A Switched-Capacitor Inverter using Series/Parallel Conversion with Inductive Load

M. Swarnalatha, G. devadas, K. sunil

Abstract - A novel switched-capacitor inverter is proposed. The proposed inverter outputs larger voltage than the input voltage by switching the capacitors in series and in parallel. The maximum output voltage is determined by the number of the capacitors. The proposed inverter, which does not need any inductors, can be smaller than a conventional two-stage unit which consists of a boost converter and an inverter bridge. Its output harmonics are reduced compared to a conventional voltage source single phase full bridge inverter. In this paper, the circuit configuration, the theoretical operation, the simulation results with MATLAB/SIMULINK, and the experimental results are shown. The experimental results accorded with the theoretical calculation and the simulation results.

Index Terms—Charge pump, multicarrier PWM, multilevel inverter, switched capacitor (SC).

I. INTRODUCTION

Recently, electrical energy systems, electric vehicles (EVs), and dispersed generation (DG) systems, etc., are focused because of the global environmental issues. The power electronics, converters and inverters, is a key technology in these systems [1]–[6].

The EVs and the grid connected DG systems need an inverter to convert dc to ac. Boost converters or transformers are widely used in these systems when the input voltage is smaller than the output voltage. However, a transformer or an inductor in the boost converter makes the system large, because the transformer and the inductor must have large and heavy magnetic cores to sustain the high power [5]. As a provision against the issue, a charge pump, which does not have any inductors, is applied to such systems [7].

A charge pump outputs a larger voltage than the input voltage with switched capacitors [7], [8]. When the several capacitors and the input voltage sources are connected in parallel, the capacitors are charged. When the several capacitors and the input voltage sources are connected in series, the capacitors are discharged. The charge pump outputs the sum of the voltages of the capacitors and the input voltage sources. However, a charge pump has many switching devices which make the system more complicated. A switched-capacitor (SC) inverter outputs multilevel voltages with switched capacitors [9], [10]. An SC inverter is similar to a charge pump in the topology. The SC inverter outputs a larger voltage than the input voltage in similar way to the charge pump. However, the SC inverter also has many switching devices which make the system complicated. On the other hand, a Marx inverter, which has less switching devices compared to the SC inverter, was proposed [11].

Marx inverter can be regarded as one of the SC inverters because of its operation principle.

In this paper, an SC inverter whose structure is simpler than the conventional SC inverter is proposed. It consists of a Marx inverter structure and an H-bridge. The proposed inverter can output larger voltage than the input voltage by switching the capacitors in series and in parallel. The proposed inverter does not have any inductors which make the system large. The output harmonics of the proposed inverter are reduced by the multilevel output.

In Section II, the circuit topology is introduced and the driving method is explained. In Section III, the determination method of the capacitance is described. In Section IV, losses in the proposed inverter are calculated and estimated. In Sections V and VI, simulation results with MATLAB/SIMULINK and the results of the circuit experiments are shown.

II. CIRCUIT DESCRIPTION

Fig. 1 shows a circuit topology of the proposed inverter, where $S_{k1}$, $S_{k2}$, $S_{k3}$ ($k = 1, 2, \ldots, 2n - 2$) are the switching devices which switch the capacitors $C_k (k = 1, 2, \ldots, 2n - 1)$ in series and in parallel. Switches $S_1 - S_n$ are in the inverter bridge. A voltage source $V_{in}$ is the input voltage source. A low pass filter is composed of an inductor $L$ and a capacitor $C$.
There are many modulation methods to drive a multilevel inverter: the space vector modulation [3], [12]–[14], the multiscarrier pulse width modulation (PWM) [3], [15], [16], the hybrid modulation [1], [3], [4], [17], the selective harmonic elimination [3], [18], [19], and the nearest level control [3]. In this paper, the multiscarrier PWM method is applied to the proposed inverter.

Fig. 2 shows the current flow in the proposed inverter \((n = 2)\) and Fig. 3 shows the modulation method of the proposed inverter \((n = 2)\). When the time \(t\) satisfies \(0 \leq t < t_1\) in Fig. 3, the switches \(S_1\) and \(S_2\) are driven by the gate-source voltage \(v_{GS1}\) and \(v_{GS2}\), respectively. While the switches \(S_1\) and \(S_2\) are switched alternately, the other switches are maintained ON or OFF state as shown in Fig. 3. Therefore, the states shown in Fig. 2(a) and (b) are switched alternately and the bus voltage \(v_{bus}\) takes 0 or \(V_{in}\).

When the time \(t\) satisfies \(t_1 \leq t < t_2\) in Fig. 3, the switches \(S_{a1}\), \(S_{b1}\), and \(S_{c1}\) are driven by the gate-source voltage \(v_{GSa1}\), \(v_{GSb1}\), and \(v_{GSc1}\), respectively. While the switches \(S_{a1}\), \(S_{b1}\), and \(S_{c1}\) are switched alternately, the other switches are maintained ON or OFF state as shown in Fig. 3. Therefore, the states shown in Fig. 2(c) and (d) are switched alternately. The capacitor \(C_3\) is charged by the current \(-i_{C3}\) as shown in Fig. 2(c) during the state shown in Fig. 2(c). Therefore, the proposed inverter outputs \(v_{bus}\) alternately in this term.

\[
v_{bus} = V_{in} + V_{C1} \tag{1}
\]

Where \(V_{C1}\) is the voltage of the capacitor \(C_1\). Therefore, the proposed inverter outputs \(v_{bus}\) or \(V_{in} + V_{C1}\) alternately in this term.
\[ V_{bus} = V_{in} + V_{C1} + V_{C3} \]  \hspace{1cm} (2)

Where \( V_{C3} \) is the voltage of the capacitor \( C3 \). Therefore, the proposed inverter outputs \( V_{in} + V_{C1} \) or \( V_{in} + V_{C1} + V_{C3} \) alternately in this term. After \( t = t3 \), the four states shown in Fig. 2 are repeated by turns.

Table I shows the list of the on-state switches when the proposed inverter \((n = 2)\) is driven by the modulation method shown in Fig. 3. The ideal bus voltage \( V_{bus} \) in Table I means the bus voltage on each state when \( V_{C1} = V_{C3} = V_{in} \) is assumed. As the conventional SC inverter, the proposed inverter has a full bridge which is connected to the high voltage. Therefore, the device stress of the switches \( S1 - S4 \) in the full bridge is higher than the other switches as the conventional SC inverter.

<table>
<thead>
<tr>
<th>Relation between ( e_k ) &amp; ( e_s )</th>
<th>On state switches</th>
<th>Ideal bus voltage ( V_{bus} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>( e_0 \geq e_1 )</td>
<td>( S_1, S_2, S_3, S_4 )</td>
<td>( 3V_{in} )</td>
</tr>
<tr>
<td>( e_2 \geq e_1 )</td>
<td>( S_1, S_2, S_3, S_4 )</td>
<td>( 2V_{in} )</td>
</tr>
<tr>
<td>( e_0 \geq e_2 )</td>
<td>( S_1, S_2, S_3, S_4 )</td>
<td>( V_{in} )</td>
</tr>
<tr>
<td>( e_2 \geq e_3 )</td>
<td>( S_2, S_3, S_4, S_1 )</td>
<td>( 0 )</td>
</tr>
<tr>
<td>( e_2 \geq e_4 )</td>
<td>( S_2, S_3, S_4, S_1 )</td>
<td>( -V_{in} )</td>
</tr>
<tr>
<td>( e_0 \geq e_4 )</td>
<td>( S_1, S_2, S_3, S_4 )</td>
<td>( -3V_{in} )</td>
</tr>
<tr>
<td>( e_0 \geq e_s )</td>
<td>( S_1, S_2, S_3, S_4 )</td>
<td>( -V_{in} )</td>
</tr>
</tbody>
</table>

The proposed inverter \((n=2)\) outputs a 7-level voltage by repeating the four states as shown in Fig. 2. Because the driving waveform \( v_{gsa1} \) and \( v_{gsa2} \) change alternately as shown in Fig. 3, the capacitors \( C1 \) and \( C3 \) are equally discharged. Assuming that the number of the capacitors is \( 2n-1 \), the proposed inverter can outputs \( 4n-1 \) levels voltage waveform.

The modulation index \( M \) is defined as the following equation because the amplitude of the output voltage waveform is inversely proportional to the double amplitude of the carrier waveform.

\[ M = \frac{A_{ref}}{2A_c} \]  \hspace{1cm} (3)

In (3), \( A_{ref} \) is the amplitude of the reference waveform and \( A_c \) is the amplitude of the carrier waveform.

The proposed inverter requires 10 switching devices for the 7-level, and 16 switching devices for the 11-level. On the other hand, the conventional SC inverter requires 20 switching devices for the 7-level, and 28 switching devices for the 11-level [9]. The conventional cascaded H-bridge (CHB) inverter requires 12 switching devices for the 7-level, and 20 switching devices for the 11-level, when all the dc voltage sources take the same voltage [17]. Therefore, the proposed inverter has less number of switching devices than the conventional multilevel inverters.

### III. DETERMINATION OF CAPACITANCE

The capacitance \( C_k \) can be determined properly with considering the voltage ripple of the capacitors \( C_k \). The smaller voltage ripple of these capacitors leads to the higher efficiency. In this section, the capacitance \( C_k \) are calculated when the maximum voltage ripple is supposed to be 10% of the maximum voltages of the capacitors.

The capacitors \( C_k \) are charged when they are connected in parallel and are discharged when they are connected in series. From Fig. 3, the switches \( S_{a1} \) and \( S_{a2} \) of the proposed inverter \((n = 2)\) are symmetrically driven during the half cycle of the reference waveform. Therefore, the voltage ripple of the capacitor \( C_1 \) is focused.

Assuming that the power factor of the output load \( \cos \phi = 1 \), the longest discharging term of the capacitor \( C_1 \) in the proposed inverter \((n = 2)\) is between \( t2 \) and \( t3 \) in Fig. 3. Assuming the modulation index \( M = 3 \), the time \( t1, t2 \) and \( t3 \) in Fig. 3 are

\[ t1 = \frac{\sin^{-1}(1/3)}{2\pi f_{ref}} \]  \hspace{1cm} (4)

\[ t2 = \frac{\sin^{-1}(2/3)}{2\pi f_{ref}} \]  \hspace{1cm} (5)

\[ t3 = \frac{\pi - \sin^{-1}(2/3)}{2\pi f_{ref}} \]  \hspace{1cm} (6)

where \( f_{ref} \) is the frequency of the reference waveform. Therefore, the maximum discharge amount \( \Phi \) of the capacitor \( C_1 \) is

\[ Q_1 = \int_{t2}^{t3} I_{bus} \sin(2\pi f_{ref}t - \phi) \, dt \]  \hspace{1cm} (7)

where \( I_{bus} \) is the amplitude of the bus current waveform and \( \phi \) is the phase difference between the bus voltage waveform \( v_{bus} \) and the bus current waveform \( I_{bus} \). \( \Phi \) supposes to be less than 10% of the maximum charge of \( C_1 \). Therefore, the capacitance \( C_1 \) must satisfy

\[ C_1 > \frac{Q_1}{0.1V_{in}} \]  \hspace{1cm} (8)

When the capacitors \( C_k \) satisfy (8), the other voltage ripple which is caused by PWM is less than 10%. The peak current of the capacitor \( I_{C1} \) is calculated by

\[ I_{C1} = \frac{V_{in} - V_{C1}}{r_{C1} + 2r_{on}} \]  \hspace{1cm} (9)

where \( r_{on} \) is the equivalent series resistance (ESR) of the capacitor \( C_1 \) and \( r_{on} \) is the internal resistance of the switching devices. From (9), the peak current of the capacitor \( C_1 \) is determined by the difference between the input voltage \( V_{in} \) and the voltage of the capacitor \( V_{C1} \), and the internal resistance of the switching devices. The difference of the voltages \( V_{in} - V_{C1} \) is small when the capacitance \( C_1 \) is large. Therefore, when the switches which have small internal resistance are used the capacitance \( C_1 \) must be larger to prevent the large peak current.
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When the phase difference $\phi$ satisfies $0 < \phi < \sin^{-1}(2/3)$ i.e., $0 < \cos \phi < 1$, the current flows as shown in Fig. 4(a) and (b) before $t = \phi/2 \pi f_{\text{ref}}$. Therefore, the capacitor $C_1$ is charged by the reverse current and the voltage of the capacitor $C_1$ is increased in the state when $C_1$ is connected in series as shown in Fig. 4(b). The charge amount $Q_1$ of the capacitor $C_1$ in the state is calculated by

$$Q_1 = -\int_{t_1}^{t_2} D_{\text{sat}}(t) I_{\text{bus}} \sin(2\pi f_{\text{ref}} t - \phi) \, dt$$  \hspace{1cm} (10)$$

where $D_{\text{sat}}(t)$ is the duty ratio of the switch $Sat$. From Fig. 3, $D_{\text{sat}}(t)$ is sinusoidal function between $t_1$ and $t_2$. In addition, $D_{\text{sat}}(t_1) = 0$ and $D_{\text{sat}}(t_2) = 1$. Therefore, $D_{\text{sat}}(t)$ is calculated by

$$D_{\text{sat}}(t) = 3 \sin(2\pi f_{\text{ref}} t) - 1.$$  \hspace{1cm} (11)$$

From (7) and (10), the maximum discharge amount $Q_1$ is larger than the charge amount $Q_1$. Therefore, the voltage ripple of the capacitor $C_1$ is determined by $Q_1$ when $0 < \cos \phi < 1$.

When the power factor $\cos \phi$ satisfies $\cos \phi \leq 0.745$, there is the term when the current direction becomes reverse in all states of the switching devices as shown in Fig. 4. Therefore, the maximum discharge amount $Q$ is calculated by

$$Q_1 - \int_{t_1}^{t_2} I_{\text{bus}} \sin(2\pi f_{\text{ref}} t - \phi) \, dt.$$  \hspace{1cm} (12)$$

From (7) and (12), the maximum discharge amount $Q_1$ is reduced. However, $Q_1$ is larger than the charge amount $Q_1$ because the input current is larger than the reverse current with an inductive load. Therefore, voltage ripple of the capacitor $C_1$ is also determined by $Q_1$ when the power factor $\cos \phi \leq 0.745$. The maximum discharge amount $Q_1$ takes the largest value when $\cos \phi = 1$ because the peak current is accorded to the peak voltage. Hence, when the capacitance $C_1$ is determined for $\cos \phi = 1$, the proposed inverter can maintain the output waveform for $\cos \phi < 1$.

IV. CALCULATION OF LOSSES

In this section, the power losses of the proposed inverter ($n=2$) are calculated. In the calculation, the following losses are considered:

- switching losses;
- conduction losses of the switches;
- conduction loss of the output filter;
- conduction losses and losses caused by the voltage ripple of the capacitors $\omega C$.

These losses are calculated about the proposed inverter ($n=2$).

A. Switching Losses

In this section, switching losses are calculated from the charge and the discharge of the parasitic capacitance [20]. From Fig. 3, the switches $S_1$ and $S_2$ are switched ON/OFF at the carrier frequency $f$ when the reference waveform $e_s$ satisfies

$$|e_s| < \frac{1}{M} A_{\text{ref}}.$$  \hspace{1cm} (13)$$

Therefore, the switches $S_1$ and $S_2$ are switched ON/OFF when the time $t$ satisfies $0 < t < t_1$ or $t_4 < t < t_5$ in Fig. 3. When the reference waveform $e_s$ does not satisfy (13), the switches $S_1$ and $S_2$ are maintained ON or OFF. Therefore, when the carrier waveforms and the reference waveform are not synchronous waveforms, the average number of switching transitions $N_{s1}$ and $N_{s2}$ in one period of reference waveform is calculated as

$$N_{s1} = N_{s2} = \frac{t_1 + (t_5 - t_4)}{t_5} \cdot \frac{2 f}{f_{\text{ref}}}.$$  \hspace{1cm} (14)$$

where $f$ is the switching frequency. When the carrier waveforms and the reference waveform are synchronous waveforms, the number of switching transitions is the maximum even number less than $N_{s1}$ or $N_{s1} + 4$. The number of switching transitions is variable depending on the frequency modulation ratio $f / f_{\text{ref}}$. Therefore, when the synchronous waveforms are used in this modulation, the individual analysis is needed in each frequency modulation ratio.
In this paper, the switching losses are calculated with the average number of switching transitions \(N\), because the high switching frequency \(f\) is assumed. For one second, this switching terms repeat \(f\) times. Therefore, the switches \(S_1\) and \(S_2\) are switched \(N_{S_1,ref}\) and \(N_{S_2,ref}\) times in one second, respectively. The same argument can be applied to the switches \(S_3\) and \(S_4\). The energy loss \(E_{loss}\) as the following equation occurs in one switching [20]

\[ E_{loss} = \frac{1}{2} C_s V_{in}^2 \]  

(15)

where \(C_s\) is the parasitic capacitance of the switching devices. Therefore, the switching losses of these switches \(P_{loss}(k = 1, 2, 3, 4)\) are given as

\[ P_{loss} = \frac{2 \sin^{-1}(1/M)}{2\pi} . C_s V_{in}^2 f. \]  

(16)

The switches \(S_{ak}, S_{bk},\) and \(S_{ck}\) \((k = 1, 2)\) are operated alternately in each cycle of the reference waveform. Therefore, the switching losses of these switches can be calculated with considering the half cycle of the reference waveform. If the voltage ripple of the capacitors \(\Delta V_{c}\) is ignored, i.e., the voltage of the capacitors \(V_{c}\) is assumed, the switching losses of the switches \(S_{ak}, S_{bk},\) and \(S_{ck}\) \((k = 1, 2)\) are calculated as the following equations by the same calculation method to the (16)

\[ P_{loss} = \frac{2 \sin^{-1}(1/M) - \sin^{-1}(1/M)}{2\pi} . C_s V_{in}^2 f. \]  

(17)

\[ P_{loss} = \frac{\pi - 2 \sin^{-1}(1/M)}{2\pi} . C_s V_{in}^2 f. \]  

(18)

All switching devices are switched when the voltage of each device is \(V_{in}\). Therefore, the switching losses of the proposed inverter are smaller than the conventional voltage source full bridge inverter.

B. Conduction Losses of the Switches

There are three states on the proposed inverter \((n = 2)\) as shown in Fig. 2; the state when all capacitors are connected in parallel, the state when one of the capacitors is connected in series, and the state when all capacitors are connected in series. However, it is obvious that the bus current \(i_{bus}\) flows in 4 switches on each state from Fig. 2. Therefore, the total conduction loss \(P_{con}\) of the switches is calculated by

\[ P_{con} = 4 \cdot \frac{2\pi f_{ref}}{\pi} \int_{0}^{\frac{\pi}{2}} r_m \delta t_{bus} \ dt \]  

(19)

where \(r_m\) is the internal resistance of the switching devices. On the other hand, the current flows in 6 switches in the conventional 7-level CHB inverter because the current flows in 2 switches in each H-bridge. Therefore, the total conduction loss of the switches in the conventional 7-level CHB inverter \(P_{con}\) is calculated as

\[ P_{con} = 6 \cdot \frac{2\pi f_{ref}}{\pi} \int_{0}^{\frac{\pi}{2}} r_m \delta t_{bus} \ dt. \]  

(20)

From (19) and (20), it is obvious that the conduction losses of the proposed inverter are less than the conventional 7-level CHB inverter when the same switching devices are used.

C. Conduction Loss of the Output Filter

The conduction losses of the filter inductance \(P_l\) and the filter capacitance \(P_c\) are calculated as the following equations:

\[ P_l = r_1 l_{bus}^2 \]  

(21)

\[ P_c = r_c C^2. \]  

(22)

In (21), \(r_1\) is the ESR of the filter inductance \(L\). In (22), \(r_c\) and \(C\) are the ESR and the capacitance of the filter inductance \(C\).

D. Losses of the Capacitors \(\Delta V_{c}\)

When the capacitors \(C(k = 2)\) are connected in parallel, the losses occur by the difference between the input voltage \(V_{in}\) and the voltages of the capacitors \(V_{c}\). The voltage ripple of the capacitors \(\Delta V_c\) is calculated by

\[ \Delta V_{c} = \frac{1}{C_k} \int_{0}^{t} \frac{\delta V_{c}}{0}. \]  

(23)

where \(\delta V_{c}\) is the current of the capacitor \(C_k\) and \(r_2\) are the time when the capacitors \(C_k\) are connected in series as shown in Fig. 3. Therefore, the losses \(P_{cap}\) by this voltage ripple are calculated as the following equation.

\[ P_{cap} = \sum_{k=1}^{2n-1} C_k \Delta V_{c}^2 f_{ref}. \]  

(24)
From (23) and (24), the loss $P_{dc}$ is inversely proportional to the capacitance $C_k$, which means the larger capacitance leads to the higher efficiency. When the capacitors $C_k$ are connected in series, the losses occur by the internal resistance $r_{sc}$. The conduction losses of these capacitors $P_{ac}$ are calculated by the following equation:

$$P_{ac} = \frac{2\pi f_{ref}}{\pi} \sum_{k=1}^{n} \int_{\frac{k-1}{2n}}^{\frac{k}{2n}} r_{sc} C_k^2 \, dt.$$  \hspace{1cm} (25)

V. SIMULATION RESULTS

Simulation was performed under the two conditions. One was for a low power inverter under the same condition with the circuit experiments. The other one was for a high power inverter.

The low power inverter simulation was performed with MATLAB/SIMULINK ver. R2009a under the following conditions. The MOSFET models whose internal resistance $R_{on} = 0.54 \, [\Omega]$ and the snubber resistance $R_s = 105 \, [\Omega]$ were used as the switching devices. The input voltage $V_{in} = 8.00 \, [V]$, the output resistance $R = 50.0 \, [\Omega]$, a filter capacitance $C$, and a filter inductance $L$ were $C = 0.450 \, [\mu F]$ and $L = 1.13 \, [mH]$, the modulation index $M = 3.00$, the switching frequency $f_s = 40.0 \, [kHz]$, and the reference waveform frequency $f_{ref} = 1.00 \, [kHz]$. From (8), the capacitance $C_1$ and $C_3$ are calculated to $C_1 = C_3 = 143 \, [\mu F]$, which have ESRs $r_{sc} = 800 \, [m\Omega]$.

Fig. 5 shows the simulated voltage waveforms of the proposed inverter ($n = 2$) designed for low power at 5.76 [W].

The voltages of the capacitors $V_{cap}$ are changed between 6.73 [V] and 7.51 [V]. Therefore, the voltage of the step in the bus voltage waveform decreases to about 90% as shown at $t = 8.11$ to 8.40 [ms] in Fig. 5(a). It is caused by the 10% voltage drop at the capacitors $C_k$. This is accorded with the theoretical calculation. The theoretical amplitude of the output waveform is

$$MVI = 24 \, [V].$$  \hspace{1cm} (26)

From Fig. 5(b) and (26), it is confirmed that the amplitude of the output waveform in the simulation is smaller than the theoretical amplitude. It is also caused by the voltage reduction of the capacitors $C_k$.

The high power inverter simulation was performed under the following conditions. The IGBT/Diode models whose internal resistance $R_{on} = 65.0 \, [m\Omega]$ and the snubber resistance $R_s = 105 \, [\Omega]$ were used as the switching devices. $V_{in} = 100 \, [V]$, $R = 10.0 \, [\Omega]$, $C = 2.5 \, [\mu F]$, $L = 225 \, [\mu H]$, $M = 3.00$, $f_s = 40.0 \, [kHz]$ and $f_{ref} = 1.00 \, [kHz]$. From (8), $C_1$ and $C_3$ are calculated to $C_1 = C_3 = 712 \, [\mu F]$. These capacitors ESRs $r_{sc}$ are $800 \, [m\Omega]$.

Fig. 6 shows the simulated voltage waveforms of the proposed inverter ($n = 2$) designed for high power at 4.50 [kW], switching frequency $f_s = 40 \, [kHz]$ and reference waveform frequency $f_{ref} = 1 \, [kHz]$. (a) Bus voltage waveform $V_{bus}$ and (b) the output voltage waveform $V_{out}$.
In both spectra of the bus voltage waveforms of the low power and the high power proposed inverters, the magnitude at 40 [kHz] is larger than the other frequency components. This is caused by the switching frequency \( f = 40 \) [kHz]. The calculated total harmonic distortions (THDs) of the 5.76 [W] inverter and of the 4.50 [kW] inverter are 18.9% and 18.7%. On the other hand, the THD of the conventional voltage source single phase full bridge inverter is 64.3% under the same condition of the low power simulation. The THDs of the proposed inverters are reduced compared to the conventional single phase full bridge inverter.

VII. CONCLUSION

In this paper, a novel boost switched-capacitor inverter was proposed. The circuit topology was introduced. The modulation method, the determination method of the capacitance, and the loss calculation of the proposed inverter were shown. The circuit operation of the proposed inverter was confirmed by the simulation results and the experimental results with a resistive load and an inductive load.

The proposed inverter outputs a larger voltage than the input voltage by switching the capacitors in series and in parallel. The inverter can operate with an inductive load. The structure of the inverter is simpler than the conventional switched-capacitor inverters. THD of the output waveform of the inverter is reduced compared to the conventional single phase full bridge inverter as the conventional multilevel inverter.

REFERENCES