A safe commutation strategy based on LRCD snubber circuit for five-level active neutral point clamped inverter

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Abstract

Spike voltage generated during the turn-off transient is one of the key threats to IGBT. When the traditional RCD snubber is applied in the five-level active neutral point clamped (5L-ANPC) inverter to clamp the spike voltage, a transient large current caused by the charging of the snubber capacitor will increase power losses and current stresses of IGBTs. This study proposes a LRCD snubber for the 5L-ANPC inverter. It can effectively suppress the spike voltages and the transient large currents of IGBTs, thus reducing the power loss. For the 5L-ANPC inverter, the traditional commutation method also poses a threat to IGBTs. Near the zero-crossing point of the output voltage, the traditional commutation method will cause the IGBTs that operate at the carrier frequency to withstand three times of the rated voltage. This study proposes a safe commutation strategy based on the LRCD snubber. When the output voltage passes through zero-crossing point, the proposed commutation strategy uses four transition switching states to avoid the breakdown of IGBTs induced by overvoltage. It can ensure that all the IGBTs operate in the safe operating area (SOA). Experimental results are presented to verify the validity of the proposed method.

1 | INTRODUCTION

In medium-voltage and high-power application, multilevel topology has some advantages such as low voltage stress of switches, small $\frac{dv}{dt}$ and high system efficiency. The classic multilevel topologies include neutral point clamped (NPC) topology, flying capacitor (FC) topology, cascaded H-bridge (CHB) topology and modular multilevel topology (MMC) [1–3]. However, with the number of voltage levels increase, the NPC converter is difficult to achieve voltage balance for some operating conditions that involve large modulation index and active load currents [4]. The FC converter needs a lot of capacitors. The CHB converter requires multiple isolated DC power supplies. The MMC converter cannot avoid low-frequency oscillation of the capacitor voltages [5–8].

Compared with the above topologies, five-level active neutral point clamped (5L-ANPC) inverter has the advantages of less passive components and higher system efficiency. In addition, the capacitor voltages of the 5L-ANPC inverter can be maintained and balanced within a limited margin of their reference values, no matter what the modulation index and power factor angle are. Therefore, 5L-ANPC inverter has broad application prospects in the fields of renewable energy, electric vehicle charging devices, and motor drives [9–11]. Figure 1 shows the single-phase topology of the 5L-ANPC inverter, where $C_u$ and $C_d$ are dc-link capacitors. $C_f$ is an FC. $T_1$–$T_{12}$ are IGBTs, and $DF_1$–$DF_{12}$ are freewheeling diodes. If the dc-link voltage is $4E$, then the rated voltages of dc-link capacitors and FCs are $2E$ and $E$, respectively. $T_1$–$T_8$ require two switches connected in series, so that the rated voltage of each IGBT is $E$ [12].

With the power level improvement, the spike voltage of IGBT generated during the turn-off transient is higher and more obvious. The spike voltage will increase the power losses and the voltage stresses of IGBT. A passive or active snubber is required to clamp the spike voltage [13–27].

Passive snubber uses capacitor, resistor, and diode to suppress the spike voltage. C snubber is the simplest passive snubber consisting of a single capacitor. However, C snubber is not suitable for suppressing the spike voltage of the series-connected switches of the 5L-ANPC inverter [19]. RC snubber, which is directly connected in parallel with switch, consists of a resistor and a capacitor. However, the voltage drop
across the resistor will increase the overvoltage of IGBT during turn-off, and is undesirable [19]. RCD is another widely used passive snubber. It can not only suppress the spike voltage but also reduce the unbalanced voltage distribution between the series-connected switches [20–22]. However, RCD snubber will cause a transient large current due to the charging of snubber capacitor, increasing the power losses and the current stresses of switches [13].

Active snubber usually uses auxiliary switches to return the energy stored in the snubber capacitor back to the main circuit to reduce the power loss [23–26]. However, active snubber needs additional control circuits, which increases the complexity and cost of the system.

For the 5L-ANPC inverter, the experimental results show that the traditional commutation method will cause $T_9$ and $T_{10}$ to withstand about three times of the rated voltage when the output voltage passes through zero-crossing point. Section 2 analyses the reason in detail.

This study analyses the application of RCD snubber in the 5L-ANPC inverter. The conclusion shows that some switches will withstand high current stresses due to the existence of transient large currents. To overcome the shortcomings of the RCD snubber, based on the Undeland circuit [27], this study proposes a LRCD snubber for the 5L-ANPC inverter. The LRCD snubber can suppress the spike voltages and reduce the current stresses of IGBTs, thus improving the system efficiency. In addition, this study analyses the proposed commutation strategy, which takes several hundreds of nanoseconds. Therefore, when the switching state is $\phi 1$ to $\phi 8$ of the 5L-ANPC inverter are listed in Table 1. The series-connected switches should be operated simultaneously. $(T_1, T_3), (T_5, T_7), (T_9, T_{10})$ and $(T_{11}, T_{12})$ are complementary switch pairs, which should be operated in a complementary way. When the output voltage is positive, $T_1, T_2, T_5, T_6$ are turned on, and $T_3, T_4, T_7, T_8$ are turned off. When the output voltage is negative, $T_1, T_2, T_3, T_4$ are turned off and, $T_5, T_6, T_7, T_8$ are turned on. So, the series-connected switches can be operated at the fundamental frequency. It can be noticed that voltage levels $E$ and $-E$ each correspond to two redundant switching states. The switching states $\phi 2$ and $\phi 3$ generate the same voltage level $E$ but with reverse $I_f$. Besides, the switching states $\phi 6$ and $\phi 7$ also generate the same voltage level $-E$ but with reverse $I_f$. Therefore, the redundant switching states can be used to control the FC voltage [12].

### 2.1 Analysis of traditional commutation method

The reference direction of the load current $I_f$ is shown in Figure 1. When the output voltage changes from positive to negative, the switching state changes from $\phi 4$ to $\phi 5$ [12]. When the switching state is $\phi 4$ and $I_f > 0$, the current path is $T_5 \rightarrow T_6 \rightarrow D_{F10} \rightarrow D_{F12}$. With $O$ as the zero-potential point, the potential of each node at this time is shown in Figure 2(a). After $T_1, T_2, T_5$, and $T_6$ are turned off, $T_3, T_4, T_7,$ and $T_8$ cannot be turned on immediately due to dead-time. The current path during the dead-time is $D_{F2} \rightarrow D_{F8} \rightarrow D_{F10} \rightarrow D_{F12}$. During the dead-time, the voltages of $T_1$ and $T_2$ will rise from 0 to $E/2$, which takes several hundreds of nanoseconds. Therefore, when

### TABLE 1 Switching states of the five-level active neutral point clamped (5L-ANPC) inverter

| Output voltage | $T_1$ | $T_3$ | $T_5$ | $T_7$ | $T_9$ | $T_{10}$ | $T_{11}$ | $T_{12}$ | $I_f$ | Switching state |
|----------------|------|------|------|------|------|--------|--------|--------|-----|---------------|
| $2E$           | 1    | 1    | 1    | 1    | 0    | $I_L$  | $\phi 1$|        |     |               |
| $E$            | 1    | 1    | 0    | 1    | 0    | $-I_L$ | $\phi 2$|        |     |               |
| $-E$           | 0    | 0    | 1    | 1    | 0    | 0      | $\phi 3$|        |     |               |
| $-2E$          | 0    | 0    | 0    | 0    | 0    | 0      | $\phi 4$|        |     |               |

**FIGURE 1** Single-phase circuit of the five-level active neutral point clamped (5L-ANPC) inverter

**TABLE 1 Switching states of the five-level active neutral point clamped (5L-ANPC) inverter**

| Output voltage | $T_1$ | $T_3$ | $T_5$ | $T_7$ | $T_9$ | $T_{10}$ | $T_{11}$ | $T_{12}$ | $I_f$ | Switching state |
|----------------|------|------|------|------|------|--------|--------|--------|-----|---------------|
| $2E$           | 1    | 1    | 1    | 1    | 0    | $I_L$  | $\phi 1$|        |     |               |
| $E$            | 1    | 1    | 0    | 1    | 0    | $-I_L$ | $\phi 2$|        |     |               |
| $-E$           | 0    | 0    | 1    | 1    | 0    | 0      | $\phi 3$|        |     |               |
| $-2E$          | 0    | 0    | 0    | 0    | 0    | 0      | $\phi 4$|        |     |               |
FIGURE 2 The diagram of current path: (a) the current path of the switching state $S_4$, (b) the potential of each node when the voltage of $T_1$ is zero during the dead time, (c) the potential of each node after the voltage of $T_1$ reaches $E/2$ during the dead time.

FIGURE 3 5L-ANPC topology with RCD snubbers. (a) RCD snubber, (b) The path of the transient large current.

The $I_L$ has just completed the commutation, the potential of the collector of the $T_9$ is still $2E$, as shown in Figure 2(b). At this time, the voltage stress of $T_9$ is about $3E$. When the voltages of $T_1$ and $T_2$ reach $E/2$, the voltage stress of $T_9$ is about $2E$, as shown in Figure 2(c). Therefore, the maximum voltage stress of $T_9$ can reach about $3E$ during the rising process of the voltages of $T_1$ and $T_2$.

Similarly, when the output voltage changes from negative to positive, the switching state changes from $S_5$ to $S_4$ [12]. If $I_L < 0$, the maximum voltage stress of $T_{10}$ can reach about $3E$ during the rising process of the voltages across the $T_3$ and $T_4$.

2.2 Analysis of the transient large current path of RCD snubber

The traditional RCD snubber is shown in Figure 3(a). It consists of capacitor $C_s$, diode $D_s$, resistor $R_s$ and static voltage balancing resistor $R$.

Figure 3(b) shows the single-phase 5L-ANPC topology with RCD snubbers, which may cause transient large currents due to the charging of snubber capacitors. Taking the switching state
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3.2 The snubber process of the LRCD snubber when the output voltage passes through zero-crossing point

When the output voltage changes from positive to negative, the switching sequence is $\text{St}2 \to \text{Sm1} \to \text{Sm2} \to \text{Sm6}$. The commutation process is divided into four stages. If $I_L > 0$, the schematic curves of the snubber capacitor voltages during the commutation process are shown in Figure 5.

Stage 3 ($0 < t < t_1$) starts with the switching state $\text{St}2$. If $I_L > 0$, the path of $I_L$ is $T_1 \to T_2 \to T_9 \to T_{10} \to D_{F5} \to D_{F6} \to C_1$, as indicated by the blue solid lines in Figure 3(b).

At $t = t_1$, stage 2 ($t_1 < t < t_2$) begins. $T_1$, $T_2$, $T_3$, and $T_4$ are turned off, and the switching state changes from $\text{St}2$ to $\text{Sm1}$. $C_{s1}$, $C_{s2}$, $C_{s5}$, and $C_{s6}$ start to be charged, while $C_{s3}$, $C_{s4}$, $C_{s7}$, $C_{s8}$, and $C_{s10}$ start to be discharged. Because $C_L$ is much larger than other snubber capacitors, the peak of charging current $I_{C_L}$ is set as $I_{C_L} = 0.5 I_L$. The inductance of $L_1$ and $L_2$ are $L_1 = L_2 = 0.1 \mu\text{H}$ and the resistance of $R_{L1}$ and $R_{L2}$ are $R_{L1} = R_{L2} = 0.1 \Omega$. The capacitance of $C_{s1}$ and $C_{s2}$ are $C_{s1} = C_{s2} = 0.5 \mu\text{F}$ and the capacitance of $C_{s5}$ and $C_{s6}$ are $C_{s5} = C_{s6} = 1 \mu\text{F}$. The capacitance of $C_{s3}$ and $C_{s4}$ are $C_{s3} = C_{s4} = 2 \mu\text{F}$. The resistance of $R_S$ is $R_S = 100 \Omega$. The switching frequency of $T_1$ is $f_{sw} = 20 \text{kHz}$. The output voltage of $T_1$ is $U_{O1} = 100 \text{V}$.
than $C_s$, the voltages of $C_{L1}$ and $C_{L2}$ are almost constant during this process. Therefore, the induced voltages of $L_1$ and $L_2$ and the currents through $RL_1$ and $RL_2$ can be neglected. At this time, the current path is indicated by the red solid lines in Figure 6(a). The equivalent circuit is shown in Figure 7(a). According to Figure 7(a), the approximate expressions of snubber capacitor currents can be written as:

$$
\begin{align*}
I_{G1}(t) &= \frac{I_L}{3} + \frac{I_L}{4} e^{-2(t-t_1)/(R,C_s)} + \frac{5I_L}{12} e^{-6(t-t_1)/(R,C_s)} \\
I_{G3}(t) &= \frac{I_L}{3} - \frac{I_L}{4} e^{-2(t-t_1)/(R,C_s)} - \frac{I_L}{12} e^{-6(t-t_1)/(R,C_s)} \\
I_{G5}(t) &= \frac{I_L}{6} + \frac{I_L}{4} e^{-2(t-t_1)/(R,C_s)} - \frac{5I_L}{12} e^{-6(t-t_1)/(R,C_s)} \\
I_{G7}(t) &= \frac{I_L}{6} - \frac{I_L}{4} e^{-2(t-t_1)/(R,C_s)} + \frac{I_L}{12} e^{-6(t-t_1)/(R,C_s)} \\
I_{G10}(t) &= \frac{I_L}{3} - \frac{I_L}{4} e^{-6(t-t_1)/(R,C_s)}
\end{align*}
$$

(1)

According to Equation (1), the snubber capacitor voltages can be written as:

$$
\begin{align*}
U_{c1}(t) &= \frac{I_L}{3C_s} (t-t_1) + \frac{7I_L R_s}{36} - \frac{I_L R_s}{8} e^{-2(t-t_1)/(R,C_s)} \\
&\quad - \frac{5I_L R_s}{72} e^{-6(t-t_1)/(R,C_s)} \\
U_{c3}(t) &= E - \frac{I_L}{3C_s} (t-t_1) + \frac{5I_L R_s}{36} - \frac{I_L R_s}{8} e^{-2(t-t_1)/(R,C_s)} \\
&\quad - \frac{I_L R_s}{72} e^{-6(t-t_1)/(R,C_s)} \\
U_{c5}(t) &= \frac{I_L}{6C_s} (t-t_1) + \frac{I_L R_s}{18} - \frac{I_L R_s}{8} e^{-2(t-t_1)/(R,C_s)} \\
&\quad + \frac{5I_L R_s}{72} e^{-6(t-t_1)/(R,C_s)} \\
U_{c7}(t) &= E - \frac{I_L}{6C_s} (t-t_1) + \frac{I_L R_s}{9} - \frac{I_L R_s}{8} e^{-2(t-t_1)/(R,C_s)} \\
&\quad + \frac{I_L R_s}{72} e^{-6(t-t_1)/(R,C_s)} \\
U_{c10}(t) &= E - \frac{I_L}{3C_s} (t-t_1) + \frac{I_L R_s}{18} \left(1 - e^{-2(t-t_1)/(R,C_s)}\right)
\end{align*}
$$

(2)

Since the time constant $R_s C_s$ is very small (hundreds of nanoseconds), the exponential terms in Equation (2) will decay rapidly to 0. Therefore, snubber capacitor voltages $U_{c10}(t)$ vary approximately linearly, as shown in Figure 5. Since $U_{T3}(t) = U_{c3}(t) - I_{c3}(t) R_s$, $U_{T3}(t)$ can be written as:

$$
U_{T3}(t) = E - \frac{I_L}{3C_s} (t-t_1) - \frac{7I_L R_s}{36} + \frac{I_L R_s}{8} e^{-2(t-t_1)/(R,C_s)} \\
&\quad + \frac{5I_L R_s}{72} e^{-6(t-t_1)/(R,C_s)}
$$

(3)

At $t = t_2$, $U_{c1}(t) = E$, $U_{T3}(t) = 0$, $D_{F3}$ and $D_{F4}$ turns on. $I_{L1}$ commutates to path $D_{F3} \rightarrow D_{F4} \rightarrow T_9 \rightarrow C_{L1} \rightarrow D_{F12}$. At this time, $I_{L1} \approx I_L$ and starts to decrease. The energy stored in $L_1$ is transferred to $C_{L1}$, $C_{L1}$ and $C_{L2}$. The current path is indicated by the red solid lines in Figure 6(b). The equivalent discharging circuit of $L_1$ is shown in Figure 7(b). According to Figure 7(b), the current expression of $L_1$ can be written as:

$$
\begin{align*}
I_{L1}(t) &= A_1 e^{\theta 1} + A_2 e^{\theta 2} \\
A_1 &= \frac{E}{p_2 - p_1} I_{\text{max}} A_2 &= \frac{p_1}{p_2 - p_1} I_{\text{max}} \\
\theta_1 &= -I + \sqrt{I^2 - 4R_s^2 I_{\text{LC}}} \\
\theta_2 &= -I - \sqrt{I^2 - 4R_s^2 I_{\text{LC}}} \\
p_1 &= \frac{-I + \sqrt{I^2 - 4R_s^2 I_{\text{LC}}}}{2R_s I_{\text{LC}}} \\
p_2 &= \frac{-I - \sqrt{I^2 - 4R_s^2 I_{\text{LC}}}}{2R_s I_{\text{LC}}}
\end{align*}
$$

(4)
FIGURE 6 Snubber process of the LRCD snubber when the output voltage changes from positive to negative (a) \( t_1 < t < t_2 \), (b) \( t_2 < t < t_3 \), (c) \( t_3 < t < t_4 \), (d) \( t_5 < t < t_6 \).

FIGURE 7 Equivalent circuit diagram: (a) equivalent circuit of Figure 6(a,b) equivalent discharging circuit of the inductor

\[
\begin{align*}
\text{Equation (4), the maximum induced voltage of the inductor } L_1 & \text{ is:} \\
\Delta U & = L_1 \left| \frac{dI_{L1}(t)}{dt} \right|_{\text{max}} = \frac{LR_1I_{\text{max}}}{\sqrt{L^2 - 4R_1^2LC}} \left| \phi_{\text{p1}} - \phi_{\text{p2}} \right|_{\text{max}}
\end{align*}
\]

According to Equation (5), the maximum induced voltage \( \Delta U_1 \) of \( L_1 \) can be obtained. Therefore, the spike voltages across \( T_1 \) and \( T_2 \) are \( \Delta U_1/2 + E \). Since \( C_\text{L} \) is much larger than \( C_s \), most of the energy stored in \( L_1 \) is transferred to \( C_{\text{L1}} \). So, the spike voltages across \( T_1 \) and \( T_2 \) are mainly clamped by \( C_{\text{L1}} \). When the snubber capacitor voltages reach steady state, \( U_{C_s1}(t) = U_{C_s2}(t) = E \), \( U_{C_s3}(t) = U_{C_s4}(t) = U_{C_s9}(t) = 0 \), \( U_{C_s5}(t) = U_{C_s6}(t) = U_{C_s7}(t) = U_{C_s8}(t) = E/2 \), as shown in Figure 5. It can be seen that \( T_9 \) and \( T_{10} \) operate in the SOA.

At \( t = t_3 \), stage 3 \( (t_3 < t < t_4) \) begins. \( T_3 \) and \( T_4 \) are turned on, and the switching state changes from \( \text{Sn1} \) to \( \text{Sn2} \). The path of \( I_L \) remains unchanged.
The snubber process at T9 turn-off

Then stage 4 (\(t_4 < t < t_5\)) begins when \(t = t_4\). T7 and T8 are turned on, and the switching state changes from Sm2 to Sm6. C_{g7}, C_{s8} start to be discharged, while C_{s5}, C_{s6} and C_{s10} start to be charged. At this time, the current paths are indicated by the red solid line in Figure 6(c). According to Figure 6(c), the charging circuit of C_{s5}, C_{s6} and C_{s10} is a second-order circuit. The currents through L_2, C_{s5} and C_{s10} can be written as:

\[
\begin{align*}
I_{C_{s5}}(t) & = \frac{E}{3} \sqrt{\frac{3C_{s}}{2L}} \sin\left(\sqrt{\frac{2}{3L}C_{s}}(t - t_4)\right) \\
I_{C_{s10}}(t) & = \frac{2E}{3} \sqrt{\frac{3C_{s}}{2L}} \sin\left(\sqrt{\frac{2}{3L}C_{s}}(t - t_4)\right)
\end{align*}
\]

According to Equation (6), the snubber capacitor voltages can be written as:

\[
\begin{align*}
U_{C_{s5}}(t) & = E - \frac{E}{2} \cos\left(\sqrt{\frac{2}{3L}C_{s}}(t - t_4)\right) \\
U_{C_{s10}}(t) & = E - E \cos\left(\sqrt{\frac{2}{3L}C_{s}}(t - t_4)\right)
\end{align*}
\]

When \(t = t_5\), \(U_{C_{s5}}(t) = U_{C_{s6}}(t) = U_{C_{s10}}(t) = E\), and D_{f12} turns on. The energy stored in L_2 is transferred to C_{s5}, C_{s6}, C_{s10} and C_{l2}. The current paths are indicated by the red solid lines in Figure 6(d). The equivalent discharging circuit of L_2 is the same as Figure 7(b), so the current expression of L_2 can be expressed by Equation (4). But at this time, \(C\), \(I_{\text{max}}\) and \(y\) are respectively:

\[
\begin{align*}
C &= \frac{3C_{s}}{2} + C_{l} \\
I_{\text{max}} &= E \sqrt{\frac{3C_{s}}{2L}} \\
y &= t - t_5
\end{align*}
\]

According to Equations (4), (5) and (8), the maximum induced voltage \(\Delta U_{l2}\) of L_2 can be obtained. The spike voltages across T_5 and T_6 are \(E + \Delta U_{l2}/2\), while that across T_{10} is \(E + \Delta U_{l2}\). Since \(C_{s}\) is much larger than \(C_{l}\), most of the energy stored in L_2 is transferred to \(C_{l}\). Therefore, the spike voltages across T_5, T_6 and T_{10} are mainly clamped by \(C_{l}\).

3.3 The snubber process at T9 turn-off

Take the switching state changes from Sm2 to Sm6 as an example, the snubber process is divided into three stages. The schematic curves of the snubber capacitor voltages during the whole snubber process are shown in Figure 8.

Stage 1 (\(0 < t < t_4\)) starts with the switching state Sm2. If \(I_{l} > 0\), the path of \(I_{l}\) is \(L_1 \rightarrow T_1 \rightarrow T_2 \rightarrow T_9 \rightarrow C_l \rightarrow D_{f12}\). At this time, the snubber capacitor voltages \(U_{C_{s1}}(t) = U_{C_{s2}}(t) = U_{C_{s3}}(t) = U_{C_{s6}}(t) = U_{C_{s9}}(t) = 0\), and \(U_{C_{s4}}(t) = U_{C_{s7}}(t) = U_{C_{s8}}(t) = U_{C_{s10}}(t) = E\) as shown in Figure 8.

At \(t = t_4\), T_9 is turned off and stage 2 (\(t_4 < t < t_5\)) begins. \(C_{s9}\) starts to be charged and \(C_{s10}\) starts to be discharged. Due to the induced voltage of L_1, D_{f11} turns on. Since \(C_{l}\) is much larger than \(C_{s}\), the voltage of \(C_{l}\) is almost constant during this process. Therefore, the induced voltage of L_1 and the current through \(R_{1}\) are enough small to be neglected. The current path is indicated by the red solid lines in Figure 9(a). The equivalent circuit of this process is shown in Figure 10. According to Figure 10, the approximate expressions of the voltages and currents of \(C_{s9}\) and \(C_{s10}\) can be written as:

\[
\begin{align*}
U_{C_{s9}}(t) &= \frac{I_{l}}{2} \left(1 + e^{-2(t-t_4)/(R_{l}C_{s})}\right) \\
U_{C_{s10}}(t) &= \frac{I_{l}}{2} \left(1 - e^{-2(t-t_4)/(R_{l}C_{s})}\right) \\
U_{C_{s9}}(t) &= \frac{I_{l}R_{l}}{2C_{s}}(t-t_4) + \frac{I_{l}R_{l}}{4} \left(1 - e^{-2(t-t_4)/(R_{l}C_{s})}\right) \\
U_{C_{s10}}(t) &= E - \frac{I_{l}R_{l}}{2C_{s}}(t-t_4) + \frac{I_{l}R_{l}}{4} \left(1 - e^{-2(t-t_4)/(R_{l}C_{s})}\right)
\end{align*}
\]

Since the time constant \(R_{l}C_{s}\) is very small (hundreds of nanoseconds), the exponential terms in Equation (9) will decay rapidly to 0. Therefore, snubber capacitor voltages \(U_{C_{s1}}(t)\) and \(U_{C_{s10}}(t)\) vary approximately linearly, as shown in Figure 8. The voltages across T_9 and T_{10} can be written as:

\[
\begin{align*}
U_{T_{9}}(t) &= \frac{I_{l}}{2C_{s}}(t-t_4) + \frac{I_{l}R_{l}}{4} \left(1 - e^{-2(t-t_4)/(R_{l}C_{s})}\right) \\
U_{T_{10}}(t) &= E - \frac{I_{l}R_{l}}{2C_{s}}(t-t_4) - \frac{I_{l}R_{l}}{4} \left(1 - e^{-2(t-t_4)/(R_{l}C_{s})}\right)
\end{align*}
\]
to decrease. The energy stored in $L_1$ is transferred to $C_{L+1}, C_{L+3}, C_{L+2}$, and $C_{L+0}$. The current paths are indicated by the red solid lines in Figure 9(b). The equivalent discharging circuit of the inductor $L_1$ is the same as Figure 7(b), so the current expression of $L_1$ can be expressed by Equation (4). But at this time, $C_I, I_{max}$ and $y$ are respectively:

$$
\begin{align*}
C &= \frac{3C_I}{2} + C_L, \\
I_{max} &\approx I_L, \\
y &= t - t_2
\end{align*}
$$

According to Equations (4), (5) and (11), the maximum induced voltage $\Delta U_3$ of $L_1$ can be obtained. Therefore, the spike voltages across $T_3$ and $T_4$ are $E + \Delta U_3/2$, while that across $T_9$ is $E + \Delta U_3$.

At $t = t_3$, $T_{10}$ is turned on and stage 3 ($t_3 < t < t_4$) begins. The path of $I_L$ is $T_5 \rightarrow T_6 \rightarrow D_{F10} \rightarrow D_{F12}$.

### 3.4 The snubber process at $T_{10}$ turn-off

Take the switching state changes from $St_4$ to $St_2$ as an example, the snubber process is divided into two stages. The schematic curves of the snubber capacitor voltages during the whole snubber process are shown in Figure 11.

Stage 1 ($0 < t < t_1$) starts with the switching state $St_4$. If $I_L > 0$, the path of $I_L$ is $T_9 \rightarrow T_6 \rightarrow D_{F10} \rightarrow D_{F12}$. At this time, the snubber capacitor voltages $U_{c1}(t) = U_{c2}(t) = U_{c3}(t) = U_{c4}(t) = U_{c5}(t) = U_{c6}(t) = U_{c7}(t) = U_{c8}(t) = U_{c9}(t) = E$ as shown in Figure 11.

At $t = t_1$, $T_{10}$ is turned off. The path of $I_L$ remains unchanged during the dead-time.

At $t = t_2$, $T_9$ is turned on and stages 2 ($t_2 < t < t_3$) begins. $C_{L+9}$ starts to be discharged. In this process, the current rise rate of $L_1$ is:

$$
\frac{dI_{L1}(t)}{dt} = E
$$

Therefore, the induced voltage of $L_1$ is $E$, which will cause $C_{s3}$ and $C_{s4}$ to be discharged to $E/2$, as shown in Figure 11. The current path is indicated by the red solid line in Figure 12(a).

When $t = t_3$, the current through $D_{T10}$ is 0, while that through $L_1$ is approximately equal to $I_L$, and continues to increase. At this time, $C_{s3}$, $C_{s4}$ and $C_{s10}$ start to be charged. The current paths are indicated by the red solid line in Figure 12(b). The equivalent circuit is shown in Figure 13. According to Figure 13, the expressions of voltages and currents can be written as:

\[
\begin{align*}
I_{L1}(t) &= E\sqrt{3} \frac{3C_s}{2L} \sin \left( \sqrt{\frac{2}{3LC_s}} (t - t_3) \right) + I_L \\
I_{C3}(t) &= \frac{E}{3} \sqrt{3} \frac{3C_s}{2L} \sin \left( \sqrt{\frac{2}{3LC_s}} (t - t_3) \right) \\
I_{C10}(t) &= \frac{2E}{3} \sqrt{3} \frac{3C_s}{2L} \sin \left( \sqrt{\frac{2}{3LC_s}} (t - t_3) \right) \\
U_{C3}(t) &= E - \frac{E}{2} \cos \left( \sqrt{\frac{2}{3LC_s}} (t - t_3) \right) \\
U_{C10}(t) &= E - E \cos \left( \sqrt{\frac{2}{3LC_s}} (t - t_3) \right)
\end{align*}
\] (13)

At $t = t_4$, $U_{C3}(t) = U_{C4}(t) = U_{C10}(t) = E$ and $D_{T1}$ turns on. The current through $L_1$ starts to decrease until it is equal to $I_L$. The energy stored in $L_1$ is transferred to $C_{s3}$, $C_{s4}$ and $C_{s10}$. The current paths are indicated by the red solid lines in Figure 12(c). The equivalent charging circuit of $C_{s3}$, $C_{s4}$ and $C_{s10}$ is the same as Figure 7(b), so the current expression of $L_1$ can be expressed by Equation (4). But at this time, $C$, $I_{max}$ and $y$ are respectively:

\[
\begin{align*}
C &= C_L + \frac{3C_s}{2} \\
I_{max} &= E \sqrt{\frac{3C_s}{2L}} + I_L \\
I_{L1} &= I_L \\
y &= t - t_4
\end{align*}
\] (14)

According to Equations (4), (5) and (14), the maximum induced voltage $\Delta U_4$ of $L_1$ can be obtained. Therefore, the spike voltages across $T_3$ and $T_4$ are $E + \Delta U_4/2$, while that across $T_{10}$ is $E + \Delta U_4$.

4 LRCD SNUBBER PARAMETER DESIGN

According to the analysis of the LRCD snubber, the turn-off spike voltages of IGBTs are determined by the induced voltages $\Delta U$ of the inductors $L_1$ and $L_2$. According to Equation (5), the induced voltage $\Delta U$ is directly proportional to the maximum current $I_{max}$ through the inductors. Since $C_L$ is much large

![Figure 12](image_url) Snubber process at $T_{10}$ turn-off. (a) $t_2 < t < t_3$, (b) $t_3 < t < t_4$, (c) $t_4 < t < t_5$
than $C_s$, $\Delta U$ is mainly clamped by parameters $L$, $R_L$, and $C_L$, as shown in Figure 14.

$R_{L1}$ and $R_{L2}$ are the energy-reset resistors which can help exhaust the energy stored in inductors. The energy stored in inductors should be fully released before the next switching cycle begins. According to Equation (4), when $I_L(T_i) = 0$, $T_i$ is:

$$T_i = \frac{R_L L C_s}{\ln \left( \frac{L - \sqrt{L^2 - 4R_L^2L C_s}}{L + \sqrt{L^2 - 4R_L^2L C_s}} \right)}$$  \hspace{1cm} (15)

where $C_s \approx C_L$. According to Equation (15), the relationship between $T_i$ and the snubber parameters is shown in Figure 15. The powers of $R_{L1}$ and $R_{L2}$ are:

$$P_{RL} = \frac{I^2 L}{4} f_L I_{L,peak}^2 + \frac{3}{4} C_s f_L E^2 + f_L I_{L,peak}^2 \sin^2(\phi) + \frac{3C_s f_L E^2}{2}$$  \hspace{1cm} (16)

where $I_{L,peak}$ is the peak value of $I_L$, $\phi$ is the power factor angle, $f_L$ is the switching frequency of IGBT, and $f_z$ is the fundamental frequency.

The parameters $L$ and $C_s$ determine the maximum current stress of IGBT. According to Equations (8), (11) and (14), the maximum current stress of IGBT is:

$$I_{pk} = I_{L,peak} + \frac{E}{\sqrt{3}} \sqrt{\frac{C_s}{2L}}$$  \hspace{1cm} (17)

According to Equation (17), choosing appropriate parameters $L$ and $C_s$ can reduce the current stress of IGBT. According to Equations (2, 7, 9) and (13), it can be seen that the maximum $\frac{dU}{dt}$ of IGBT is:

$$\frac{dU}{dt}\bigg|_{\text{max}} = \max \left\{ \frac{I_{L,peak} L C_s}{C_s}, E \sqrt{\frac{2}{3LC_s}} \right\}$$  \hspace{1cm} (18)

After IGBT is turned on, its snubber capacitor starts to be discharged. The minimum conduction time of IGBT shall be greater than five times of the discharging time constant of the snubber capacitor [26]. Therefore, the parameter $R_s$ should satisfy:

$$R_s < \frac{T_{\text{on(min)}}}{5C_s}$$  \hspace{1cm} (19)

where $T_{\text{on(min)}}$ is the minimum conduction time of IGBT. The powers of $R_{s1}$–$R_{s8}$ are:

$$P_{R_{s1}} = \frac{1}{2} f_z C_s E^2$$  \hspace{1cm} (20)

The powers of $R_{s9}$ and $R_{s10}$ are:

$$P_{R_{s9}} = \frac{1}{2} f_z C_s E^2$$  \hspace{1cm} (21)

The capacitance of $C_1$ can be calculated according to Equation (22):

$$C_1 = \frac{L_{\text{par}}}{f_z} (u_{\text{pi}} - E^2)$$  \hspace{1cm} (22)
TABLE 3  Experimental parameters

| Parameter         | Value  |
|-------------------|--------|
| DC-link voltage   | 200 V  |
| Peak value of rated load current | 10 A   |
| Capacitor $C_s$, $C_d$ and $C_f$ | 1.3 mF |
| R-L load         | 7 Ω 6 mH |
| Switching frequency $f_s$ | 2 kHz |
| Fundamental frequency $f_L$ | 50 Hz |
| Modulation index | 0.9    |

where $L_{par}$ is the parasitic inductance of the two-level half-bridge cell of the 5L-ANPC inverter, $u_{spi}$ is the spike voltages of $T_{11}$ and $T_{12}$.

In engineering application, the maximum overvoltage of switches should be determined according to the rated voltages of switches and the DC-link voltage of the 5L-ANPC inverter. Then select a suitable set of LRCD snubber parameters. It should be noted that the $T_i$ should be kept as small as possible to ensure that the response speed of the LRCD snubber meets the requirements.

5 EXPERIMENTAL RESULTS

A low-power three-phase 5L-ANPC prototype has been constructed to verify the proposed safe commutation strategy based on the LRCD snubber, as shown in Figure 16. The experimental parameters are shown in Table 3. The modulation algorithm used in this study is PS-PWM considering voltage balancing of the dc-link capacitors and the FCs [12]. The IGBTs used in the experimental prototype are Infineon IKW50N60T. Its rated voltage and rated current are 600 V and 50 A, respectively.

Since the dc-link voltage is 200 V, the rated voltage of each IGBT is 50 V. In this study, the spike voltage of single IGBT is set to be less than 1.2 times of its rated voltage, that is 60 V. According to the parameter design method in Section 4, the parameters of the LRCD snubber selected in this study are: $L = 1$ uH, $C_L = 10$ uF, $C_s = 0.22$ uF, $R_s = 2$ Ω, $R_L = 1$ Ω. The experimental results are shown in Figures 17–27.
The IGBTs are mounted on the printed circuit board. Therefore, we cannot directly measure the current of each IGBT with a current probe. According to Figure 1, the current stresses of IGBTs can be measured indirectly by measuring \( I_p \), \( I_o \), \( I_n \) and \( I_f \). For example, when the \( I_p \) is positive, it will flow through T1, T2 and T9, when the \( I_p \) is negative, it will flow through DF9, DF2 and DF1. Therefore, the current stresses of all the switches can be measured indirectly by measuring \( I_p \), \( I_o \), \( I_n \) and \( I_f \). Figure 17(a) shows the current waveforms of the 5L-ANPC inverter with RCD snubbers. The peak values of \( I_p \), \( I_o \), \( I_n \) and \( I_f \) are 77 A, 72.6 A, 65.5 A and 78 A, respectively. Figure 17(b) shows the current waveforms of the 5L-ANPC inverter with LRCD snubber. The peak values of \( I_p \), \( I_o \), \( I_n \) and \( I_f \) are 14.9 A, 12.1 A, 16.4 A and 29.8 A, respectively. The experimental results show that the RCD snubber will cause transient large currents due to the charging of the snubber capacitors, which is consistent with the analysis in Section 2. Compared with RCD snubber, the LRCD snubber can effectively reduce the current stresses of IGBTs.

When the switching frequency is 2 kHz, the inverter efficiency is shown in Figure 18(a). The efficiency of the 5L-ANPC inverter without any snubber at maximum output power (2.2 kW) is 94.93%, while those of the 5L-ANPC inverters with LRCD snubber and RCD snubber are respectively 91.85% and 89.39%. The experimental results show that the efficiency of the LRCD snubber is higher than that of the RCD snubber.

Figure 19 shows the voltage waveforms of the traditional commutation method when the snubber circuit is not applied. Near the zero-crossing point of the output voltage, the voltage stress of T9 or T10 is about three times of the rated voltage, which is consistent with the analysis in Section 2. The maximum voltage stress of T9 is about 148.5 V as shown in Figure 19(b). Figure 20 shows the experimental waveforms of the traditional commutation method when the LRCD snubber is applied. The maximum voltage stress of T9 is about 95.7 V, which is less than three times of the rated voltage due to the clamping effect of the snubber capacitor.

Figure 21 shows the experimental waveforms of the traditional commutation method when the RCD snubber is applied. The maximum voltage stress of T9 is about 104.6 V, which is less than three times of the rated voltage due to the clamping effect of the snubber capacitor. The experimental results show that the T9 and T10 will be breakdown when using the traditional commutation method. The snubber circuits cannot ensure T9 and T10 operate in SOA.

Figure 22 shows the voltage waveforms of the proposed safe commutation strategy. According to Figure 22(a), when the
conventional RCD snubber is applied, the maximum voltage stresses of \( T_1, T_3, T_5, T_7, T_9 \) and \( T_{10} \) are 56.8 V, 58.2 V, 59.7 V, 57.5 V, 77.6 V and 77.3 V, respectively. According to Figure 22(b), when the proposed LRCD snubber is applied, the maximum voltage stresses of \( T_1, T_3, T_5, T_7, T_9 \) and \( T_{10} \) are 53.6 V, 53.2 V, 55.8 V, 56.0 V, 59.0 V and 56.8 V, respectively. The maximum overvoltage of the proposed LRCD snubber is 11.16% of \( E \), which is less than the maximum design value and can be more reduced by the values of \( C_L \) and \( C_s \). The experimental results show that the proposed LRCD snubber can suppress the turn-off spike voltages better than the conventional RCD snubber. This is because the RCD snubber will increase the current stresses of switches, which will lead to the increase of the turn-off spike voltages. Most importantly, the experimental results verify the validity of the proposed commutation strategy. All the switches of the 5L-ANPC inverter operate in the SOA when the proposed commutation strategy based on the LRCD snubber is adopted.

When the output voltage changes from positive to negative, the experimental waveforms of the LRCD snubber are shown in Figure 23. The commutation process is divided into four stages. In stage 1, the switching state is \( \delta 2 \). The voltages across \( T_1, T_3, T_5, T_7, T_9 \) and \( T_{10} \) are 0.8 V, 50.8 V, 0.5 V, 50.6 V, 3.1 V and 43.6 V, respectively. In stage 2, \( T_1, T_2, T_5 \) and \( T_6 \) are turned off. The voltages across \( T_1 \) and \( T_5 \) increase approximately linearly, while that across \( T_3, T_7 \) and \( T_{10} \) decrease approximately linearly, which is consistent with the conclusion in Section 3.2. The spike voltages of \( T_1 \) is 51.8 V. In stage 3, \( T_3 \) and \( T_4 \) are turned on, and the voltage stresses of all the IGBTs remain unchanged. In stage 4, \( T_7 \) and \( T_8 \) are turned on. The voltages across \( T_3 \) and \( T_{10} \) start to increase, and the spike voltages of \( T_5 \) and \( T_{10} \) are 53 V and
FIGURE 24  Experimental waveforms of the LRCD snubber at T9 turn-off

FIGURE 25  Experimental waveforms of the LRCD snubber at T10 turn-off

55.8 V, respectively. It can be seen that T9 and T10 operate in SOA during whole the commutation process.

In summary, when the output voltage changes from positive to negative, the voltage waveforms of IGBTs are basically consistent with the theoretical analysis in Section 3.2. The spike voltages across T1–T10 are less than 60 V.

The experimental waveforms at T9 turn-off are shown in Figure 24. After T9 is turned off, the voltages across T9 increase approximately linearly while that across T10 decrease approximately linearly, which is consistent with the conclusion in Section 3.3. The spike voltages across T3 and T9 are 52.6 V and 56.5 V, respectively, during the whole commutation process. The voltage waveforms of T9 and T10 are basically consistent with the theoretical analysis in Section 3.3.

The experimental waveforms at T10 turn-off are shown in Figure 25. After T10 is turned off, Cs3 is discharged to 30.6 V due to the induced voltage of L1, more than the theoretical value of 25 V. This is because the power of the experimental inverter is small. The spike voltages of T3 and T10 are 51.3 V and 52.3 V, respectively, during the whole commutation process. The voltage waveforms of T3, T9 and T10 are basically consistent with the theoretical analysis in Section 3.4.

The voltage equalisation of capacitors is the precondition for stable operation of a multilevel converter. Figure 26 shows the phase voltage and capacitor voltages, where \( U_{fa}, U_{fb} \) and \( U_{fc} \) are FC voltages of phase a, b and c, respectively. The experimental results show that all the capacitor voltages can be maintained balanced within a limited margin of their reference values when the proposed LRCD snubber is applied.

Figure 27 shows the temperature curves of switches. According to the symmetry of the 5L-ANPC topology, the losses of T1, T3, T7 and T9 are equal, while those of switches T3–T6 and switches T9–T12 are also equal. So, Figure 27 only shows the temperature curves of T1, T3, T9 and their freewheeling diodes. The stable temperature of T1 is about 37.6°C, while those of the T3 and T9 are respectively 32.5°C and 49.3°C.

6 CONCLUSION

This study proposes a safe commutation strategy based on LRCD snubber for the 5L-ANPC inverter. By experimental results, its validity is verified. Putting all accounts together, the
proposed safe commutation strategy based on the LRCD snubber has the following features: (1) The proposed LRCD snubber can suppress the spike voltages across IGBTs of the 5L-ANPC inverter. (2) Compared with the RCD snubber, the proposed LRCD snubber can reduce current stresses of the switches, thus improving system efficiency. (3) The safe commutation strategy can prevent T9 and T10 from being overvoltage breakdown. It is considered that with proper designs of parameter conditions, the proposed commutation strategy based on the LRCD snubber can be used to medium-voltage and high-power 5L-ANPC inverters. The future work is to apply the proposed LRCD snubber to a medium-voltage 5L-ANPC inverter and study its effect on the switching transient characteristic of IGBT.

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