New Control Method for Power Decoupling of Electrolytic Capacitor-less Photovoltaic Micro-Inverter with Primary Side Regulation

Mohammad Sameer Irfan*, Jong-Hyun Shin* and Joung-Hu Park†

Abstract – This paper presents a novel power decoupling control scheme with the bidirectional buck-boost converter for primary-side regulation photovoltaic (PV) micro-inverter. With the proposed power decoupling control scheme, small-capacitance film capacitors are used to overcome the life-span and reliability limitations of the large-capacitance electrolytic capacitors. Then, an improved flyback PV inverter is employed in continuous conduction mode with primary-side regulation for the PV power conditioning. The proposed power-decoupling controller shares the reference for primary side current regulation of the flyback PV inverter. The decoupling controller shapes the input current of the bidirectional buck-boost converter. The shared reference eliminates the phase-delay between the input current to the bidirectional buck-boost converter and the double frequency current at the PV primary current. The elimination of the phase-delay in dynamic response enhances the ripple rejection capability of the power decoupling buck-boost converter even with small film capacitor. With proposed power decoupling control scheme, the additional advantage of the primary-side regulation of flyback PV inverter is that there is no need to have an extra current sensor for obtaining the ripple-current reference of the decoupling current-controller of the power-decoupling buck-boost converter. Therefore, the proposed power decoupling control scheme is cost-effective as well as the size benefit. A new transient analysis is carried out which includes the source voltage dynamics instead of considering the source voltage as a pure voltage source. For verification of the proposed control scheme, simulation and experimental results are presented.

Keywords: Active power decoupling, Bidirectional buck-boost converter, Primary side regulation, Film capacitor

1. Introduction

Because of the limited source of fossil fuels, now, the researchers are focusing on alternate sources of energy. From the perspective of lifetime and maintenance cost of the sources, renewable energy sources are considered as one of the most preferred options. Among the usage of the renewable energy sources, the utilization of photovoltaic system has grasped a growth rate of higher pace in worldwide [1, 2]. In photovoltaic power generation system engineering industry, the main portion is assigned to the factor of product related with grid-connected power conditioning systems.

The PV systems for grid-connected inverter are categorized into three sub-divisions: central inverter, string inverter and AC-module inverters. Micro-inverter are preferred because of the enhanced maximum power point tracking (MPPT) efficiency of the system, lower installation cost, improved power scavenging, plug-n-play operation and better modularity and flexibility. There are different configurations of micro-inverter, shown in Fig. 1. In Fig. 1(a), the DC-link voltage is stepped up to a high voltage which is mainly because of a boost converter. The boosted voltage degrades the efficiency and the necessity of high

---

* Corresponding Author: Dept. of Electrical Engineering, Soongsil University, Seoul, Korea. (wait4u@ssu.ac.kr)
* Dept. of Electrical Engineering, Soongsil University, Seoul, Korea. (msameerifran@gmail.com, shin3082@naver.com)

† Corresponding Author: Dept. of Electrical Engineering, Soongsil University, Seoul, Korea.
frequency switching in the following-stage inverter degrades the power density of the system. In Fig. 1(b), the configuration is a type without DC-link. The secondary side is a cyclone converter, which suffers from the similar disadvantages mentioned in Fig. 1(a). In Fig 1(c), it is called “pseudo DC-link” type. In this configuration, the primary makes 120 Hz rectified voltage (or current) instead of DC waveform. Then, the secondary side full-bridge inverter just changes the current direction alternatively by the grid-frequency switching to convert the rectified waveform to a perfect AC. So, there are reduced losses due to the low frequency switching of the inverter. Hence, the configuration of Fig. 1(c) is preferred.

However, the main challenge in achieving the successful market share in PV industry is the limited lifetime and the reliability of the micro-inverter. It is because of the well-known factor of the life-time-limiting, poor reliability and, low-frequency bandwidth component in the inverter, electrolytic capacitors. Therefore, various inverter topologies have been proposed to eliminate the electrolytic capacitors and to replace them with long lifespan film capacitors [2-6]. A schematic diagram of the conventional power decoupling scheme for the film capacitor applications of the PV power conditioner in the pseudo DC-link configuration in Fig. 1(c), is shown in Fig. 2.

The conventional pseudo DC-link flyback inverter operates generally in discontinuous conduction mode or at boundary condition because of its simple control as it has no right half plane (RHP) zero in the output-current controller design. However, it suffers from a poor efficiency in these modes [7-9, 11]. From efficiency perspective, continuous conduction mode is preferred; however the control design became an issue to solve. Therefore, conventional topology commonly developed has been a DCM operation with secondary (rectified or AC) current sensing as shown Fig. 2. In [10], primary side regulation was proposed to overcome the control design problem for flyback PV inverter. Therefore, the primary side regulation of the flyback inverter has been considered as one of the state-of-the-art configurations in PV micro-inverter topology. Still, there is some reliability issue in the PV systems, electrolytic-less power decoupling methods have become a key research topic for the system.

Generally, with passive power decoupling methods, a
huge electrolytic capacitor is needed in PV inverter (position C in Fig. 2). However, electrolytic capacitors’ implementation offers disadvantages of short lifespan and low reliability. So, with electrolytic capacitors, micro-inverters suffer from the aforementioned disadvantages of electrolytic capacitors. The limitations caused by electrolytic capacitors can be overcome by power decoupling circuits which rejects the fundamental AC-output ripple using film capacitors, as film capacitors have a greater lifetime and higher reliability than those of electrolytic capacitors [3-6]. The basic idea of power decoupling circuit is to bypass the ripple power into alternative energy storage with competent size and long life span [1]. In power decoupling realization for grid-connection applications, the required information to ripple reject is the double frequency, 120Hz, component in the DC power rail (see Fig. 2). With the conventional power decoupling configuration, the switch current (i\textsubscript{sw}) sensing is needed for the double-frequency information in the current to compensate the discharging current of the capacitor (C in Fig. 2) by charging current of the decoupling circuit. However, the switch current contains DC, switching and double frequency component; therefore a band-pass filter is needed in the sensing process to extract the double frequency component information from the switch current. This filter produces a phase-delay to the reference of double frequency component information which reduces the power decoupling capability by a mismatch between the reference (\textit{i}_{ref}) and real current component (\textit{i}_{sw}).

The proposed power decoupling control scheme is shown in Fig. 3. In the proposed power decoupling control scheme, due to the primary-side regulation [10,12], the reference of the primary switch current regulation (\textit{i}_{ref}) can be shared with the power decoupling controller as a reference for controlling the compensating current of the bidirectional buck-boost converter in the power decoupling circuit (i\textsubscript{cap}). In the proposed power decoupling control scheme, there is no usage of a band-pass filter to extract the double frequency component information for the reference, and hence the mismatch from a phase-delay is eliminated which enhances the power decoupling capability of the controller scheme in very effective manner. Another advantage is, from the reference sharing, there is no need to add an extra current sensor to the flyback switch for the power decoupling controller to obtain the reference of the compensating current in the power decoupling circuit. Hence, an advantage of reducing the component and cost is achieved.

The hardware, 100W prototype, is tested to verify the proposed power decoupling control scheme. The paper is summarized as: section II describes the power decoupling concept with the bidirectional buck-boost converter, section III describes the transient analysis and controller design. Section IV includes simulation’s and experiment’s verification. Lastly, the paper is concluded in section V.

2. Power Decoupling Concept and Bidirectional Buck-Boost Converter

The instantaneous output current through the grid-connected single-phase inverter and the voltage across it are given as,

\[
\begin{align*}
\nu(t) &= V\sin(\omega t) \\
i(t) &= I\sin(\omega t)
\end{align*}
\]

where \(\omega\) is the grid frequency, and V and I are the voltage and current amplitudes, respectively. Then, the expression for instantaneous output power \(P_\text{d}(t)\) is given as,

\[
P_\text{d}(t) = \frac{1}{2}VI + \frac{1}{2}VI\cos(2\omega t)
\]

The expression for instantaneous power reveals that the power is oscillating in sinusoidal pattern around the average DC power, \(\frac{1}{2}VI\), with the double line-frequency in peak power value, \(\frac{1}{2}VI\). It can be observed in the Fig. 4. The double line frequency power oscillation degrades the system efficiency. So, to manage the power oscillation, an electrolytic high-capacitance decoupling capacitor is conventionally used. The decoupling capacitance is given by [1]:

\[
C = \frac{P}{2\pi f_{\text{grid}}V_p\Delta V_{pv}}
\]

where \(V_{pv}\) voltage across PV, \(\Delta V_{pv}\) - double frequency voltage ripple, \(f_{\text{grid}}\) - grid frequency, 60Hz and P-inverter output power. From (3), the decoupling capacitance is inversely proportional to \(V_{pv}\) and \(\Delta V_{pv}\). For example, the decoupling capacitor required for 100W power is of capacitance, 6.631mF, with 40V as voltage across PV and 1V as 120Hz component of voltage. The capacitance of such high value is conventionally available in electrolytic type capacitors, and electrolytic capacitors limit the lifespan and reliability of the circuit. For having a low capacitance, a power decoupling circuit is introduced with film capacitor as a power decoupling capacitor. The usage of film capacitor increases the reliability and lifetime of the circuit. Bidirectional buck-boost converter is preferred as the power decoupling circuit, as it has a low part count.

In configuration of a pseudo DC-link topology inverter, there is only one coupling node for having the power decoupling capacitor, the PV string output [2]. The schematic of flyback micro-inverter with proposed controller for bidirectional buck-boost converter as a power decoupling circuit is shown in Fig. 3. In order to perform power decoupling by the bidirectional buck-boost converter, the power should be transferred into the converter from the inverter when the instantaneous (but switching-frequency averaged) power (\(I_{sw}\) in Fig. 4) is smaller than the average power (\(I_{pv}\) in Fig. 4). When the instantaneous power is
greater than the average, the power should be transferred from the bidirectional buck-boost converter to the DC-link. To achieve the operation, therefore, the currents through the flyback inverter and through the bidirectional buck-boost converter should be as shown in Fig. 4. The input current to the flyback micro-inverter, $i_{sw}$, contains DC component and double frequency component. The input current to buck-boost converter, $i_L$, must be controlled in such a way that the PV current ($i_{PV}$) contains only DC component. The current equation for $i_{sw}$ is given as in Eq. (4)[10]. Therefore, the input current to buck-boost converter, $i_L$, is given as in Eq. (4), to eliminate the double frequency component from the PV source. For power decoupling, the buck-boost converter has two modes of operation, boost mode and buck mode, according to direction of the current, $i_L$. In boost mode, the decoupling capacitor charges, while in buck mode, the decoupling capacitor discharges. The capacitance is inversely proportional to the level of the average voltage and its instantaneous swing. Hence, for decrease of the power decoupling capacitance, both of the higher swing voltage and the average voltage are desirable in the design process.

Now let’s consider the practical cases such that there is some phase difference between the reference and the real buck-boost inductor current. In first, the current relationships between the PV, capacitor and the buck-boost inductor current are derived by node analysis. The results are as follows.

$$I_{sw} = k_{sw} \sin^2 \omega t$$
$$I_L = I_{sw} - I_{pv}$$

From Eq. (4), it reveals that the double frequency component of the inductor current and switch current must be in phase to get the power decoupling well performed. If there is phase delay between them, then it will produce a ripple in the PV current. Let the phase delay between them be $\varphi$, then the inductor current equation be as below in (5).

$$I_L = -\frac{k_{sw}}{2} \cos(2\omega t - \varphi)$$

From Eq. (4) and (5),

$$I_{pv} = \frac{k_{sw}}{2} + k_{sw} \sin\left(\frac{\varphi}{2}\right) \sin\left(2\omega t - \frac{\varphi}{2}\right)$$

Eq. (6) states that PV current is DC if there is no phase delay between $I_{sw}$ and $I_L$. If not, then the ripple current caused by the phase delay, is absorbed by the DC-link capacitance, $C_{dc}$ as Eq. (7).

$$I_c = k_{sw} \sin\left(\frac{\varphi}{2}\right) \sin\left(2\omega t - \frac{\varphi}{2}\right)$$

In phasor form, can be written as in Eq. (7a).

$$i_c = \frac{k_{sw}}{2} \sin\left(\frac{\varphi}{2}\right) \angle\left(-\frac{\varphi}{2}\right)$$

And, hence, $v_c = \frac{k_{sw}}{2\omega C_{dc}} \angle\left(-\frac{\varphi}{2} - 90\right)$

In Eq. (7b), $v_c$ represents the voltage ripple corresponding to the current ripple by phase delay. Therefore, the DC-link capacitance can be obtained by the following Eq. (8).

$$C_{dc} = \frac{k_{sw} \sin^{2} \frac{\varphi}{2}}{2\omega \Delta V_c}$$

Next section, the buck-boost current and voltage controller design procedure will be introduced using the state space averaging technique for the small-signal analysis.

3. Transient Analysis and Controller Design

3.1 Modelling of Flyback inverter in primary-current loop

The control block diagram for flyback inverter is shown in Fig. 5. From the block diagram, the loop gain of the flyback switch current controller is presented as Eq. (9).

$$T_{fly}(s) = G_{swd}(s)S_f G_{clamp}(s)FM$$

Then, the closed loop input impedance of the flyback inverter can be obtained as in Eq. (10).

![Fig. 4. Power Decoupling concept with buck-boost converter as a power decoupling circuit.](image-url)

![Fig. 5. Control Block Diagram for flyback switch current](image-url)
\[ \frac{\bar{v}_{pv}}{i_{sw}} = Z_{in}f_{ly}(s) \left( 1 + T_{fly}(s) \right) = Z_{fly}(s) \] (10)

The loop gain, \( T_{fly}(s) \), is usually designed high in bandwidth which implies that the closed loop impedance, \( Z_{fly}(s) \), is high. So, with a buck-boost converter as a power decoupling circuit, the flyback inverter behaves as high impedance compared to the PV resistance \( R_{in} \). Therefore, the flyback inverter’s influence on the buck-boost converter is negligible.

3.2 State-space average model of buck-boost converter

In this state-space averaged model, the PV-source dynamic effect is considered. In analysis, equivalent circuit mode of the PV source is taken as a voltage source, \( V_{pv} \), in series with a resistor, \( R_{in} \). For taking into account the power flow through the buck-boost converter, a load resistance \( R \) is considered at the output of buck-boost converter.

The state variables of the buck-boost converter are DC-link voltage, \( v_{pv} \), input (inductor) current to the converter, \( i_{L} \), and the output voltage, \( v_{opd} \). All the state-variables are defined in Fig. 3. The buck-boost converter has two operational modes I, on-time, and II, off-time [13]. The state-equations for the two modes are as follows:

**Mode I**

\[
\begin{align*}
dv_{pv} \over dt &= -\frac{1}{C_{dc}}i_{L} - \frac{1}{R_{in}C_{dc}}v_{pv} + \frac{1}{R_{in}C_{dc}}v_{in} \\
dv_{opd} \over dt &= -\frac{1}{C_{pd}}i_{L} - \frac{1}{R_{in}C_{pd}}v_{opd}
\end{align*}
\] (11)

**Mode II**

\[
\begin{align*}
dv_{pv} \over dt &= \frac{1}{L}v_{pv} \\
dv_{opd} \over dt &= -\frac{1}{C_{pd}}i_{L} - \frac{1}{R_{in}C_{pd}}v_{pv} + \frac{1}{R_{in}C_{dc}}v_{in}
\end{align*}
\] (12)

From (11) and (12), the averaged model equation is in (13),

\[
\begin{bmatrix}
i_{L} \\
v_{pv} \\
v_{opd}
\end{bmatrix}
= A \begin{bmatrix}
i_{L} \\
v_{pv} \\
v_{opd}
\end{bmatrix}
+ B \begin{bmatrix} v_{in} \end{bmatrix}
\]

where \( A, B \) and \( C \) are given in Eq. (14),

\[
A = \begin{bmatrix} 0 & 1 & -\frac{d}{L} \\
-\frac{1}{C_{dc}} & -\frac{1}{R_{in}C_{dc}} & 0 \\
\frac{d}{C_{pd}} & 0 & -\frac{1}{R_{in}C_{pd}} \end{bmatrix}
\]

\[
B = \begin{bmatrix} 0 \\
1 \\
0 \end{bmatrix}
\]

and \( C = \begin{bmatrix} 1 & 0 & 0 \\
0 & 1 & 0 \\
0 & 0 & 1 \end{bmatrix} \)
(14)

3.3 Small signal model

From the generalized state-space average model, the small-signal model can be obtained. The small signal model is obtained in (16) and (17) by having perturbation around the variables, state-variables (\( v_{opd} \), \( i_{L} \), and \( v_{opd} \)), control variable (\( d \)) and input variable (\( v_{in} \)) as shown in (15).

\[
\begin{align*}
v_{in} &= V_{in} + \bar{v}_{in} and |\bar{v}_{in}| \ll V_{in} \\
v_{pv} &= V_{pv} + \bar{v}_{pv} and |\bar{v}_{pv}| \ll V_{pv} \\
i_{L} &= i_{L} + \bar{i}_{L} and |\bar{i}_{L}| \ll i_{L} \\
d &= D + \bar{d} and |\bar{d}| \ll D \\
v_{opd} &= V_{opd} + \bar{v}_{opd} and |\bar{v}_{opd}| \ll V_{opd}
\end{align*}
\] (15)

\[
\begin{bmatrix}
i_{L} \\
v_{pv} \\
v_{opd}
\end{bmatrix}
= \begin{bmatrix} 0 & 1 & -\frac{D}{L} \\
-\frac{1}{C_{dc}} & -\frac{1}{R_{in}C_{dc}} & 0 \\
\frac{D}{C_{pd}} & 0 & -\frac{1}{R_{in}C_{pd}} \end{bmatrix}
\begin{bmatrix}
\bar{i}_{L} \\
\bar{v}_{pv} \\
\bar{v}_{opd}
\end{bmatrix}
\] (16)

\[
\begin{bmatrix}
\bar{i}_{L} \\
\bar{v}_{pv} \\
\bar{v}_{opd}
\end{bmatrix}
= \begin{bmatrix} 1 & 0 & 0 \\
0 & 1 & 0 \\
0 & 0 & 1 \end{bmatrix}
\begin{bmatrix}
i_{L} \\
v_{pv} \\
v_{opd}
\end{bmatrix}
\] (17)

The transfer functions, \( \frac{\bar{i}_{L}}{d} \), \( \frac{\bar{v}_{pv}}{d} \) and \( \frac{\bar{v}_{opd}}{d} \), are obtained in Eq. (18), (19) and (20), respectively.

3.4 Controller design

The control block diagram of the bidirectional buck boost converter for the power decoupling is shown in Fig. 6. The proposed power decoupling scheme needs two loop controllers. The inner current loop is for the inductor current to track the current reference (\( i_{L,ref} \)) from the flyback primary so that the voltage ripple across PV is
minimized. The outer voltage loop is to prevent the average voltage \( V_{\text{opd}} \) of the decoupling capacitor from passing over the safe level.

From Fig. 4, it can be seen that the main concern of the controller design for power decoupling is based on the currents dynamics. For regulation of the input current, or the inductor current \( i_L \), in the bidirectional buck-boost converter, the controller design should guarantee the stability and the bandwidth. In addition to the current control loop, the voltage loop of the decoupling capacitor is required to be slower than the inductor current loop by at least an order in bandwidth for preventing the dynamic interaction. In the controller design procedure, the current control loop is designed first.

The current controller is shown in Fig. 7. The controller has three poles and two zeroes. The controller is supposed to be designed to track double line frequency, 120 Hz, without phase delay, so that the loop-gain bandwidth should be at least 10 times higher than the line frequency. By the open-loop transfer function, \( G_{\text{id}}(s) \), the poles and zeroes of the controller in (21) can be obtained to satisfy the phase margin and the bandwidth for guaranteeing the dynamic response and stability.

\[
\begin{align*}
   f_{x1} &= \frac{1}{2\pi C_1(R_1 + R_{11})} \\
   f_{x2} &= \frac{1}{2\pi C_2 R_2} \\
   f_{p1} &= \frac{1}{2\pi C_1 R_1} \\
   f_{p2} &= \frac{1}{2\pi R_2(C_2 || C_3)}
\end{align*}
\]

(21)

The transfer function of the current controller, three poles and two zeroes controller, is given in Eq. (22).

\[
G_{\text{ci}}(s) = \frac{(1 + sC_1(R_1 + R_{11}))(1 + sC_2R_2)}{[sR_1(C_2 + C_3)][1 + sC_1R_1][1 + sR_2(C_2 || C_3)]}
\]

(22)

Then the final current loop gain is determined as (17).

\[
T_{L_i}(s) = G_{\text{id}}(s)G_{\text{ci}}(s)S_e
\]

(23)

where \( S_e \) is the sensing gain.

The frequency response of the \( G_{\text{id}}(s) \) and the close-loop gain, \( T_{L_i}(s) \), are shown in Fig. 8. From the frequency response of loop gain, the system is stable as it has a phase margin of 79.9° which is greater than 60° as required for stability. The current controller has a bandwidth of 1.84 kHz.
The parameter values for the frequency response and controller parameters are shown in Table 1 and Table 2, respectively. After designing the current controller, the next step is the voltage controller to maintain the DC component of decoupling capacitor voltage at a specified value. Aforementioned, the controller should be slow as compared to the current controller. And also, the voltage controller should pass only the DC component. So, a low pass filter is needed to pass the component of the decoupling capacitor voltage.

The cut-off frequency of the low-pass filter is 159 Hz. The voltage controller is integral controller for tracking the DC component of the voltage. The frequency response of the voltage loop gain is shown in Fig. 9. From the frequency response of the loop gain, the voltage controller is slow as compared to current controller as it has bandwidth of 7 Hz.

For checking the robustness of parameter variations, the frequency response of the loop gain is shown in Fig. 10 for ±10% change in $C_{dc}$ and $L$. From the corresponding frequency response, it reveals that the change in the DC-link capacitance and inductance of buck-boost converter has not much impact on the phase margin, 79.9°, and hence the controller is stable from the variation aspect of DC-link capacitance and the inductance. The transfer function of flyback switch current has single pole [10], so the transformer parameters have no influence on the controller stability.

The ability of the controller for power decoupling concept can be known by the input impedance for closed-loop of the buck-boost converter. The impedance of the buck-boost converter can be obtained from the control block diagram. The expression of the closed-loop impedance is as in (24).

$$Z_{\text{closed}} = Z_{\text{in}}(s)$$

(24)

The frequency response of the closed loop input impedance of the buck-boost converter is shown in Fig. 11 and indicating the value of $Z_{\text{closed}}(s)$ as 10.3dB at 120Hz. The value of the input impedance of buck-boost converter is very small as compared to PV-source resistance, $R_{\text{in}}$, in parallel with the converter. It means that the 120Hz current
component will flow through the power decoupling circuit with the smaller impedance than that of the PV side. Therefore, it confirms the effectiveness of the controller for the buck-boost converter as power decoupling circuit.

4. Verification

4.1 Verification of model by simulation

The verification of the transfer functions, \( G_{id}(s) \), \( G_{vpv}(s) \) and \( G_{vopd}(s) \), can be done by comparison between their frequency responses from the mathematical model derivations and the numerical PSIM AC-sweep simulations of the bidirectional buck-boost converter. The parameters for the frequency response from the numerical simulation are given in Table 3. The frequency responses from the expression and the numerical simulations are shown in Fig. 12 and Fig. 13, respectively. The similarity of the corresponding frequency responses of \( G_{vopd}(s) \) and \( G_{vpv}(s) \) validates the expressions for bidirectional buck-boost converter. The similarity of the corresponding frequency responses of \( G_{id}(s) \) also validates the expressions for bidirectional buck-boost converter. PSIM software is used to simulate the proposed power control scheme. The parameters for the simulation are given in Table 3. The key simulation waveforms for the
Table 3. Parameters for simulation

<table>
<thead>
<tr>
<th>Symbol</th>
<th>QUANTITY</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( V_{in} )</td>
<td>Source Voltage</td>
<td>60V</td>
</tr>
<tr>
<td>( R_{in} )</td>
<td>Source resistance</td>
<td>10Ω</td>
</tr>
<tr>
<td>( C_{dc} )</td>
<td>Capacitance</td>
<td>200µF</td>
</tr>
<tr>
<td>( L_{m} )</td>
<td>Magnetizing Inductance</td>
<td>100µH</td>
</tr>
<tr>
<td>( n )</td>
<td>Turn ratio of the flyback transformer</td>
<td>4</td>
</tr>
<tr>
<td>( V_o )</td>
<td>Output voltage of inverter</td>
<td>220V(rms)</td>
</tr>
<tr>
<td>( f_{grid} )</td>
<td>Grid frequency</td>
<td>60 Hz</td>
</tr>
<tr>
<td>( C_{pd} )</td>
<td>Decoupling Capacitance</td>
<td>15 µF</td>
</tr>
</tbody>
</table>

Fig. 14. Key waveforms of the simulation results (From top: inductor current and the reference, switch current and the reference, decoupling capacitor voltage, PV inverter output, PV film capacitor voltage)

Fig. 15. Hardware prototype for the experimental test

proposed power decoupling control scheme are shown in Fig. 14. From the figure, the DC component of the decoupling capacitance voltage (\( V_{opd} \)) is maintained at 150V with a peak-to-peak voltage swing of 150 V. The PV capacitor (\( V_{PV} \)) has the voltage swing of 1 V, in spite of the small capacitance, which means the power decoupling capacitor absorbs well the ripple power from the PV capacitor through the power-decoupling buck-boost. The small capacitance allows the capacitor can be replaced by film capacitors.

4.2 Verification by experimental results

Fig. 15 shows the hardware test set-up for the experimental verification. Fig. 16 shows the experimental waveforms of the converter for 100W system. The agreement among the corresponding waveforms of simulation and experiment successfully validates the analysis and controller design guidelines. The switch current, \( I_{sw} \), which is low-pass-filtered to remove the switching-frequency component follows well the current reference in the Fig. 16(a) and, also, the inductor current, \( I_L \), is well following its reference in Fig. 16(b). From the equation of the power decoupling capacitance (3), the voltage ripple across the decoupling capacitor is 115 V with the decoupling capacitance as 15µF. The PV capacitor’s average voltage is 30V with ripple of 1.5 V. From the experimental waveforms, it can be seen that the operational behavior is stable with the proposed controller. The power decoupling circuit with proposed controller has a reduced capacitance of 15µF from 6.6mF. Therefore, the controller has fully shown its effectiveness for power decoupling concept.

The experiment waveforms of the instantaneous flyback switch current, PV current and the inductor one are shown in Fig. 17. The flyback switch current is switching...
waveforms, not filtered, and the inductor current has almost the same phase as the flyback, so the PV current is almost DC waveform. The waveforms verify the theoretical power-decoupling behavior as in Fig. 4.

5. Conclusion

A novel power decoupling control scheme for primary-side regulated flyback PV inverter with the bidirectional buck-boost as a power decoupling circuit is presented in this paper. The transient analysis with the PV-source dynamics is carried out to have a more accurate result in controller design. Film capacitors are implemented to overcome the life-span and reliability limitations of electrolytic capacitors.

In the proposed control scheme, the reference for primary-side regulation of flyback PV inverter has been used as the reference of input current of the bidirectional buck-boost converter. Due to the utilization of the primary-side current reference of flyback PV inverter, in the input current of the bidirectional buck-boost converter, the phase-difference of the current is removed.

It enhances the power decoupling capability to a great extent as observed in the simulation and experimental results. Moreover, there is an additional advantage of removal of one current sensor with primary-side regulation of the flyback PV inverter which is needed to have the reference for input current of the bidirectional buck-boost converter in conventional scheme. Therefore, the proposed control scheme is cost and size effective. The design guidelines are presented by using a small signal modeling. With the proposed power-decoupling control scheme, the PV capacitor voltage double frequency ripple is reduced to less than 5% on 30V average voltage by having power decoupling capacitance as 15µF. It shows the effectiveness of the proposed control scheme. The simulation and experimental results are shown and both the results are in agreement and validating the proposed control scheme for power decoupling application. With the bidirectional buck-boost converter as power decoupling circuit, the proposed power decoupling control scheme are successfully verified as a promising candidate.

Nomenclature

- $V_{pv}$: PV voltage
- $R_{in}$: PV source resistance
- $i_{pv}$: PV current
- $i_{sw}$: Flyback switch current
- $C_{dc}$: DC link Capacitance
- $L_m$: Magnetizing Inductance
- $i_L$: Inductor Current
- $i_{L-ref}$: Reference current for buck-boost Inductor
- $L$: Buck-Boost Inductor
- $C_{pd}$: Power Decoupling capacitance
- $V_{pd}$: Voltage across $C_{pd}$
- $V_{pd-ref}$: Reference for average voltage across $C_{pd}$
- $G_{ld}(s)$: Control to $i_L$ transfer function of buck-boost converter
- $G_{vopd}(s)$: Control to input voltage of the Buck-boost converter transfer function
- $G_{vout}(s)$: Control to output voltage of the buck-boost converter transfer function
- $G_{iL}(s)$: Input voltage to current transfer function of the buck-boost converter
- $L_{pd}$: Current Controller for buck-boost converter
- $LPF$: Low pass filter
- $Z_{in}(s)$: Open loop input impedance of the buck boost converter
- $Z_{in-closed}(s)$: Closed loop input impedance of the buck boost converter
- $G_{vbd}(s)$: Voltage controller for the buck boost converter
- $Z_{in-bd}(s)$: Open loop input impedance of flyback inverter
- $T_{pd}(s)$: Loop gain of flyback switch current controller
- $Z_{in-closed}(s)$: Closed loop input impedance of flyback inverter
- $G_{isvd}(s)$: Control to flyback switch current transfer function
- $G_{isvd}(s)$: PV voltage to flyback switch current transfer function
- $G_{fly}(s)$: Flyback switch current controller

Acknowledgment

This work was supported by “Human Resources Program in Energy Technology” of the Korea Institute of Energy Technology Evaluation and Planning, granted financial resource from the Ministry of Trade, Industry & Energy, Republic of Korea. (No. 20164010201010)

References

Mohammad Sameer Irfan, Jong-Hyun Shin and Joung-Hu Park

Mohammad Sameer Irfan received B.Tech (Electrical Engineering) and M.Tech (Power System and Drives) from Aligarh Muslim University, Uttar Pradesh, India, in 2004 and 2008 respectively. He is currently pursuing PhD in Soongsil University, Seoul, South Korea. His current research interests include analysis, design and applications of dc-dc converters and renewable energy applications.

Jong-Hyun Shin received his B.S. and M.S degree in 2013 and 2015 from the Department of Electrical Engineering of Soongsil University, Seoul, Korea. His current research interests include the analysis of high-frequency switching converters.

Joung-Hu Park received the B.S., M.S., and Ph.D. degrees in electrical engineering from the Department of Electrical Engineering and Computer Science, Seoul National University, Seoul, South Korea, in 1999, 2001, and 2006, respectively. He is currently an Associate Professor at Soongsil University, Seoul, South Korea. From August 2004 to August 2005, he was a Visiting Scholar at Virginia Tech, Blacksburg, VA, USA, and from July 2015 to June 2016, he was a Visiting Assistant Professor at the University of British Columbia, Vancouver, Canada.