REDUCTION OF SWITCHING LOSS IN GRID-CONNECTED INVERTERS USING A VARIABLE SWITCHING CYCLE

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ABSTRACT
Distributed generations (DGs) converting renewable energy sources such as solar, wind power, and biomass into grid systems have become very popular in the world recently. The DG using power inverter injects significant harmonics into the grid and can cause instability of the system. Thus, the stringent grid standards are imposed by utility companies to maintain the grid stability. These lead to requirement of reducing current harmonics of grid-connected inverters to achieve compliance with such grid codes. Increasing the switching frequency of the sinusoidal pulse-width modulation (SPWM) of inverters is a common method for reducing the total harmonic distortion (THD) of the current; however, this increases switching losses. This paper is to propose an SPWM technique, with a variable switching cycle in each half of the fundamental period, for reducing the switching losses in grid-connected inverters without increasing current harmonics. The simulation results also validate the performance of the proposed technique compared to those of the total demand distortion based technique and the constant ripple technique.

Key words: Electromagnetic Interference (EMI), Sinusoidal Pulse-Width Modulation (SPWM), Switching Loss, Total Demand Distortion (TDD), Total Harmonic Distortion (THD).

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1. INTRODUCTION
The application of DGs using renewable energy leads to a rapid development of grid-connected inverter systems [1] toward achieving sustainable development with its enormous potential [2]. However, grid-connected inverters insert significant current harmonics into the power network and adversely affect the power quality of the
system. Thus, to meet the stringent IEEE standards 929-2000 [3] and 1547-2009 [4], [5], harmonic reduction is considered as a primary task of engineers who design inverter systems.

Grid-connected inverters using SPWM are widely used in renewable energy converters [6]-[9]. Increasing an inductance of the filter is one of the popular methods for reducing the output current harmonics; however, the size and cost of the inverters increase as a result. Increasing the switching frequency is another method but results in higher switching loss and the overheating of components [10].

An SPWM-based variable switching frequency technique, proposed in [11] for reducing the switching loss of inverter, requires an accurate model of the ripple current; additionally, the complicated calculations result in issues with robustness and a low dynamic response. Moreover, the high ripple caused by the very low switching frequency of the near zero current is a big drawback to digital electronic devices and electric motors. Besides, a very high range of switching frequencies (16-90 kHz) is not suitable for the real power semiconductor switches of grid-connected inverters due to the limit of maximum switching frequency. Further, the performance in a grid-connected system was also not considered in this work. Another technique using a variable switching frequency, proposed in [12], is based on the estimated model of the TDD. However, the filter parameter requirement and the computation time render the performance of the method as poor in terms of the dynamic response and robustness.

A method in [13] uses an SPWM technique in which selective harmonics are injected into the modulating signal to provide a large amount of sinusoidal information in areas of greater sampling. Then, the modulating wave is compared to a triangular carrier with a variable frequency over a period of the modulator. However, the switching loss has not considered in this method. It thus is impossible to optimize the grid-connected inverters.

Multilevel inverters in [14]-[15] are also adopted to reduce the harmonic content, but controlling them is complicated during modulation. In addition, the excessive numbers of power switches and DC sources increase the cost, and such equipment suffers from problems related to capacitor voltage balance, etc. A technique proposed in [16] to reduce the switching loss and the current THD depends on the measured current errors and current sensors. Therefore, this method also has poor robustness.

Moreover, variable switching frequency PWM methods are usually not analyzed quantitatively or rigorously in terms of the switching loss compared to constant switching frequency PWM [17]-[19]. To maintain a constant root-mean-square (rms) ripple current and to reduce the EMI noise, the switching frequency of every sector of the space vector PWM (SVPWM) in [18] is redistributed in a nonlinearity based on the current ripple prediction. In contrast, that of [19] is linearly increased and decreased in each half of a sector to spread an acoustic noise spectrum and reduce magnitudes of the dominant noise components. However, the switching loss and application for SPWM have not been considered quantitatively.

Recently, as alternative methods instead of solving nonlinear transcendental equations, many heuristic methods [20], such as particle swarm optimization (PSO) [21]-[22], ant colony optimization (ACO) [23], an artificial bee colony algorithm [24], and a genetic algorithm [25] have been proposed for eliminating selective harmonics in SPWM; however, the switching loss has not been considered in these methods.

This paper proposes an SPWM technique, with a variable switching cycle, for reducing the switching loss of inverters. In the proposed technique, the optimal switching cycle of the inverter in every half of a fundamental period is determined
improving a spread of noise spectrum in [19] for reducing the switching loss under the constraint of current THD being equal to that of the constant switching cycle method. The results of the implementation of a grid-connected inverter system using the proposed method are also compared to those of the TDD method in [12] and the constant ripple method (the hysteresis control) in [26]-[27].

2. CURRENT RIPPLE AND SWITCHING LOSS ANALYSIS

Both the current THD and switching loss of the inverter depend on the switching frequency. In conventional SPWM techniques with a constant switching frequency, a higher switching frequency leads to a higher switching loss and a lower current THD and vice versa [28]. Therefore, selecting the optimal switching frequency to reduce the switching loss without increasing the current THD of the inverter is a difficult problem of SPWM.

To test the proposed technique, an H-bridge, grid-connected, single-phase inverter with unipolar SPWM as in Fig. 1 is used. To analyze the current ripples, the assumptions are made: the switching frequency of the inverter is much higher than that of the modulated signal and the effect of dead time is negligible. The losses of the IGBTs and diodes produce conduction, switching, and other losses. It is also assumed that the conduction loss is not dependent on the switching frequency of the inverter and that the switching loss linearly depends on the switched instantaneous current and switching frequency of one switching cycle.

![Figure 1 H-bridge grid-connected inverter.](image)

The inverter output current includes the fundamental current and the ripple current based on the superposition principle.

The output current of the inverter increases in every positive half of the switching cycle of the carrier wave and decreases in the negative half, as seen from the waveforms in Fig. 2. In the positive half cycle of the carrier cycle, the increase in the peak-to-peak current ripple \( i_{L1} \) can be calculated as

\[
i_{L1} = \left[ 1 - d(t) \right] \frac{d(t)}{L_f} T_s \frac{V_{dc}}{2}
\]

(1)

where \( L_f \) is the inductance of the output filter of the inverter, \( V_{dc} \) is the DC input voltage value of the inverter, \( d(t) \) is the duty cycle, and \( T_s \) is the cycle of the carrier wave.

The decrease in the current ripple \( i_{L2} \) can be similarly calculated as

\[
i_{L2} = \left[ 1 + d(t) \right] \frac{d(t)}{L_f} T_s \frac{V_{dc}}{2}
\]

(2)

Adding (1) and (2) for both the positive and negative half cycles of \( d(t) \) yields the peak-to-peak current ripple as
The output current waveform of the unipolar, H-bridge, single-phase inverter is shown in Fig. 3.

\[
\Delta I_p = \left[ 1 - \left| d(t) \right| \right] \frac{V_{dc}}{2 L_f} T_s
\]  

(3)

Figure 2 Carrier wave, voltage ripple, and ripple current.

The rms value of the current ripple in each switching cycle can be defined as

\[
\Delta I_p = \frac{\Delta I_p}{\sqrt{3}} = T_s V_{dc} \frac{1 - \left| d(t) \right|}{2 \sqrt{3} L_f} d(t)
\]  

(4)

Assuming that the phase angle is ignored in this analysis. Then, the modulated signal is expressed as

\[d(t) = m \sin(\omega t)\]  

(5)

where \(m\) is the modulation index of the amplitude and \(\omega\) is the angular frequency of the grid source.

Substituting (5) for \(d(t)\) into (4) yields the rms value of current ripple as

\[
\Delta I_p = \frac{T_s V_{dc}}{L_f 2 \sqrt{3}} \left[ 1 - m \sin(\omega t) \right] \left| m \sin(\omega t) \right|
\]  

(6)

The rms value of the current ripple in half of the fundamental period is calculated as

\[
\Delta I = \sqrt{\frac{1}{\pi} \int_0^{\pi} \Delta I_p^2 d(\omega t)}
\]  

(7)

Figure 3 Output current of inverter (m=0.97). (a) Output current and fundamental current. (b) Current ripple.
The current THD has the following relation to the rms current ripple:

\[ THD = \frac{\Delta I}{I_1} \]  

(8)

where \( I_1 \) is the rms value of the fundamental current.

For a constant frequency carrier wave, the switching loss is expressed as

\[ \Delta P_{sw} = C_1 |i(t)| \frac{1}{T_s} \]  

(9)

where the constant \( C_1 \) depends on the DC voltage \( V_{dc} \) and \( i(t) \) is the value of the instantaneous current flowing in the power switches. The average switching loss in half of the fundamental period is considered to depend linearly on the switched instantaneous fundamental current \( i_1(t) \) and switching frequency in one switching cycle. Then, the average switching loss in half of the fundamental period \( \Delta P_{sw} \) can be approximated as

\[ \Delta P_{sw} = C_1 I_1 \sqrt{2} \frac{\int_0^\pi |\sin(\alpha)| d(\alpha)}{T_s} \]  

(10)

3. THE PROPOSED TECHNIQUE

As mentioned above, the spread of acoustic noise spectrum in [19] for SVPWM is used to reduce amplitude of individual harmonics. The switching cycle linearly increases and decreases in each half of sector of SVPWM as (11) and Fig. 4, where \( T_s \) is the fixed switching cycle. The switching frequency changes between two-third and two of the fixed frequency in each half of sector. Fraction \( k \) is proposed to be constant as 0.5 for all sectors.

\[ T_s(\alpha) = \begin{cases} T_s \left[ 1 + k \left( 1 - \frac{12\alpha}{\pi} \right) \right] , & 0 \leq \alpha \leq \frac{\pi}{6} \\ T_s \left[ 1 - k \left( 3 - \frac{12\alpha}{\pi} \right) \right] , & \frac{\pi}{6} < \alpha \leq \frac{\pi}{3} \end{cases} \]  

(11)

Figure 4 Switching cycle change in every sector (\( T_s=50 \mu s \)).

However, this method has not been quantitatively considered in term of switching loss of a grid-connected inverter system. In addition, it is applied for SVPWM, not for SPWM.

Based on the ripple description in (6), this paper proposes to improve the variable switching cycle \( T_{ss} \) in [19] to be suitable for the SPWM as given by (12). An illustration for the constant frequency of 1 kHz and the fraction \( C_3=0.5 \) is shown in Fig. 5.
In (12), $T_{sc}$ is a constant switching cycle, and $C_3$ is a fraction. To improve the performance, both $T_{sc}$ and $C_3$ are adjusted to match the load level to achieve a current THD similar to that of the TDD method.

\[
T_{ss}(t) = \begin{cases} 
T_{sc} \left[ 1 + C_3 \left( 1 - \frac{12t}{\pi} \right) \right] & 0 \leq t \leq \frac{\pi}{6} \\
T_{sc} \left[ 1 - C_3 \left( 3 - \frac{12t}{\pi} \right) \right] & \frac{2\pi}{6} \leq t \leq \frac{4\pi}{6} \\
T_{sc} \left[ 1 + C_3 \left( 1 - \frac{12t}{\pi} \right) \right] & \frac{5\pi}{6} \leq t \leq \frac{\pi}{6} 
\end{cases}
\]

(12)

**Figure 5** Switching cycle distribution in half the fundamental period

($T_{sc}=1000 \mu s; C_3=0.5$).

(a) Cycle. (b) Frequency.

4. SIMULATION RESULTS

The grid-connected inverter system of Fig. 1 is implemented in MATLAB/Simulink with the parameters in Table I and the control diagram in Fig. 6. The active power reference $P_{ref}$ is stepwise varied from 3.0 kW to 1.5 kW at time $t=0.2$ s. The reactive power reference $Q_{ref}$ is also stepwise varied 0.0 kVar to 1.0 kVar at time $t=0.35$ s.

The constant $C_1$ is defined as $2.49433\times10^{-4}$ according to determination of [29].

**TABLE I THE PARAMETERS OF THE SYSTEM**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Value</th>
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</thead>
<tbody>
<tr>
<td>Inductance of filter</td>
<td>$L_f$</td>
<td>2.5 mH</td>
</tr>
<tr>
<td>Resistance of $L_f$</td>
<td>$R_f$</td>
<td>0.2 $\Omega$</td>
</tr>
<tr>
<td>Inductance of grid source</td>
<td>$L_g$</td>
<td>0.01 mH</td>
</tr>
<tr>
<td>Resistance of $L_g$</td>
<td>$R_g$</td>
<td>0.01 $\Omega$</td>
</tr>
<tr>
<td>DC voltage value</td>
<td>$V_{dc}$</td>
<td>350 V</td>
</tr>
<tr>
<td>Grid source voltage</td>
<td>$V_{ac}$</td>
<td>220 V</td>
</tr>
<tr>
<td>Constant</td>
<td>$C_1$</td>
<td>$2.494\times10^{-4}$</td>
</tr>
<tr>
<td>Capacitor of filter</td>
<td>$C_f$</td>
<td>1 $\mu$F</td>
</tr>
<tr>
<td>Fundamental frequency</td>
<td>$f$</td>
<td>50 Hz</td>
</tr>
</tbody>
</table>
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Figure 6 The control circuit diagram

1. The Fixed Switching Frequency of 10 kHz

Figure 7 Responses of the active and reactive powers.

Figure 8 Responses of the grid voltage and current.
Figure 9 Switching loss and THD for the fixed cycle technique. (a) Instantaneous and average switching loss. (b) Current THD.

2. The TDD-Based Variable Switching Frequency

Figure 10 The TDD technique. (a) Instantaneous and average switching loss. (b) Current THD.

Fig. 11. The THD of the TDD.
(a) For t<0.2 s. (b) For 0.2≤t<0.35 s. (c) For 0.35≤t<0.5 s.

Fig. 12. The constant ripple THD.
(a) For t<0.2 s. (b) For 0.2≤t<0.35 s. (c) For 0.35≤t<0.5 s.
3. The Constant Ripple

In the constant ripple method [26]-[27] (the hysteresis current control), the switching cycle \( T_s \) is defined by replacing the peak-to-peak current ripple in (6) with the constant current ripple \( \Delta I_p \text{const} \) and solving for \( T_s \):

\[
T_s = \frac{\Delta I_p \text{const} V_{dc}}{L \omega m} \sqrt{3} \frac{1}{1 - m \sin(\omega t) \left| m \omega \sin(\omega t) \right|}
\]  

(13)

Figura 13 The constant ripple. (a) Instantaneous and average switching loss. (b) Current THD.

4. The Proposed Technique

In the proposed technique, the switching cycle \( T_s \) and the fraction \( C_3 \) are optimally adjusted to obtain the similar current THDs to those of the TDD in the three intervals, respectively.

In the first interval \( 0 < t < 0.2 \) s, the switching cycle \( T_s \) and the fraction \( C_3 \) are chosen as 200 \( \mu s \) and 0.5, respectively. In the second interval \( 0.2 < t < 0.35 \) s, \( T_s \) and \( C_3 \) are adjusted as 87 \( \mu s \) and 0.25, respectively. In the last interval \( 0.35 < t < 0.5 \) s, \( T_s \) and \( C_3 \) are also adjusted as 98 \( \mu s \) and 0.53, respectively.

Figure 14 The proposed technique. (a) Instantaneous and average switching loss. (b) Current THD.
ent lags behind the voltage when both the active power of 3 kW, the higher current results in the higher average switching loss and reactive power current in Fig. 8. Conversely, the lower current THD of 2.

The simulation results of the proposed THD show that the current THD is 4.71% in Fig. 8(a) and the lower current THD of 2.68% in Fig. 8(b). Conversely, in the second interval (0.2-0.35 s), with an active power of 1.5 kW, the lower current leads to a lower switching loss of 15.37 W and a higher current THD of 5.58% which

<table>
<thead>
<tr>
<th>Switching cycle</th>
<th>0&lt;t&lt;0.2s</th>
<th>0.2s&lt;t&lt;0.35s</th>
<th>0.35s&lt;t&lt;0.5s</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Switching Loss (W)</td>
<td>THD (%)</td>
<td>Switching loss (W)</td>
</tr>
<tr>
<td>Constant</td>
<td>30.71</td>
<td>2.68</td>
<td>15.37</td>
</tr>
<tr>
<td>TDD</td>
<td>15.35</td>
<td>4.73</td>
<td>17.53</td>
</tr>
<tr>
<td>Constant ripple</td>
<td>14.21</td>
<td>4.71</td>
<td>17.10</td>
</tr>
<tr>
<td>Proposed</td>
<td>13.65</td>
<td>4.71</td>
<td>16.14</td>
</tr>
</tbody>
</table>

5. DISCUSSIONS
The simulation results of the grid-connected inverter system are shown in Figs. 7-16 and Table II. The responses of the active and reactive powers are shown in Fig. 7 for the fixed switching frequency of 10 kHz. The responses of the grid voltage and current in Fig. 8 show that the current lags behind the voltage when both the active and reactive powers are injected into the grid. In the first interval (0-0.2 s), with an active power of 3 kW, the higher current results in the higher average switching loss of 30.71 W in Fig. 9(a) and the lower current THD of 2.68% in Fig. 9(b). Conversely, in the second interval (0.2-0.35 s), with an active power of 1.5 kW, the lower current leads to a lower switching loss of 15.37 W and a higher current THD of 5.58% which
For the TDD-based variable switching frequency, the carrier frequency is changed to reduce the switching loss in Fig. 10(b) while meeting the TDD limit (5%) specified in [4]. To reduce the average switching loss from 30.71 W to 15.35 W to improve efficiency in the first interval, the TDD technique must decrease the carrier frequency from 10 kHz to 5 kHz, and the THD of 4.73% in Fig. 10(b) continues to meet the requirement. On the other hand, the switching frequency is increased from 10 kHz to 11.4 kHz in the second interval to decrease the THD from 5.58% to 4.78%, therein causing the switching loss to increase from 15.37 W to 17.53 W. Similarly, to reduce the switching loss from 18.45 W to 18.34 W in the third interval, the switching frequency is also decreased from 10 kHz to 9.95 kHz; however, this also increases the THD from 4.57% to 4.6%. Moreover, the current spectrum in Fig. 11 shows that some individual harmonics are still significantly high (up to 2.5%), although the THD remains below the limit. Therefore, these harmonics can also cause some noise in communications. In this paper, the switching loss of the TDD is chosen as the benchmark to compare to the constant ripple technique and the proposed technique.

To obtain the current THDs in Fig. 13(b) similar to those of the TDD technique, the switching loss of the constant ripple in Fig. 13(a) is equal to 14.21 W, 17.1 W, and 16.9 W in the three intervals, respectively. The spectrum of the constant ripple in Fig. 12 covers in a wide range and this leads to reducing individual harmonic amplitude significantly.

The results of the proposed technique are shown in Figs. 14-15. With the similar current THDs compared to those of the TDD, the average switching loss in Fig. 14(a) is 13.65 W, 16.14 W, and 16.8 W for the three intervals, respectively. These mean that the switching loss of the proposed technique in Table II is lower than that of the TDD technique for the first, second, and third intervals, respectively. For the case of unity power factor (resistance load or active power only), the instantaneous and average switching losses in half of the fundamental period are shown in Fig. 16(b). Similarly, for the case of non-unity power factor (R-L load or active and reactive power), the losses are also shown in Fig. 16(c).

For the third interval, the slightly higher THD of the proposed 4.71% in Fig.15(c) results in the slightly lower switching loss of the proposed compared to that of the constant ripple in Table II.

The switching cycle near zero current, \( \omega t=0.15\pi \), and \( \omega t=0.85\pi \) of the proposed method in Fig. 16(a) is much lower than that of the constant ripple cycle. This causes the current ripples to be greatly reduced near zero current, whereas the instantaneous switching loss in Figs. 16(b) and 16(c) see minimal increases. Similarly, the switching cycle near the peak current (\( \omega t=0.5\pi \)), is also higher than that of the constant ripple method. This also leads to the significant reduction in the instantaneous switching loss, whilst the current ripples increase insignificantly. The low switching frequency near the high instantaneous current of the proposed also contributes to increasing the lifetime of the semiconductor devices.

6. CONCLUSIONS
Reducing the current THD of grid-connected inverters is a pressing requirement for meeting stringent grid codes. Selecting the optimal switching cycle by striking a
balance between the switching loss and current THD is a difficult and complicated problem.

The authors improved the spread of acoustic noise spectrum technique in [19] for space vector PWM to SPWM to enable a suitable comparison with under the same conditions. It is observed that by using the SPWM technique with a variable switching cycle in every half of the fundamental period, the average switching loss can be significantly decreased compared to what can be achieved with a constant switching cycle. The performance of the proposed method is outstanding compared to the TDD and constant ripple techniques, although the current THDs remain equal in all cases.

In addition, the low switching frequency of the semiconductor switches near the peak current can help decrease the thermal stress, thus increasing the inverter’s lifetime. The current ripple, which is significantly reduced near zero output current, also helps improve the power quality. This enables the use of grid-connected devices that use phase-lock loops, such as digital electronic meters and synchronization detectors, for the recognition of phase angle and frequency to ensure proper operation.

To address the changing load currents under practical conditions, the cycle $T_{sc}$ and fraction $C_3$ can be prepared offline for various current levels using the lookup table in MATLAB.

The simulation results were implemented to verify the theory. For a grid-connected inverter, operation at a non-unity power factor was also verified by injecting the reactive power into the grid.

With the proposed technique, the current THD can be significantly reduced for the same given switching loss. In addition, this approach can be applied to three-phase inverters and other PWM techniques.

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REFERENCES
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