have only one topology extension scheme.

The switched-capacitor technique provides a good way to solve the above-mentioned shortcomings. There is no additional complicated control circuit in the switched-capacitor multilevel inverters (SCMLIs) [19]. They are able to boost small input dc voltages to high ac voltages by switching the pre-charged capacitors in series [20]-[21]. Such inverters also benefit from their small size, light weight, large power density and low harmonic components.

2) Literature review

Various SCMLIs have been proposed in the open literature. The SCMLIs presented in [22]-[25] can maintain the balance of capacitor voltages. In addition, they are featured that only one dc power source is employed with the capability of voltage boosting. However, using H-bridges to achieve the voltage polarity conversion will limit their application because the H-bridge increases the maximum voltage stress (MVS) of the switches and also, increases the capital cost.

The H-bridge has been removed in [26]-[29] without affecting the voltage polarity conversion. However, there are still switches need to withstand the peak value of the output voltage, which may also limit their applications in high voltage systems. The topology proposed in [30] reduces the MVS of switches, however, the MVS of switches is still twice the dc source voltage. The topologies proposed in [31]-[32] can effectively reduce the voltage stress of the switches. However, a large number of switches and independent power sources are required by the ones in [31] and [32], respectively. In [33]-[36], the proposed inverters limit the MVS of switches to the input power voltage utilizing proper devices. However, more improvement can be made to use fewer capacitors in the topologies proposed in [33]-[35].

3) Contribution and paper organization.

To overcome the shortcomings of the above-mentioned inverters in terms of using numerous devices, high MVS of switches, and lacking voltage boosting and capacitor voltage self-balancing, an MLI based on the switched-capacitor technique has been proposed in this paper. Comparing with conventional MLIs, the proposed inverter can boost the input power voltage and realize capacitor voltage self-balancing. Moreover, the MVS of inverter switches is limited to less than the input power voltage. The components used in the proposed MLI are reduced compared to low voltage stress SCMLIs. In addition, the proposed inverter is also superior in its flexible extensibility and capability of supplying inductive loads. The low voltage stress is one of the excellent features that made the proposed inverter an attractive alternative for medium- and high-voltage photovoltaic power generation applications.

The rest of the paper is organized as: the circuit configuration and operation principles of the proposed inverter and analysis of capacitors are described in Section II. Section III gives an analysis of power losses and efficiency. The topology extension and comparison analysis are presented in Section IV. In Section V, the feasibility and effectiveness of the proposed MLI have been validated through experiment. Section VI draws the conclusion and closes the paper.

II. PROPOSED MULTILEVEL INVERTER

A. Circuit Configuration

The topologies proposed in [33]-[36] are shown in Fig. 1. The common feature of them is that they only use one dc power source, two T-type voltage-dividing capacitors to achieve a seven-level inverter. The two T-type capacitors are used to generate a voltage level of $V_{dc}/2$, making the step voltage of the output waveforms have a smaller value than $V_{dc}$, which can effectively reduce the total harmonic distortion (THD) of the output voltage. However, the two T-type capacitors are not fully utilized because they are only used for voltage division and cannot be connected in series with the dc power source to supply the load. Therefore, these topologies require additional capacitors to achieve multilevel outputs.

![Fig. 1. Topologies proposed in [33]-[36]. (a) Topology in [33]; (b) Topology in [34]; (c) Topology in [35]; (d) Topology in [36].](image)

In order to reduce the number of capacitors and thereby maximize the utilization of the two T-type capacitors, a new T-type switched-capacitor inverter has been proposed in this paper.
paper, as shown in Fig. 2. In this topology, the two capacitors are not only employed to generate the voltage level of $V_{dc}/2$, but also act as the switched capacitors, which eliminates the need for additional capacitors. In addition, the proposed topology can be regarded as a reconstruction of the topology proposed in [35]. The dc source and two T-type voltage-dividing capacitors are integrated to reduce the additional switched capacitor. Comparing to the topology in [35], the front-to-front connected switches $S_3$ and $S_6$ are needed to achieve a bidirectional current path. However, the proposed topology, in turn, reduces one capacitor at the expense of adding one switch. It is worth mentioning that the topology proposed in [35] has the potential to further reduce the number of capacitors. For instance, the two T-type voltage-dividing capacitors could be replaced by two split dc sources. In this case, only one capacitor is needed to achieve a seven-level output. Similar changes can be made in [33], [34] and [36]. However, the use of multiple dc sources may limit the applications of such designs. This is because that one important driving force of using the switched capacitor technique for multilevel inverters is to reduce the use of independent dc sources.

In the proposed inverter, the dc power source (e.g. photovoltaic panels, batteries, and fuel cells) supplies the capacitors and load. The switched-capacitor structure guarantees the boost capacity and enables a voltage boost gain of 1.5. With the help of the H-bridge, the two capacitors can operate alternately to achieve the voltage polarity conversion on the load. In this case, the H-bridge used for changing the voltage polarity in conventional SCMLIs can be removed. This feature leads to a remarkable decrease of the MVS of switches. The proposed inverter can a seven-level output: $\pm 3V_{dc}/2$, $\pm V_{dc}$, $\pm V_{dc}/2$ and 0.

**B. Operating Principle**

This section describes the operating principle of the seven-level inverter. Different output levels are obtained by controlling the on and off states of each switch. All operating states under different modes are shown in Table I. It is noted that 0 and 1 refer to the off and on states of the related switches. The states of the capacitors are shown by “C”, “D” and “–”, which indicate the charging, discharging and rest states.

<table>
<thead>
<tr>
<th>$V_n$</th>
<th>Switches</th>
<th>Capacitors</th>
</tr>
</thead>
<tbody>
<tr>
<td>$3V_{dc}/2$</td>
<td>0110000100</td>
<td>-</td>
</tr>
<tr>
<td>$V_{dc}$</td>
<td>0110000011</td>
<td>-</td>
</tr>
<tr>
<td>$V_{dc}/2$</td>
<td>0101100100</td>
<td>C</td>
</tr>
<tr>
<td>0</td>
<td>0101000011</td>
<td>-</td>
</tr>
<tr>
<td>$-V_{dc}/2$</td>
<td>1001100000</td>
<td>D</td>
</tr>
<tr>
<td>$-V_{dc}$</td>
<td>1001000011</td>
<td>-</td>
</tr>
<tr>
<td>$-3V_{dc}/2$</td>
<td>1001001000</td>
<td>D</td>
</tr>
</tbody>
</table>

The current paths in each operating mode are shown in Fig. 3. It is noted here, the two operating modes with the output voltage of $\pm V_{dc}/2$ contain two current paths. One path is that the capacitor supplies the load. The other is that the power source charges the capacitors in series. It can be seen from Fig. 3 that the load current can flow reversely under each operating mode and the reversed current path is the same as the forward current path. It proves that the proposed inverter has the capacity of supplying inductive loads.

**C. Modulation Strategy**

The pulse width modulation (PWM) is mainly divided into three types: carrier wave PWM, selective harmonic eliminated PWM (SHE-PWM) and the space-vector PWM (SV-PWM) [37]-[38]. The SHE-PWM can reduce the switching frequency, thereby reducing the switching losses, and improving the utilization of the dc voltage. However, it is complicated to implement. The SV-PWM method is suitable for inverters which output three to five voltage levels. However, it is not suitable for inverters output more than five voltage levels due to its complexity. In this paper, the phase disposition PWM (PD-PWM), which is one type of carrier wave PWMs, is
adopted to generate the driving signals for all switches. The advantage of this method is that it is easy to implement, thereby greatly reducing the complexity of the control circuit. For a seven-level inverter, six triangular carriers and a sinusoidal modulation wave are required to generate pulses. Different logical combinations of these pulses control the on and off states of each switch. The schematic diagram of the PD-PWM is shown in Fig. 4. The six triangular carriers have the same amplitude ($A_c$) and frequency ($f_o$) but different offsets. The amplitude of the sinusoidal modulation wave is $A_M$ and the frequency is $f_c$. The modulation index is determined by the amplitude of the carrier and reference waveforms. Hence, the modulation index $M$ can be defined as

$$M = \frac{A_M}{3A_c}. \quad (1)$$

The proposed inverter is able to adjust its output accordingly when $M$ changes. The relationship between different values of $M$ and the output level is shown in Table II.

<table>
<thead>
<tr>
<th>$M$</th>
<th>Output level</th>
</tr>
</thead>
<tbody>
<tr>
<td>$0 &lt; M \leq 1/3$</td>
<td>3</td>
</tr>
<tr>
<td>$1/3 &lt; M \leq 2/3$</td>
<td>5</td>
</tr>
<tr>
<td>$2/3 &lt; M \leq 1$</td>
<td>7</td>
</tr>
</tbody>
</table>

**D. Analysis of Capacitor Balancing**

As shown in Table I, $C_1$ is charged when the output voltage is $V_{dc}/2$ and discharged when the output voltage is $-V_{dc}/2$ and $3V_{dc}/2$. $C_2$ is charged when the output voltage is $-V_{dc}/2$ and discharged when the output voltage is $V_{dc}/2$ and $3V_{dc}/2$. It can be seen from Fig. 4, the working states of the two capacitors are symmetrical in one cycle. $C_2$ works in the positive half-cycle, then the voltage will drop and the voltage of $C_1$ rises. This is because the two capacitors are connected in series with the dc power source when they are charged. $C_1$ works in the negative half-cycle, then the voltage will drop and the voltage of $C_2$ rises. The voltages of two capacitors can return to the initial state after a cycle deviating from the set voltage. Hence, the voltages of the two capacitors are self-balanced which is one of the features of this inverter.

**E. Analysis of the Capacitors**

The capacitors in SCMLIs play an important role in power transmission and conversion. Their voltage ripples should be controlled within a reasonable range. The voltage ripple of capacitors is related to their capacitance, load value and discharging periods. A low voltage ripple can improve the quality of the output voltage, reduce ripple losses and improve the efficiency of inverters. The two capacitors in the proposed seven-level inverter have the same capacitance. The selection of the two capacitors are the same and therefore, only capacitor $C_2$ is analyzed as follows.

To determine the capacitance of a capacitor, it is necessary to learn its maximum discharge amount. It can be seen from Fig. 4 that $C_2$ is discharged when the output voltage is $3V_{dc}/2$ and $V_{dc}/2$. Therefore, the maximum discharge amount of $C_2$ is the sum of the discharge amount in the three periods of $0-t_1$, $t_2-t_3$ and $t_4-t_5$.

The instants $t_i (i=1,2,3,4,5)$ are intersection points of the sine and triangle waves, which can be calculated by the following equations:

$$t_1 = \frac{\arcsin\left(\frac{1}{3M}\right)}{2\pi f_o}, \quad \text{(2)}$$

$$t_2 = \frac{\arcsin\left(\frac{2}{3M}\right)}{2\pi f_o}, \quad \text{(3)}$$

$$t_3 = \frac{\pi - \arcsin\left(\frac{2}{3M}\right)}{2\pi f_o}, \quad \text{(4)}$$

$$t_4 = \frac{\pi - \arcsin\left(\frac{1}{3M}\right)}{2\pi f_o}, \quad \text{(5)}$$

$$t_5 = \frac{\pi}{2\pi f_o}. \quad \text{(6)}$$

In (2)-(6), $M$ is the modulation index ($M=0.9$ is used in this study) and $f_o$ is output frequency. The discharge amount of the capacitor during the period from $0$ to $t_1$ can be calculated by:
\[ \Delta Q_1 = \int_0^{t_1} I_o \sin(2\pi f_o t) dt, \]  
(7)

where \( \Delta Q_1 \) is the discharge amount during the period of 0-\( t_1 \), \( I_o \) is the output current. Using the same calculation method, the discharge amount \( \Delta Q_2 \) during the period of \( t_2-t_3 \) can be obtained as:

\[ \Delta Q_2 = \int_{t_2}^{t_3} I_o \sin(2\pi f_o t) dt. \]  
(8)

The variables in (8) are the same as the ones in (7). As the operating status of the capacitor is the same during 0-\( t_1 \) and \( t_2-t_3 \), the discharge amount during the two periods is the same as well. Therefore, the maximum discharge amount of \( C_2 \) is given by:

\[ \Delta Q_{C2} = 2\Delta Q_1 + \Delta Q_2. \]  
(9)

Assuming \( k \) is the factor describing the maximum acceptable voltage ripple, the capacitance can be obtained:

\[ C_2 = \frac{\Delta Q_{C2}}{k V_{C2}} , \]  
(10)

where \( V_{C2} \) is the rated voltage of \( C_2 \).

Considering the symmetrical characteristics of the working states of \( C_1 \) and \( C_2 \) in the positive and negative half-cycles, the voltage ripples of the two capacitors are the same. Thus, \( C_2 \) is chosen as an example. It can be seen from (9) that the maximum continuous discharge amount of \( C_2 \) is \( \Delta Q_{C2} \). Therefore, the voltage ripple of \( C_2 \) can be given by:

\[ \Delta V_{C2} = \frac{\Delta Q_{C2}}{C_2} , \]  
(11)

where \( \Delta V_{C2} \) is the voltage ripple of \( C_2 \). As can be seen from (11), the ripple can be controlled within a reasonable range by setting a proper capacitance.

III. POWER LOSSES ANALYSIS

Three types of losses are considered for the switched-capacitor inverters, which include the ripple losses of capacitors (\( P_{\text{rip}} \)), conduction losses (\( P_{\text{con}} \)) and switching losses (\( P_{\text{sw}} \)).

A. Ripple Losses of Capacitors

\( P_{\text{rip}} \) is caused by the voltage fluctuation of the capacitors. \( C_2 \) is still taken as an example, due to the symmetrical working states of the two capacitors. As shown in (11), the voltage ripple of \( C_2 \) is \( \Delta V_{C2} \), therefore the \( P_{\text{rip}} \) can be calculated from:

\[ P_{\text{rip}} = f_o C_2 \Delta V_{C2}^2 . \]  
(12)

A further calculation can be obtained as:

\[ P_{\text{rip}} = f_o \frac{\Delta Q_{C2}^2}{C_2} . \]  
(13)

B. Conduction Losses

The conduction losses of the inverters are caused by the parasitic parameters of devices, such as the on-state resistance of the switches (\( r_s \)), and the equivalent series resistance of each capacitor (\( ESR_C \)). Fig. 5 shows the equivalent circuit of supplying the load. In Fig. 5, \( V_o \), \( r_o \) and \( R_o \) are the output voltage, the equivalent parasitic resistance of the devices and the load resistance, respectively.

\[ \Delta Q_1 = \int_0^{t_1} I_o \sin(2\pi f_o t) dt , \]

\[ \Delta Q_2 = \int_{t_2}^{t_3} I_o \sin(2\pi f_o t) dt . \]

According to Fig. 4, in the interval of \([0, t_1]\), the output level changes between 0 and \( V_{dc}/2 \). Therefore, the power losses during \([0, t_1]\) can be calculated from:

\[ P_{0.0.5V_{dc}} = \int_0^{t_1} [I_o \sin(2\pi f_o t)]^2 \times (ERS_C + 3r_s) \frac{A_{ref} \sin(2\pi f_o t)}{A_c} dt . \]  
(14)

Similarly, the power losses of the other two working modes can be obtained in the same way.

\[ P_{0.5V_{dc}} = \int_{t_2}^{t_3} [I_o \sin(2\pi f_o t)]^2 \times 4r_s \frac{A_{ref} \sin(2\pi f_o t) - A_c}{A_c} dt . \]  
(15)

\[ P_{0.5V_{dc}} = \int_{0.01}^{0.5V_{dc}} [I_o \sin(2\pi f_o t)]^2 \times (ERS_C + 3r_s) \frac{A_{ref} \sin(2\pi f_o t) - 2A_c}{A_c} dt . \]  
(16)

Therefore, the \( P_{\text{con}} \) can be obtained as:

\[ P_{\text{con}} = 4(P_{0.0.5V_{dc}} + P_{0.5V_{dc}} + P_{0.5V_{dc}}) . \]  
(17)

C. Switching Losses

The switching losses can be calculated according to the charging and discharging processes of the parasitic capacitor \( C_s \) in the switches, it is assumed that the capacitance of the parasitic capacitor is linear. The voltage of the parasitic capacitor is gradually charged to \( V_s \) when the switches are turned off, where \( V_s \) is the MVS of the switches, and the \( V_s \) of each switch is shown in Table IV.

\[ \Delta Q_{C2} = \int_0^{t_1} I_o \sin(2\pi f_o t) dt , \]

\[ \Delta Q_2 = \int_{t_2}^{t_3} I_o \sin(2\pi f_o t) dt . \]

The equivalent parameters in the three working modes of the positive half cycle are shown in Table III. The value of \( j \) in Table III indicate that the output \( V_o \) is \( j \times \) times of \( V_{dc}/2 \).

<table>
<thead>
<tr>
<th>( j )</th>
<th>( V_o )</th>
<th>( r_{eq} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
<td>( 4r_s )</td>
</tr>
<tr>
<td>1</td>
<td>( V_{dc}/2 )</td>
<td>( ESR_c + 3r_s )</td>
</tr>
<tr>
<td>2</td>
<td>( V_{dc} )</td>
<td>( 4r_s )</td>
</tr>
<tr>
<td>3</td>
<td>( 3V_{dc}/2 )</td>
<td>( ESR_c + 3r_s )</td>
</tr>
</tbody>
</table>

\[ P_{0.0.5V_{dc}} = \int_0^{t_1} [I_o \sin(2\pi f_o t)]^2 \times (ERS_C + 3r_s) \frac{A_{ref} \sin(2\pi f_o t)}{A_c} dt . \]  
(14)

\[ P_{0.5V_{dc}} = \int_{t_2}^{t_3} [I_o \sin(2\pi f_o t)]^2 \times 4r_s \frac{A_{ref} \sin(2\pi f_o t) - A_c}{A_c} dt . \]  
(15)

\[ P_{0.5V_{dc}} = \int_{0.01}^{0.5V_{dc}} [I_o \sin(2\pi f_o t)]^2 \times (ERS_C + 3r_s) \frac{A_{ref} \sin(2\pi f_o t) - 2A_c}{A_c} dt . \]  
(16)

\[ P_{\text{con}} = 4(P_{0.0.5V_{dc}} + P_{0.5V_{dc}} + P_{0.5V_{dc}}) . \]  
(17)

\[ \Delta Q_{C2} = \int_0^{t_1} I_o \sin(2\pi f_o t) dt , \]

\[ \Delta Q_2 = \int_{t_2}^{t_3} I_o \sin(2\pi f_o t) dt . \]

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<td>( V_{dc}/2 )</td>
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</tr>
<tr>
<td>2</td>
<td>( V_{dc} )</td>
<td>( 4r_s )</td>
</tr>
<tr>
<td>3</td>
<td>( 3V_{dc}/2 )</td>
<td>( ESR_c + 3r_s )</td>
</tr>
</tbody>
</table>

\[ P_{\text{con}} = 4(P_{0.0.5V_{dc}} + P_{0.5V_{dc}} + P_{0.5V_{dc}}) . \]  
(17)

\[ \Delta Q_{C2} = \int_0^{t_1} I_o \sin(2\pi f_o t) dt , \]

\[ \Delta Q_2 = \int_{t_2}^{t_3} I_o \sin(2\pi f_o t) dt . \]

\[ \Delta Q_{C2} = \int_0^{t_1} I_o \sin(2\pi f_o t) dt , \]

\[ \Delta Q_2 = \int_{t_2}^{t_3} I_o \sin(2\pi f_o t) dt . \]
where \( N_i \) is the switching transitions in one period of the reference waveform.

As can be seen from Fig. 4, the switches are repeatedly turned on or off in the corresponding intervals. The \( N_i \) of each switch can be roughly obtained as the ratio of \( f_c \) and \( f_o \) if the switch works in the whole period. However, the switches in the proposed inverter only work in the specific intervals. Therefore, the \( N_i \) of each switch can be calculated as:

\[
N_i = \frac{t_i f_c}{T_s f_o},
\]

where \( t_i \) is the working time of the switches, and it can be obtained from Figs. 3 and 4. \( T_s \) is the time of one cycle. Therefore, \( P_{sw} \) can be calculated from:

\[
P_{sw} = 50 \sum_{i=1}^{10} C_0 V_{dc}^2 t_i f_c.
\]

In summary, the efficiency of the proposed inverter can be calculated as:

\[
\eta = \frac{P_o}{P_o + P_{rip} + P_{con} + P_{sw}},
\]

where \( \eta \) and \( P_o \) are the efficiency and output power of the proposed inverter.

The aforementioned losses can be obtained numerically at different output power. It should be noted that \( r_s, ESR_{C_i}, f_o, f_c, C_i \) are considered 5 m\( \Omega \), 60 m\( \Omega \), 50 Hz, 5 kHz, and 500 pF, respectively, for the losses calculations. The theoretical efficiency of the proposed seven-level inverter is shown in Fig. 6(a). Under the condition of a 30 V dc source and 30 Ω load, the three types of losses can be obtained as: \( P_{rip} = 0.49 \text{ W}, P_{con} = 0.12 \text{ W} \) and \( P_{sw} = 0.3 \text{ W} \). The ratio of the three types of losses is illustrated in Fig. 6(b).

IV. TOPOLOGY EXTENSION AND COMPARISON ANALYSIS

An advantage of the proposed inverter is that it is extensible. There are two ways to extend the topology. Each type has its own characteristics and can be used for different applications.

![Extension module 1](image)

**A. Extending the H-bridge**

It can be seen from Fig. 7(a) that the left side of the proposed inverter is an H-bridge. The first type of extensions is achieved by using multiple H-bridges which are connected through switches and diodes, as shown in Fig. 7(a). It is worth mentioning that the independent power sources inside the extension H-bridges can be replaced by capacitors, which can enhance the applicability of the proposed inverter. In addition, thanks to the right-side capacitors, the output level under this extension can be significantly increased compared to conventional CHB MLIs. In fact, the total output level can be increased by 4 for every additional H-bridge. In the extended topology, the output level of the inverter, \( N_{level1} \), has a quantitative relationship with the number of the extended H-bridges \( n \) and the number of switches \( y \):

\[
N_{level1} = 4n + 7 = \frac{4y + 5}{5}.
\]

**B. Connecting Multiple Inverters in Series**

The other way to extend the inverter is to connect multiple inverters in series. The extended topology with two inverters is shown in Fig. 7(b). Modular design can be achieved without using additional power devices, such as the switches and diodes used in the above extension, which is one of the benefits of this extension. According to the analysis in the last section, the highest output level of a single inverter is \( 3V_{dc}/2 \). Then, the dc source voltage of the series-connected inverters can be set in proportion to increasing the output levels. To ensure the output voltage increases step-by-step, a suitable ratio is 1:7. Assuming \( x \) inverters are connected in series, the proportional relationship between each power source is:

![Extension module 2](image)
Similarly, the quantitative relationship between the output level $N_{\text{level}}$ with the number of connected inverters ($x$) and the number of switches ($y$) is:

$$N_{\text{level}} = 7^x = 7^{10y}.$$  \hfill (25)

C. Comparison of the Two Extensions

Both extensions have their pros and cons. First of all, in the first extension, the output level increases linearly with the number of H-Bridges and switches. The output level in the second extension increases exponentially with the number of the connected inverters and switches. Although the growth rate of the first extension is considerable, the one of the second extension is significantly higher. Secondly, in both extensions, the voltage stress of each switch does not exceed the input source voltage. However, in the second extension, the voltage stress of the switches will increase gradually with the increase of the number of the connected inverters. It is because that the voltage of the dc sources in the series-connected inverter will increase. Finally, the first extension requires only one dc source, while the second requires multiple dc sources. The advantages of the two extensions make them suitable for different applications.

D. Topology Comparison

Table V shows the comparison of the proposed inverter with the recently proposed hybrid MLIs. The comparison has been made under the condition that all inverters have the same output level. These hybrid MLIs are able to limit the voltage stress on the switches to a low value, which is also a feature of the proposed inverter. However, more devices are used in the hybrid MLIs compare to the proposed inverter. In addition, most of them do not have the ability to boost the input voltage. It can be seen from Table V, the proposed inverter solves the above shortcomings, and has a better comprehensive performance.

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Levels</td>
<td>7</td>
<td>7</td>
<td>7</td>
<td>7</td>
<td>7</td>
</tr>
<tr>
<td>Switches</td>
<td>18</td>
<td>16</td>
<td>8</td>
<td>12</td>
<td>10</td>
</tr>
<tr>
<td>Capacitors</td>
<td>4</td>
<td>3</td>
<td>3</td>
<td>0</td>
<td>2</td>
</tr>
<tr>
<td>DC sources</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>3</td>
<td>1</td>
</tr>
<tr>
<td>Voltage boosting</td>
<td>no</td>
<td>no</td>
<td>no</td>
<td>no</td>
<td>yes</td>
</tr>
</tbody>
</table>

In order to further evaluate the superiority of the proposed inverter, it has been comprehensively analyzed and compared with other recently proposed SCMLIs. The main indicators for the comparison are the number of power devices, MVS, and total standing voltage (TSV), etc. The comparison analysis has been carried out under the condition that their output level is $2m+1$ (or $m$ steps). The results are shown in Fig. 8 and Table VI. “$E_i$” and “$E_H$” represent the first and second types of the extension. $V_{\text{step}}$ is the voltage of each step.

It can be seen from Table VI and Fig. 8 that the proposed inverter shows excellent performance amongst the SCMLIs, especially in terms of the switch MVS. Although the value of MVS in [31] is the smallest, $V_{\text{step}}$ in [31] is twice the $V_{\text{step}}$ of the proposed inverter, and therefore, both inverters have the same MVS. However, with the extension of the inverter, the

\[ V_{dc1} : V_{dc2} : \cdots : V_{dcx} = 7 : \cdots : 7^{x-1}. \]  \hfill (24)
number of switches in [31] increases significantly, and the number of capacitors is also higher than the proposed inverter. In terms of the used capacitors, the topology in [23] is also excellent. However, [23] and the second extension of the proposed inverter need to use multiple dc power sources, which limits its application. Only the SCMLIs in [31] are superior to the proposed inverter in terms of TSV. However, the SCMLI in [31] requires a large number of switches.

The gain of the proposed inverter is half that of other inverters when the output is at the same level due to the existence of the voltage-dividing capacitors. If the same output voltage is achieved, the dc power source of the proposed inverter has to be doubled. However, the presence of the voltage-dividing capacitors makes it possible to effectively increase the number of output levels when the components used are equivalent, thereby substantially achieving the same gain. In addition, the two proposed extensions have their advantages which can be seen from Table VI. The first extension is better in reducing the voltage stress of the switches, while the second extension is better in reducing the number of devices. This shows good agreement with the previous analysis of the topology extension.

Table VII shows the comparison of the proposed inverter with the recently proposed SCMLIs feature low voltage stress and T type capacitors. The comparison is performed under the condition that the output level of all inverters is seven. It can be seen from Table VII that the proposed SCMLI is the optimal one in terms of the number of devices. In addition, the \( Q_m \) in Table VII is the discharge amount of capacitors, the benchmark discharge amount of the capacitor working at 0.5\( V_{dc} \) is defined as \( Q_u \). Then, the capacitor discharge amount will be 2\( Q_u \) when it works at \( V_{dc} \), and so forth. It can be seen from the \( Q_m \) of these inverters that the proposed inverter is superior in terms of the number of capacitors and the maximum discharge capacity. Therefore, the proposed inverter has the advantage of using a smaller capacitance.

### Table VII

**Comparison with Low Voltage Stress SCMLIs**

<table>
<thead>
<tr>
<th>Items</th>
<th>[33]</th>
<th>[34]</th>
<th>[35]</th>
<th>[36]</th>
<th>proposed</th>
</tr>
</thead>
<tbody>
<tr>
<td>Levels</td>
<td>7</td>
<td>7</td>
<td>7</td>
<td>7</td>
<td>7</td>
</tr>
<tr>
<td>Switches</td>
<td>10</td>
<td>10</td>
<td>10</td>
<td>9</td>
<td>10</td>
</tr>
<tr>
<td>Diodes</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>Capacitors</td>
<td>4</td>
<td>3</td>
<td>4</td>
<td>3</td>
<td>2</td>
</tr>
<tr>
<td>MVS</td>
<td>( V_{dc} )</td>
<td>( V_{dc} )</td>
<td>( V_{dc} )</td>
<td>( V_{dc} )</td>
<td>( V_{dc} )</td>
</tr>
<tr>
<td>( Q_m )</td>
<td>2( Q_u )</td>
<td>5( Q_u )</td>
<td>5( Q_u )</td>
<td>5( Q_u )</td>
<td>4( Q_u )</td>
</tr>
</tbody>
</table>

**V. SIMULATION VERIFICATION AND EXPERIMENTAL VALIDATION**

### A. Simulations

To verify the effectiveness of the proposed inverter, a model of a seven-level inverter is built in MATLAB/Simulink. Parameters in the simulations are shown in Table VIII.

The simulation results are shown in Figs. 9 and 10. It can be seen from Fig. 9 that the inverter outputs 7-level PWM waves, and the current lags the voltage by about 8.9°. The fast Fourier transform (FFT) of the output voltage is given in Fig. 10. The THD is 16.26%. The 20th harmonic components are larger than others because the carrier frequency is 2 kHz. As other harmonics are much less than the 20th harmonics, this can simplify the filter design.

### TABLE VIII

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input DC source (( V_{dc} ))</td>
<td>30 V</td>
</tr>
<tr>
<td>Output frequency (( f_o ))</td>
<td>50 Hz</td>
</tr>
<tr>
<td>Carrier frequency</td>
<td>2 kHz</td>
</tr>
<tr>
<td>Modulation index (( M ))</td>
<td>0.9</td>
</tr>
<tr>
<td>Capacitors (( C_1, C_2 ))</td>
<td>2200 ( \mu )F</td>
</tr>
<tr>
<td>Loads</td>
<td>30 Ω &amp; 15 mH</td>
</tr>
</tbody>
</table>

**Fig. 9.** Simulation results of the output voltage, current and capacitor voltage.

**Fig. 10.** THD of the output voltage.

### B. Steady-state Analysis

In order to validate the feasibility of the proposed inverter, a seven-level prototype has been constructed to test its steady-state and dynamic performance. The parameters of the prototype are listed in Table IX and the experimental platform is shown in Fig. 11. The waveforms of voltage and current, the voltage waveforms of the two capacitors and the voltage stress of switches are studied.
Experiment has been conducted in an R-L load (30 Ω & 15 mH) to test the steady-state performance of the proposed inverter. The waveforms of voltage and current are shown in Fig. 12(a). The voltage of each step is 15 V and the maximum output voltage is 45 V. As the voltage of the input source used is 30 V, a boost gain of 1.5 is achieved which matches the previous analysis. The steady-state capacitor voltages are shown in Fig. 12(b). It can be seen that the two capacitors are self-balancing with small voltage ripples, which is consistent with the analysis of capacitor voltage balancing. Under the same resistance condition, increase the inductance value to 60 mH. The results are shown in Fig. 12(c). It is obvious that the output current lags behind the voltage which proves the capability of supplying inductive loads.

Fig. 13 shows the voltage stress of the switches. The MVS of the switches is limited to the input power voltage $V_{dc}$, which is an advantage of the proposed inverter. A total of ten switches are used in the proposed seven-level inverter. The voltage stress of $S_1$ to $S_8$ is $V_{dc}$, and is $V_{dc}/2$ of $S_9$ and $S_{10}$.

The experimental results of the current stress of the switches are shown in Fig. 14. Thanks to the small voltage ripples of the capacitors, all switch current stress is small. The peak value of the current stress of the switches in the capacitor charging circuit is only about 1 A.

### C. Dynamic Analysis

![Fig. 12. Steady-state experimental waveforms. (a) output voltage and current. (b) capacitor voltage. (c) load voltage and current under 200Hz.](image)

![Fig. 13. Voltage stress of switches. (a) $S_1$, $S_2$, $S_3$, $S_4$ (b) $S_5$, $S_6$, $S_7$, $S_8$ (c) $S_9$, $S_{10}$.](image)
An inverter with excellent performance should be tolerant to the changes of working conditions, such as sudden changes of the load, input voltage, frequency and amplitude of modulation waves, etc. To further test the performance of the inverter in dynamic conditions, the following experiments have been performed. Fig. 15 shows the output voltage under the condition that the amplitude of modulation wave changes. It can be seen from Fig. 15 that the output voltage changes from seven-level to five-level and then to three-level when $M$ changes from 0.9 to 0.6 and then to 0.3. The transient processes complete fast, which proves the excellent dynamic performance of the proposed inverter. The experimental results also show good agreement with the analysis in Table II.

The proposed inverter can properly adapt to the conditions when the frequency of the modulation wave changes. The output voltage and current when frequency changes are shown in Fig. 16. It is observed that the inverter can transit accordingly with a fast transient response in both cases (50 to 100Hz and 100 to 50Hz).

The experimental results under the condition of a sudden change of the load are shown in Fig. 17. The setting of the load change is: from no-load condition changes to an $R$-$L$ load (30Ω & 15mH), and then changes to a resistive load (50Ω). The results illustrate that the inverter performs well when the load changes.
Load affects the discharging current of the capacitors, and therefore, affects the discharge amount of the capacitors, which will, in turn, affects the voltage ripples of the capacitors. Fig. 18 shows the voltage ripples of capacitors when the load changes. It can be seen that the voltage ripples of capacitors decrease with the increase of impedance.

The charging and discharging of $C_1$ and $C_2$ when the dc source voltage changes from 30 V to 20 V and the reverse processes, which can emulate the abrupt changes of the input voltage (e.g. a power source of a solar cell), has been carried. As shown in Figs. 19 and 20, the inverter performs well during the abrupt change of its input voltage. It can be seen from the experimental results that the inverter can quickly reach to a new steady-state and has good dynamic performance in adapting to changes of the input voltage.

**D. Efficiency analysis**

The efficiency v.s. load power curves at different supply voltages are shown in Fig. 21. The efficiency is higher than 95% when the output power does not exceed 250 W. When the inverter output power is 200 W under an input voltage of 30 V, the efficiency is 95.5%, which is higher than the 95.3% in [25] and 89% in [26]. In addition, it can be seen from Fig. 21 that the efficiency increases as the input voltage increases.
under the same load power. This is because the load current will decrease, and therefore reduce power losses.

E. Discussion

The voltage boosting, capacitor self-balancing and capability of supplying inductive loads of the proposed inverter have been investigated through simulation verification and experimental validation. The experimental results show that with an input dc voltage of 30 V, the amplitude of output voltage is 45 V, achieving a voltage gain of 1.5. The experiment also proves that the capacitor voltages can achieve self-balancing. In addition, the capability of supplying the inductive load is validated through tests under a load of 30 Ω and 60 mH. Compared with other SCMLIs, the proposed inverter features low MVS: the voltage stress of all switches does not exceed the dc source voltage. Moreover, the working conditions may change when the inverter is operating. For example, the amplitude and frequency of the sinusoidal modulation wave, and sudden changes in the load and voltage of input dc source. These scenarios are also tested through experiments of dynamic changing, which show that the proposed inverter is capable to respond to dynamics and stabilize in a new working state quickly.

VI. CONCLUSION

A novel T-type switched-capacitor multilevel inverter with low voltage stress and capacitor voltage self-balancing capability has been proposed in this paper. Only two capacitors are used to achieve a seven-level output without affecting the switch count.

Outcomes and features of this topology include: 1) Compared to conventional SCMLIs, the maximum voltage stress of the switches in the proposed inverter can be limited to less than the input dc voltage $V_{dc}$. The low voltage stress is one of the excellent features that make the proposed inverter an attractive alternative for medium- and high-voltage photovoltaic power generation applications. 2) With the merit of the low voltage stress, a voltage boost gain of 1.5 can be achieved as well. 3) Two topology extension schemes have been proposed to achieve a higher voltage boost gain and higher output level, which is superior in its flexible extensibility and capability of supplying inductive loads. 4) The capacitor voltages of the proposed inverter can be self-balanced which can simplify its control complexity. 5) The capacity of supplying the inductive load is ensured by the bidirectional energy loop in each working mode.

The above features of the proposed inverter have been validated through a seven-level experimental prototype. The experimental results indicate that the inverter has an excellent performance in both steady-state and dynamic conditions.

REFERENCES


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