A Three-Phase Grid-Connected Micro-Inverter for AC Photovoltaic Module Applications

Jianghua Feng, Hui Wang, Junfeng Xu, Mei Su, Weihua Gui, Xing Li

Abstract—Photovoltaic (PV) micro-inverter converts the DC from a PV panel to AC directly, which has the advantages of improved energy harvesting, friendly “plug-and-play” operation, enhanced flexibility/expandability, excellent system redundancy and no DC cabling/safety issue, therefore it is an attractive solution for grid-connected PV system. In AC PV module applications, the features like power density, reliability, efficiency and reactive power capability are essential for the micro-inverters. In order to overcome the drawbacks of the conventional micro-inverters including the power density/reliability issues caused by the bulky input capacitors and the limited output reactive power capability, a three-phase micro-inverter without energy storage capacitors is proposed in this paper. The proposed micro-inverter consists of a flyback stage, a third-harmonic injection circuit and a line-commutated current-source type inverter. The flyback stage realizes maximum power point tracking, while the third-harmonic injection circuit is responsible for output power factor correction. In addition to having the potential of achieving compact structure, long lifetime and high efficiency, the developed micro-inverter provides extended output reactive power control range and three-phase balanced output. Finally, the proposed topology and method are verified by simulations and experimental results.

Index Terms—Flyback, maximum power point tracking, photovoltaic micro-inverters, third-harmonic injection.

I. INTRODUCTION

In recent years, the issues of electrical energy shortage and environmental pollution have resulted in the increasing demand of renewable energy sources, such as wind power, hydrogen energy, solar energy and so on. Among these renewable energy sources, the photovoltaic (PV) sources have the advantages of no supply limitations, emission-free and high reliability, and thus the number of installations over the world has experienced a steep rise [1], [2]. In 2016, the total installed capacity of PV has reached to about 300 GW, with an annual increased capacity of 70 GW and an annual growth rate of 30%. Meanwhile, the average price per watt of the PV module in 2016 was $0.36/Wp, and the accumulative reduction in the price is roughly 80% during 2010-2016 [3]-[5]. Moreover, the PV sources are expected to be the biggest contributors to power generation among all renewable energy sources by 2040 [2], [6]-[12].

Generally, the grid-interactive PV systems can be divided into three categories: the centralized inverter system, the string inverter system and the AC module system [13]-[18]. Among these, the centralized and string inverter systems are superior in terms of conversion efficiency, unit cost, power density and system monitoring/maintenance, and thus become the main form in the practical applications with a great commercial success. In respect to the residential and commercial applications, the centralized and string inverter systems are used, where the PV panels are usually connected in series so as to obtain a sufficiently high DC voltage required by the central inverters. In these cases, the output currents of all PV panels are equal to the string current and hence the global maximum power point tracking (MPPT) is hard to realize when the parameters mismatch of the PV modular happens. Consequently the efficiency of energy harvesting significantly reduces. Besides, the centralized and string inverter systems also have some other disadvantages such as the safety problem due to DC wires, no expandability, and so on [16], [18].

Different from the centralized and string inverter systems, the AC module system is a kind of PV system with small capacity, which integrates the inverter into the PV panel and directly converts the DC voltage to high AC voltage compatible with the grid. By equipping each PV panel with a micro-inverter, the PV panels are operated independently in their respective maximum power point (MPP) and hence the issue of the power generation reduction caused by module mismatch is eliminated. Compared with the centralized and string inverter systems, the PV micro-inverter has the advantages of improved energy harvesting, friendly “plug-and-play” operation, enhanced flexibility and expandability, excellent system redundancy and no DC cabling/safety issue, although the challenges of reducing unit cost and enhancing conversion efficiency in the PV micro-inverter system remain. Because of these advantages, the PV micro-inverters have attracted an increasing attention and became competitive alternatives to the centralized and string inverters for PV system, and many PV micro-inverter topologies and control schemes have been proposed and discussed [16]-[33]. However, several problems of the conventional micro-inverters still exist.

Firstly, the power density and reliability issues caused by the energy storage elements are the main concerns for most of the existing PV micro-inverters. For the conventional single-phase micro-inverters, the imbalance of the instantaneous power between the PV side and the grid side causes power pulsation at twice the line-frequency, which would affect the MPPT of the
PV panel, therefore the bulky electrolytic capacitors are usually equipped at the PV input side instead of the active power decoupling techniques [18], [20], [30]-[34], which served as the energy buffering elements in many commercially micro-inverters. Although it has the advantage of the low cost, the electrolytic capacitors inevitably introduce challenges on the power density and the reliability, which is sensitive to panel-integrated micro-inverters [21]-[26]. Secondly, most of the micro-inverters consist of a quasi-single-stage architecture and utilize an unfolding bridge to realize DC/AC conversion [17], [21]-[26]. As a result, the micro-inverters could only be operated in unity output power factor condition and cannot provide reactive power compensation, which makes the PV micro-inverters fail to meet the requirements of some electric power regulations [35].

In this paper, to solve the power density/reliability issues caused by the bulky energy storage elements and improve the output reactive power control range, a three-phase micro-inverter topology without bulky energy storage capacitors is presented and investigated systematically. The proposed micro-inverter mainly consists of a DC/DC stage including four flyback converters, a third-harmonic injection circuit derived from [36]-[40] and a line-commutated current-source type inverter (CSI). The flyback converters are used to realize independent MPPT of each PV panel, and the third-harmonic injection circuit is responsible for output power factor correction (PFC) while the CSI is commutated at line-frequency so as to accomplish the DC/AC inversion. In addition to possessing the advantageous features of improved energy harvesting, “plug-and-play” operation, enhanced modularity/flexibility and excellent system redundancy, the developed micro-inverter has the potential of achieving high efficiency/reliability, extended output reactive power control range and three-phase balanced output. This paper is organized as follows: Section II introduces the topology and operating principles of the micro-inverter, followed by the mathematical verification of sinusoidal output currents; Section III presents the control algorithm for tracking MPP and the control scheme of the third-harmonic injection circuit; Section IV shows the simulation and experimental results to verify the presented methods; Section V draws the final conclusion of this paper.

II. CIRCUIT CONFIGURATION AND OPERATING PRINCIPLES

A. Circuit Configuration

The proposed 1000W three-phase micro-inverter directly connects four adjacent 250W PV panels, and the system configuration is shown in Fig. 1, which consists of a DC/DC conversion stage, an active third-harmonic current injection circuit, a three-phase line-commutated CSI and an output LC filter. The DC/DC conversion stage is composed of four flyback converters, and each flyback converter consists of two interleaved sub-converters.

It should be noted that different from the single PV input situation in [21]-[26], the developed micro-inverter has four PV input terminals and is installed directly on the back of one PV panel, and each PV panel is connected to one of the PV input terminals of the micro-inverter. Thus, each PV panel is operated independently with its own MPP. Usually, in the conventional micro-inverters, a bulky PV input capacitor is set to decouple the power pulsation at twice the line frequency. Differently, the input capacitor $C_f$ of the suggested micro-inverter is used only for providing high frequency input pulse current of the flyback converter. Therefore, ceramic capacitors with small capacitance, long lifespan and high reliability are adopted instead of the electrolytic capacitors.

The line-commutated CSI consists of three arms and three bidirectional switches. Compared with the conventional VSIs, the selected CSI has the following advantages: 1) the switches of the CSI are commutated with line frequency here, the switching losses of power switches are ignored, therefore the power losses are relatively low and high conversion efficiency is achieved; 2) the CSI realizes DC/AC inversion without any energy storage capacitors, consequently high power density and reliability of the converter are attainable.

The output filter consisting of inductor $L_c$ and film capacitor $C_f$ is mainly used for smoothing the pulse currents generated by the inverter so as to produce three-phase sinusoidal output currents. The active third-harmonic current injection circuit is composed of a third-harmonic injection inductor and a bridge leg, which constitutes the topology used for AC PV module applications. By proper control of the third-harmonic injection circuit, extended output reactive power control range is achieved.

The DC/DC stage and the CSI work together to accomplish the energy conversion process and cannot work independently. Therefore, the presented topology is similar to the conventional single-phase micro-inverters in terms of construction approach of the topology, operating principles and functionality, and may be regarded as the generalization of the conventional micro-inverter from single-phase to three-phase.

![Fig. 1. Schematic diagram of the proposed micro-inverter topology.](image)

B. Operating Principles

Assuming that the three-phase grid voltages are symmetrical and sinusoidal, the operating principles of the suggested three-phase micro-inverter are described as follows. For the flyback stage, each flyback converter tracks the respective MPP of the corresponding PV module, and the two sub-converters of each flyback converter are placed in parallel and interleaved with 180° phase-shift. By interleaving the two sub-converters, the input/output current ripples are reduced and the power range of each flyback converter is extended. The third-harmonic injection circuit is essentially a time-sharing multiplexing single-phase shunt active power filter, the switches $S_1$ and $S_2$, working in high frequency are switched to generate the desired quasi-third-harmonic current $i_r$, which injects into the output phase with the minimum absolute voltage, and then the PFC of the output side is achieved.
The line-commutated CSI can be regarded as a grid voltage selector, which connects output phase \( a, b, c \) to either node \( p, y \) or \( n \) periodically according to the instantaneous values of the output phase voltages. Take the case of \( u_a \geq u_b \geq u_c \), for example, only the switch \( S_{a+} \) in the upper arms, the switch \( S_{c-} \) in the lower arms and the bidirectional switch \( S_m \) are in ON state, in order to impose the piecewise six-pulse shape voltage \( u_{in} \) across the intermediate DC-link. In conclusion, the switches of the CSI are worked according to Table I, where \( \theta_n \) is the phase angle of the output phase voltage \( u_{in} \).

As described above, the switches of the CSI are commutated at line frequency and the DC link voltage exhibits a six-pulse shape waveform, which implies that the flyback stage, the third-harmonic injection circuit and the CSI stage can be controlled independently. Therefore, this section mainly focuses on the analysis of the operating principle of the flyback stage. Assuming that the three-phase output voltages satisfy \( u_a \geq u_b \geq u_c \) (denoted as sector 1) and only one sub-converter operating in continuous conduction mode (CCM) is included in the topology, the detailed analysis of the micro-inverter in a switching period is explained as follows.

![Fig. 2a](image1)

### Fig. 2a. Equivalent circuit of the proposed AC module micro-inverter (assuming \( u_a \geq u_b \geq u_c \)). (a) Operating stage I. (b) Operating stage II. (c) Partial key waveforms of the inverter in a switching period.

#### Table I

<table>
<thead>
<tr>
<th>Sector</th>
<th>( S_a )</th>
<th>( S_b )</th>
<th>( S_c )</th>
<th>( S_{a+} )</th>
<th>( S_{b+} )</th>
<th>( S_{c+} )</th>
<th>( S_{a-} )</th>
<th>( S_{b-} )</th>
<th>( S_{c-} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td>2</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>0</td>
<td>1</td>
<td>0</td>
</tr>
</tbody>
</table>

#### Table II

<table>
<thead>
<tr>
<th>( \theta_n )</th>
<th>( \phi )</th>
<th>( \phi_n )</th>
<th>( \phi_{inj} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>0-( \pi/3 )</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>( \pi/3-2\pi/3 )</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
</tbody>
</table>

Stage I (\( S_{n1} = 1 \)): Fig. 2a shows the equivalent circuit of the first stage. In this stage, the switch \( S_{n1} \) of flyback converter I is ON. At the same time, the rectifier diode \( D_{n1} \) is in reverse blocking state and the secondary current \( i_{n1} \) is zero. Then, the input current \( i_{in1} \), as well as the magnetizing current \( i_{n1} \), linearly increase and the magnetizing current is expressed as

\[
L_{in1} \frac{di_{n1}}{dt} = u_{pp1}
\]

where \( L_{in1} \) is the magnetizing inductance of the transformer, and \( u_{pp1} \) is the PV voltage.

Stage II (\( S_{n1} = 0 \)): The equivalent circuit of this stage is shown in Fig. 2b. In this stage, the switch \( S_{n1} \) is turned off and the input current \( i_{n1} \) is zero. The magnetizing current \( i_{n1} \) linearly decrease and the magnetizing current is expressed as

\[
L_{in1} \frac{di_{n1}}{dt} = -u_{in} / N = -(u_a - u_n) / N
\]

where \( N \) is the turns ratio of the flyback transformer. Besides, the rectifier diode \( D_{n1} \) is in conduction state and the energy stored in the primary winding is released to the output side. Omitting the effects of the snubber capacitor \( C_s \), the secondary current \( i_{n1} \) and the output current \( i_o \) could be obtained by

\[
i_o = i_{n1} = i_{n1} / N
\]

Fig. 2c shows the partial key waveforms of the micro-inverter in a switching period \( T_o \), where \( f_o \) is the switching frequency, and \( k \) and \( d \) are the duty ratios of the switches \( S_{a+} \) and \( S_{c+} \).

![Fig. 3a](image2)

### Fig. 3a. Key waveforms of the proposed micro-inverter. (a) Output voltages and output currents, DC link voltage and third-harmonic injection current. (b) Voltages and currents of the flyback converters. (c) Gating signals.
In this manner, sinusoidal output currents, controllable output power factor and global MPPT are achieved in the developed micro-inverter. The key waveforms of the micro-inverter are shown in Fig. 3.

C. Analysis of the Input and Output Currents

For grid-connected PV inverters, sinusoidal grid-connected currents, constant output power and MPPT of each PV panel should be guaranteed simultaneously in steady state. For the developed micro-inverter, the realizability of sinusoidal grid-connected currents and MPPT are carefully verified in the following discussion. Since each sub-converter is operated independently and the three-phase output voltages is symmetrical, one specific case, where the output voltage vector is located in sector 1 and only one sub-converter is included in the flyback stage, is used for analyzing the input and output currents in details.

Suppose that the three-phase output voltages are given by

\[ u_{na} = U_{om} \cos(\theta_{na}) \]
\[ u_{nb} = U_{om} \cos(\theta_{na} - 2\pi / 3) \]  
\[ u_{nc} = U_{om} \cos(\theta_{na} + 2\pi / 3) \]  

where \( U_{om} \) is the magnitude of the output phase voltage.

Sort the output voltages according to the relationships of the instantaneous values as follows:

\[ u_{max} = \max(u_{na}, u_{nb}, u_{nc}) \]
\[ u_{mid} = \text{mid}(u_{na}, u_{nb}, u_{nc}) \]
\[ u_{min} = \min(u_{na}, u_{nb}, u_{nc}) \]  

where \( \max(), \text{mid}() \) and \( \min() \) are the operator of the maximum, medium and minimum value, respectively. As can be known from the operating principles, the DC link voltage in sector 1 is expressed as

\[ u_{pm} = u_{max} - u_{min} = u_{na} - u_{nc} \]  

Referring to the equivalent circuits in Fig. 3, the duty ratio \( d \) is obtained as (7) according to the voltage-second balance principle of the magnetizing inductor \( L_{m1} \):

\[ d = \frac{u_{pm}}{(N u_{py1} + u_{pm})} \]  

Assume that the averaged value of the input current \( i_{pm1} \), denoted as \( \bar{i}_{pm1} \), tracks the constant reference current \( \bar{i}_{pm1} \) exactly in steady state. Then, according to the ampere-second balance principle of \( C_{ij} \), the averaged value of the output current \( i_{py1} \) and the averaged output power of the PV panel are deduced as

\[ \bar{I}_{py1} = \bar{I}_{pm1} = \bar{i}_{pm1} \]
\[ \bar{P}_{py1} = u_{py1} \bar{I}_{py1} = u_{py1} \bar{i}_{pm1} \]

The averaged value of the secondary current \( i_{pm1} \) and the averaged output power of flyback converter1 are expressed as

\[ \bar{I}_{pm1} = \frac{\bar{P}_{py1}}{u_{pm}} \]
\[ \bar{P}_{e} = u_{pm} \bar{I}_{pm1} = \bar{P}_{py1} \]

(9)

It can be known from (8)-(9) that the averaged output power of the PV panel and the averaged output power of flyback converter1, denoted as \( \bar{P}_{py1} \) and \( \bar{P}_{e} \) are both constant in steady state.

In each switching period, according to the voltage-second balance principle of \( L_{p} \), the duty ratio \( k \) is obtained as

\[ k = \frac{u_{mid} - u_{min}}{(u_{max} - u_{min})} \]  

(10)

The third-harmonic injection inductor current \( i_{j} \) is given by

\[ i_{j} = \frac{I_{m1} \cos(\theta_{j} - 2\pi / 3) + I_{m1} \sin(\theta_{j} - 2\pi / 3)}{[G \cos(\theta_{na} - 2\pi / 3 + \phi)] / \cos \phi} \]

\[ \tan \phi = \frac{I_{om}}{I_{m1}} \]

where \( I_{m1} \) and \( I_{om} \) are the amplitudes of the active and reactive components of the grid-connected current, respectively; \( G = 2P_{py1}/3u_{om}^{2} \) is the equivalent output conductance and \( \phi \) is the output displacement angle.

Neglecting the effects of the output filtering capacitor \( C_{e} \), it is found from Table I that the output phase current \( i_{j} \) in sector 1 is equal to the third-harmonic injection inductor current \( i_{y} \)

\[ i_{y} = \bar{i}_{pm1} - ki_{y} \]  

(12)

Using the equivalent circuits shown in Fig. 2, the output current of phase \( a \) can be expressed as

\[ i_{a} = \bar{i}_{pm} - ki_{y} \]  

(13)

In sector 1, the DC link voltage \( u_{om} \) is equal to the input line-line voltage \( u_{ac} \). According to (7), the duty ratio \( k \) is equal to \( u_{ac}/u_{om} \). Substitute (9) and (11) into (13), \( i_{a} \) is deduced as

\[ i_{a} = \frac{G u_{om} \cos(\theta_{na} + \phi)}{\cos \phi} \]  

(14)

As the three-phase output currents satisfy \( i_{x} + i_{y} + i_{z} = 0 \), the output current of phase \( c \) is obtained as

\[ i_{c} = \frac{G u_{om} \cos(\theta_{na} + 2\pi / 3 + \phi)}{\cos \phi} \]  

(15)

According to equations (8), (9), (11), (12), (14) and (15), the output power of the PV panel is constant and each PV panel achieves MPPT by adjusting its respective reference current \( \bar{i}_{pm1} \). Besides, the three-phase output currents are symmetrical and sinusoidal with a desired displacement angle. It is easy to find that the range of the output displacement angle of the suggested micro-inverter is from \(-\pi / 2\) to \(\pi / 2\). For the remaining sectors and multiple flyback converters configuration, similar results can be obtained and the derivation process is not repeated here.

III. CONTROL OF THE PROPOSED MICRO-INVERTER

As discussed in the previous section, the CSI of the micro-inverter commutates at line frequency and hence can be treated as a passive part. Therefore, only the flyback stage and the third-harmonic injection circuit need to be properly controlled so as to realize MPPT and PFC at the output side. In this paper, the flyback stage is utilized to achieve MPPT, and each flyback converter is controlled to track its MPP. Furthermore, by adopting shunt active power filtering and time-sharing multiplexing injection techniques, the third-harmonic injection circuit achieves sinusoidal three-phase grid-connected current and the expected output side power factor. The concrete analysis is given as follows.

A. MPPT control of the PV system

In general, at the certain irradiance level and ambient temperature, the PV panel has one operating point that results in a maximum output power. In order to maximally extract the output power of the PV panel, the panel-integrated micro-inverter needs to continuously track this operating point even when it changes due to the moving sun, as well as the partial shading. For a PV panel, it can be known from the
current-voltage characteristics that the PV current is determined by the PV voltage. The current and voltage together determine the output power of the PV panel, and either the PV voltage or PV current can represent the MPP of a PV panel. Therefore, the MPP tracking can be realized by regulating either the PV voltage or PV current to a value that represents the MPP.

In this work, a perturbation and observation (P&O) MPPT algorithm based on PV current regulation is adopted for the three-phase micro-inverter due to its advantages of good performance and low implementation efforts [12].

As can be seen from Fig. 1, the PV current and PV voltage can be changed by regulating the averaged input current of the corresponding flyback converter. Therefore, the MPPT of each PV panel is achieved by controlling the averaged input current of the flyback converter to track the reference PV current generated by its MPP tracker. In this manner, global MPPT of the micro-inverter is achieved.

Fig. 4 shows the control block of the MPPT controller of the micro-inverter, where \( G_D (s) \) and \( G_V (s) \) are the transfer functions of the sub-converter1 and sub-converter2 of the flyback converter1, respectively. The MPPT controller is a cascaded structure, where the outer control loop is the MPPT tracker, and the inner current control loop adopts a conventional proportional-integral (PI) controller. In the output-loop, the MPPT tracker measures the PV voltage and PV current, periodically adjusts the current reference and the direction of the perturbation based on the previous operation and the changing of the PV panel power. For the inner-loop, the PI controller controls the practical PV current to track the current reference, and the output of the inner control loop is the duty ratio of the switch in flyback converter1. In order to improve the dynamic tracking speed, the steady state duty ratio \( d^* \) is fed forward to the output of current controller, which is obtained by the DC link voltage and the PV voltage.

As can be seen from (8), the averaged value of the primary input current in a switching period is equal to the PV current. Therefore, instead of the PV current, the averaged primary input currents are used to implement the MPPT controller. Although the complexity of the current signals processing and the implementation efforts increase, the latter method obviously has the advantages of low cost of current sampling, good dynamic performance, as well as fast response to over-current fault. In order to guarantee the sinusoidal grid-connected current, the frequency of the outer-loop controller should be relatively low. Moreover, the bandwidth of the inner loop is comparatively high so as to decouple the outer and inner loops. The procedure of the controller is designed by conventional frequency-domain analysis tools, and it is not given here for brevity.

Besides, it should be noted that the mismatched parameters between the two sub-converters, such as the magnetizing inductances of the transformers, lead to undesirable current/power imbalance between the two sub-converters. Therefore, extra power balancing control loop is required. In this work, a simple PI controller is adopted to balance the power among the two sub-converters. The difference between the averaged input current of sub-converter1 and sub-converter2 is feed into the power balancing PI regulator, and the output of the PI regulator is added to the original duty ratio of sub-converter1 and subtracted from the original duty ratio of sub-converter2, respectively. As a result, current/power balance between the two sub-converters is achieved.

B. Control of the Third-Harmonic Injection Circuit

Similar to the third-harmonic injection matrix converter [37], proper control and design of the third-harmonic injection circuit are the key challenges to implement the suggested micro-inverter, because PFC at the output side of the micro-inverter is realized by proper design and control of the third-harmonic injection circuit. The detailed analysis of the controller and design procedure of the third-harmonic injection circuit has been discussed in [37], and thus is not elaborated here.

Fig. 5 depicts the control block of the whole micro-inverter system, where \( \theta_{m} \) is the phase angle of the output phase voltage with the medium value, \( P_o \) is the total output power reference and \( G_p (s) \) is the system transfer functions of the third-harmonic injection circuit. First, the output power references, generated from the MPP trackers, are added to obtain the total output power reference. Then the third-harmonic injection inductor current reference \( i^*_l \) is calculated by the output phase voltages, the total output power reference and the expected output displacement angle. A PI controller with feed-forward term is used to control the practical third-harmonic injection inductor current.

IV. SIMULATION AND EXPERIMENTAL RESULTS

In this section, the functionality and performance of the three-phase micro-inverter were firstly evaluated by simulation using Matlab/Simulink software and then were validated experimentally. Simulation study was conducted by using the same parameters as the final laboratory prototype, and the parameters of the micro-inverter are listed in Table II.

### TABLE II: SYSTEM SPECIFICATIONS OF THE MICRO-INVERTER

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power rating</td>
<td>1000 W (4 × 250 W)</td>
</tr>
<tr>
<td>Output line-line voltage</td>
<td>220 Vrms</td>
</tr>
<tr>
<td>Output frequency ( f_o )</td>
<td>50 Hz</td>
</tr>
<tr>
<td>Switching frequency ( f_s )</td>
<td>56 kHz</td>
</tr>
<tr>
<td>Magnetizing inductance</td>
<td>54 µH</td>
</tr>
<tr>
<td>Turns ratio ( N )</td>
<td>6.5</td>
</tr>
<tr>
<td>Inductor ( L_c )</td>
<td>300 µH</td>
</tr>
<tr>
<td>Capacitor ( C_F )</td>
<td>3.3 µF</td>
</tr>
<tr>
<td>Inductor ( L_{dc} )</td>
<td>1.2 mH</td>
</tr>
</tbody>
</table>
A. Simulation Results

In the following simulation results, in order to accelerate the simulation process, the micro-inverter is studied without the MPPT control, and the PV panels of the micro-inverter are replaced by constant DC voltage sources. Therefore, the output current reference of each PV panel is directly from the command instead of the MPPT tracker.

Fig. 6 depicts the waveforms of the micro-inverter working with half of the rated power. In this scenario, the output current references of the PV panels were set such that the corresponding PV output power were $P_{PV1} = 250$ W, $P_{PV2} = 125$ W, $P_{PV3} = 125$ W and $P_{PV4} = 0$ W, respectively. Besides, the expected output displacement angle $\varphi$ was set as 0. The waveforms shown in Fig. 6 consist of the output phase voltage $u_{ao}$, the output current $i_o$, the output current $i_b$ and the output current $i_c$. As can be seen from Fig. 6, the three-phase output currents are nearly sinusoidal and in phase with the output phase voltages, except for a small phase lag caused by the capacitive currents drawn by the output filtering capacitors. Therefore, it is clear that three-phase sinusoidal output currents and unity output power factor are achieved for the developed micro-inverter. Moreover, the output power of the micro-inverter is nearly 500 W, which indicates that the each flyback converter has been controlled to effectively track its individual MPP independently.

Fig. 7 shows the waveforms of the micro-inverter working in full-load condition. Therefore, it is clear that three-phase sinusoidal output currents and unity output power factor are achieved for the developed micro-inverter. Moreover, the output power of the micro-inverter is nearly 500 W, which indicates that the each flyback converter has been controlled to effectively track its individual MPP independently.

Fig. 8 illustrates the reactive power capability of the micro-inverter. The micro-inverter runs under full-load condition (1000 W) and the expected output reactive powers are 268 Var (capacitive) in Fig. 8 (a) and 577 Var (capacitive) Fig. 8 (b), and the equivalent output displacement angles are 15° and 30°, respectively.

B. Experiments

In order to validate the theoretical analysis and simulation results, a laboratory prototype of the micro-inverter with the specifications given in Table II is built, as shown in Fig. 9. The control platform of the micro-inverter is based on a floating-point Digital Signal Processor (DSP) TMS320F28335. For the flyback stage, considering the 18-55V PV panel voltage, the 220 Vrms grid voltage and a turns ratio of 6.5, the Metal Oxide Semiconductor Field-Effect Transistor (MOSFET) IRFS4321 is chosen as the switches of the flyback converters. Besides, Silicon Carbide Schottky Diode C2D05120D is
selected as the output rectification diodes due to its characteristic of zero reverse-recovery losses. For the flyback transformer, the RM14 ferrite core is adopted, and the magnetizing inductance is designed such that CCM mode is guaranteed within the whole operating range. For the line-commutated switches and the bidirectional switches of the CSI, MOSFET SPW47N60C3 and STD15N50 are chosen due to their low saturation voltages. For the switches of the third-harmonic injection circuit working at high frequency, high speed MOSFET IPB60R190C6 is used.

Fig. 9. The 1000 W three-phase micro-inverter prototype.

Fig. 10. Experimental waveforms of the micro-inverter working with half of the rated power. CH1 is the input phase voltage $u_{sa}$, CH2 is the output current $i_a$, CH3 is the output current $i_b$, and CH4 is the output current $i_c$.

Fig. 11. Experimental waveforms of the micro-inverter working in full-load condition. CH1 is the input phase voltage $u_{sa}$, CH2 is the output current $i_a$, CH3 is the output current $i_b$, CH4 is the output current $i_c$.

Figs. 10-11 demonstrate the experimental results of the micro-inverter with different output reactive power. (a) The expected output reactive powers is 268 Var. (b) The expected output reactive powers is 577 Var. CH1 is the input phase voltage $u_{sa}$, CH2 is the output current $i_a$, CH3 is the output current $i_b$, CH4 is the output current $i_c$.

Fig. 12. Experimental waveforms of the micro-inverter with different output displacement angles. (a) The expected output reactive powers is 268 Var. The expected output reactive powers is 577 Var. CH1 is the input phase voltage $u_{sa}$, CH2 is the output current $i_a$, CH3 is the output current $i_b$, CH4 is the output current $i_c$.

Figs. 10-11 correspond to the simulated results shown in Figs. 6-8, and the experimental conditions and commands are exactly the same as these in the simulation. As can be seen from Figs. 10-12, the experimental results match the simulated results well, except for slightly higher distortions of the output currents. The higher distortions of the output currents in the experiments is mainly attributed to the non-idealities of the grid voltage, the non-idealities of the power devices, the dead time effects, etc.

Fig. 13(a)-(d) present the total harmonic distortions (THDs) of the output currents corresponding to the cases in Figs. 10-12, where the number of harmonics included in the Fast Fourier Transform (FFT) analysis is 50 (up to 2.5 kHz). It should be noted that the output wires were crossed the current probes twice, in order to improve the signal-to-noise ratio of the measurement. Therefore, the measured amplitudes of the currents were also doubled. As can be seen from the FFT analyses, the THDs of the output current corresponding to the operating conditions of half-load, full-load, output displacement angle of 15° and output displacement angle of 30°, are 6.82%, 4.29%, 3.53% and 4.2%, respectively, which are acceptable in practice applications.
Fig. 13. THDs of the output current with: (a) Half of the rated power. (b) Full-load. (c) Displacement angle of 15°. (d) Displacement angle of 30°.

Fig. 14. Experimental MPPT performance of the micro-inverter. CH1 is the PV voltage \( u_{PV1} \), CH2 is the PV current \( i_{PV1} \), CH3 is the output voltage \( u_{sa} \), CH4 is the output current \( i_{a} \), M is the PV output power.

Fig. 14 presents the experimental MPPT performance of the micro-inverter. In this experiment, the flyback converter1 corresponding to PV panel1 is connected to a solar array simulator (E4361A, Agilent) whose MPP is set as 250W, and the other flyback converters are disabled to simulate the partial shading condition of the whole PV system. As can be seen from the result, the practical output power of the solar array simulator tracks the set-point quickly and accurately. Therefore, the results in Fig. 14 demonstrate the correctness of the developed MPPT algorithm. Referring to Figs. 10-14, sinusoidal output currents, controllable output power factor and independent MPPT are achieved in the prototype, thus the validity of the micro-inverter is verified experimentally.

To evaluate the efficiency feature of the presented topology, the power losses and efficiency of the overall system were firstly calculated and then were measured practically. The system power losses can be divided mainly into two categories: semiconductor losses and passive components losses. For the semiconductor devices, the blocking losses can be neglected. Thus only the conduction and switching losses are considered in this paper.

The conduction losses of the switches and diodes can be calculated by

\[
P_{C,S} = I_{S,AVG} U_{CE0} + I_{S,RMS}^2 R_s
\]

and

\[
P_{C,D} = I_{D,AVG} U_{DD0} + I_{D,RMS}^2 R_D
\]

where \( P_{C,S} \) and \( P_{C,D} \) are the conduction losses of the switch and diode, \( R_s \) and \( R_D \) are the dynamic on-resistances of the switch and diode, \( U_{CE0} \) and \( U_{DD0} \) are the forward voltages of the switch and diode when the current is zero, \( I_{S,AVG} \) and \( I_{D,AVG} \) are the average currents of the switch and diode, \( I_{S,RMS} \) and \( I_{D,RMS} \) are the RMS currents of the switch and diode.

The switching losses of the switches and diodes are estimated as

\[
P_{S,S} = \frac{E_{on,s} + E_{off,s}}{V_{BRIS}} V_{S,AVG} I_{S,AVG} f_s
\]

and

\[
P_{S,D} = \frac{E_{on,R}}{V_{BRID}} V_{D,AVG} I_{D,AVG} f_s
\]

where \( P_{S,S} \) and \( P_{S,D} \) are the switching losses of the switch and diode, \( V_{S,AVG} \) and \( V_{D,AVG} \) are the average voltages of the
switch and diode, $E_{on,R}$ and $E_{off,R}$ are the switching energies of the switch at the reference voltage $V_{SR}$ and reference current $I_{SR}$, and $E_{off,R}$ is the reverse recovery energy of the diode at the reference voltage $V_{DR}$ and reference current $I_{DR}$.

The losses of the capacitors are obtained as

$$P_C = I_{C_{\text{RMS}}}^2 \frac{\tan \delta}{2\pi C_F f_s}$$

(20)

where $I_{C_{\text{RMS}}}$ and $\tan \delta$ are the RMS current and loss factor of the input capacitors.

The losses of the inductors can be divided into two parts: core loss and winding resistance loss. Since the inductors in the presented topology operate at low frequency, only low-frequency loss in winding resistance is considered in this paper. The power loss of the inductor is calculated as

$$P_L = I_{L_{\text{RMS}}}^2 R_L$$

(21)

where $I_{L_{\text{RMS}}}$ is the RMS current of the inductor and $R_L$ is the DC resistance of the inductor.

According to the above formulas, the overall system power loss distribution is derived as shown in Fig. 15. The nominal system efficiency calculated reaches 95.63%.

Fig. 16 shows the measured efficiency of the micro-inverter with the HIOKI 3390 digital power analyzer, where the voltages of the PV panels are 30V and the desired output displacement angle is 0. It should be noted that the auxiliary circuits such as the control circuits, are all included in the test. As can be seen from Fig. 16, the maximum efficiency of 95.52% and California Efficiency of 93.7% are achieved for the proposed micro-inverter.

It is important that the PV inverter should meet the specific electro-magnetic compatibility (EMC) standards. The electro-magnetic interference (EMI) of a PV inverter is determined by many factors, such the switching frequency. Regarding to the influence of the switching frequency, it mainly reflects in two aspects. Firstly, the cut-off frequency of the differential-mode (DM) filter is usually designed to satisfy the requirement of the output current quality and increases with the switching frequency. Therefore, different switching frequency will result in different cut-off frequency and noise attenuation of the DM filter. Secondly, for the semiconductor power converters switching at high frequency, the EMI noises are mainly generated by the high $du/dt$ and $di/dt$ rates of the power switches. Usually the $du/dt$ and $di/dt$ rates of the switches, as well as the conduction emission, increase with increasing switching frequency. In Fig. 17, the conducted emission of the prototype is measured by an EMI test receiver and a line impedance stabilization network (LISN), and the result has been compared with the CISPR 11 standard. As can be seen from Fig. 17, the conducted emission levels of the micro-inverter fully meet the standards.

V. CONCLUSION

This paper presents a grid-connected three-phase micro-inverter suitable for AC PV module applications. The proposed micro-inverter uses flyback converters as the DC/DC conversion stage, and MPPT of each PV panel is achieved by proper control of the corresponding flyback converter. A third-harmonic injection circuit is controlled to synthesize the appropriate injection current so as to realize PFC at the output side. A line-commutated CSI is used to convert the DC current to three-phase AC current, which delivers the power generated from the PV panels to the grid. Simulation and experimental results clearly show that the micro-inverter could achieve sinusoidal output currents, controllable output power factor, independent MPPT and excellent efficiency. By having the potential of achieving high power density/reliability, extended reactive power control range and three-phase balanced output, the developed micro-inverter is an attractive candidate for AC PV module applications.

REFERENCES


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