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Current-fed dual-half-bridge converter directly connected with half-bridge inverter for residential photovoltaic system

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ABSTRACT

A current-fed dual-half-bridge (CF-DHB) converter directly connected with a half-bridge (HB) inverter unit is proposed for residential photovoltaic power conversion systems. The proposed inverter requires only half as many switching devices as the full-bridge two-stage inverters and does not have to use a large dc-link capacitor. However, the converter unit suffers from 120-Hz dc-link voltage fluctuation and 60-Hz capacitor voltage unbalance, which degrade the control performance of the half-bridge inverter. To suppress both dc-link voltage fluctuation and capacitor voltage unbalance, we propose to use appropriate repetitive controllers, that are used to determine the nominal duty-ratio and the phase-shift. To make the proposed CF-DHB operate at the maximum power point (MPP), we use the incremental conductance method that offers smooth transition to MPP. Experimental results show that the proposed inverter system achieved high MPP tracking efficiency, low total harmonic distortion, high power conversion efficiency, and medium power capacity.

1. Introduction

Photovoltaic (PV) energy has experienced impressive growth over the past decade due to fast depletion of fossil fuels, the concern of energy security and the greenhouse gas emission problem. PV electricity generation is scalable from small-scale residential application to large-scale solar farms/power plants. Small-scale PV electricity generation in the United States alone reached 19,467 GWh in 2016, and is expected to achieve 32,900 GWh in 2018 (U.S. Energy Information Administration, 2017). More than half of the growth in small-scale system is occurring in the residential sector, with about 32% in the commercial sector and 8% in the industrial sector. Residential PV systems have a capacity of about 0.5–3 kW, and are usually mounted on the rooftop of a residential building. The number of such installed devices is increasing rapidly (Kabir et al., 2014; Worthmann et al., 2015; Kensby et al., 2015; Wang et al., 2016; Chub et al., 2018; Liu et al., in press). They use PV modules that produce unregulated dc electricity, and power converters that convert it to useful ac outputs.

Various front-end converter with back-end inverter topologies have been proposed for residential PV systems. Among the DC/DC converters used for residential photovoltaic system, the full-bridge type current-fed dual-active-bridge (CF-DAB) converter (Shi et al., 2015; Sha et al., 2016; Shi and Li, 2017; Bal et al., 2018) is considered as the attractive one due to its low-input current ripple and high conversion efficiency. Compared to the full-bridge type CF-DAB converter, however, the current-fed dual-half-bridge (CF-DHB) converter (Meghdad et al., 2015; Bai et al., 2017) is more cost-effective because it requires fewer power components. On the other hand, if we continue to use the full-bridge inverter unit in the back-end, four active components are still required. To reduce the number of these active components by half, we propose to use the HB inverter in the back-end. Moreover, instead of using large dc-link electrolytic capacitors, we adopt the secondary-side upper/lower film capacitors in the proposed CF-DHB converter with HB inverter scheme, so cost is further reduced.

However, when the CF-DHB converter directly connected to the HB inverter is used in the grid-connected environment, the pulsating power caused by the ac grid makes the dc-link voltage and current fluctuate with twice the grid frequency. Moreover, split dc-link capacitor voltages become unbalanced because the upper capacitor charges and the lower capacitor discharges during the positive half-cycles, and they operate in reverse during the negative half-cycle. The 120-Hz dc-link voltage fluctuation and 60-Hz capacitor voltage unbalance complicate the task of designing a feedback controller for the inverter stage. Furthermore, the dc-link voltage must be measured when we design the feed-forward controller for the inverter stage. To suppress both 120-Hz dc-link voltage fluctuation and 60-Hz split capacitor voltage unbalance,

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we can use a high-gain proportional-integral (PI) controller (Chang et al., 2015) that increases the gains around 60-Hz and 120-Hz but it also increases the system gain in the high-frequency region and decreases the phase margin. To overcome this problem, the proportional resonant (PR) controller was developed (Li et al., 2015; Kuperman, 2015; Lee et al., 2016; Busada et al., 2018); it acts as a notch filter to eliminate the grid voltage disturbance at the fundamental frequency, but it provides peak gain only at selected resonant frequencies, so to track the grid fundamental frequency and suppress the subharmonic frequencies, multiple PR controllers must be used in parallel; this fact increases the complexity in digital implementation and causes heavy computation burden. Moreover, the PR controller constructs the control input by using the current error only, so the control accuracy is not much improved.

To solve these problems, we introduce repetitive control algorithms coupled with PI controllers to be used for the proposed CF-DHB converter directly connected with HB inverter. It is based on controlling the phase-shift and duty-ratio of the CF-DHB converter. To this aim, we first derive the average model of the directly-connected CF-DHB converter system, and use it to implement the repetitive controller. To make the inverter operate at the maximum power point (MPP), we use the incremental conductance method to achieve smooth transition to MPP. Tests on a prototype of the proposed control scheme verify that it achieves 99.5% maximum efficiency of MPP tracking, with low total harmonic distortion, high power conversion efficiency, and medium power capacity.

This paper is organized as follows. We present the development of the PV panel model, the directly-connected CF-DHB converter model, and the analysis on dc-link voltage fluctuation and capacitor voltage unbalance in Section 2, and propose an efficient control scheme to be implemented on the directly-connected CF-DHB converter in Section 3. We present the experimental setup and results in Section 4 and draw conclusions in Section 5.

2. Preliminaries and problem formulation

Conventional current-fed two-stage converters (Fig. 1a and b) feature low-input current ripple and high conversion efficiency, but these converters require many active components and large electrolytic capacitor. Proposed converter (Fig. 1c) takes the advantages of low-input current ripple and high conversion efficiency with smaller number of active components and no use of the electrolytic capacitor. The proposed grid-connected PV system is composed of two parts: the CF-DHB converter part and the HB inverter part. The CF-DHB converter boosts the low input voltage to high output voltage, and the HB inverter transforms the DC output of the CF-DHB converter to AC power with the line frequency. The CF-DHB converter part consists of input inductor L_{in}, low voltage side (LVS) switches S_1, S_2, capacitors C_1, C_2, ideal transformer T with turns ratio $n = N_{tp}/N_{ts}$, transformer leakage inductor L_s, high voltage side (HVS) switches S_3, S_4, capacitors C_3, C_4. While the power flows from LVS to the HVS, the circuit operates in boost mode to keep the HVS voltage at a desired high value; whereas the power flows from HVS to LVS, the circuit operates in buck mode to absorb the regenerated energy. The half-bridge inverter consists of two switches T_1, T_2, filter inductor L_f, grid side inductor L_g, and filter capacitor C_f. v_g is grid voltage and i_g is grid current.

If the transformer turns ratio is $N_{tp}/N_{ts} = 1$, then by simply regarding the transformer as leakage inductance L_s, we obtain the equivalent circuit of the CF-DHB converter (Fig. 2). The circuit uses the transformer leakage inductance L_s as the energy transfer element and interface between LVS and HVS half-bridge units. The gate signals for upper and bottom switches of each bridge are complementary and the amount of transferred power is determined by the phase-shift angle ϕ between the switches of LVS and HVS half bridge units.

2.1. PV panel modeling

The PV panel provides dc power to the proposed inverter, which converts this dc power to grid-compatible ac power. But, the voltage-current characteristic of the PV panel is highly nonlinear (Fig. 3a), and it needs to be analyzed first to design an appropriate control system for the proposed inverter (Kim et al., 2016). Therefore, we first derive the equivalent circuit of the PV panel (Fig. 3b). Applying Kirchoff's current law at Node 1 yields the output current I_{PV} [A] of the PV panel as (Bose et al., 1985; Hua et al., 1998; Hamidon et al., 2012)

$$I_{PV} = N_p I_{LG} - N_p I_D - I_{RSH}, \tag{1}$$

where I_{LG} [A] is the light generated current, I_D [A] is the diode current, and I_{RSH} [A] is the current of the equivalent shunt resistance. These parameters are given as

$$I_{LG} = I_{SCR}\frac{G}{1000\ \mathrm{W/m^2}} + J_o(T_c - T_{ref}), \tag{2}$$

$$I_D = I_o\left[\exp\left\{\frac{q\left(V_{PV} + \frac{N_s R_s I_{PV}}{N_p}\right)}{nkT_c N_s}\right\} - 1\right], \tag{3}$$

$$I_{RSH} = \frac{V_{PV} + I_{PV}\frac{N_s}{N_p}R_s}{\frac{N_s}{N_p}R_{SH}}, \tag{4}$$

where I_{SCR} [A] is the short circuit current at the reference state; G [W/m^2] is the solar irradiance; J_o [A/K] is the temperature coefficient; T_c [K] is the cell temperature; T_{ref} [K] is the reference temperature; I_o [A] is the inverse saturation current; $q = 1.602 \times 10^{-19}$ C is the charge on an electron; N_s is the number of PV cells in series; N_p is the number of PV cells in parallel; n [unit] is the diode quality factor; $k = 1.381 \times 10^{-23}$ J/K is the Boltzmann constant; R_s [Ω] is the equivalent series resistance; R_{SH} [Ω] is the equivalent shunt resistance; I_{PV} [A] is the PV panel current and V_{PV} [V] is the PV panel voltage. Substituting (2)–(4) into (1) yields

$$I_{PV} = N_p I_{LG} - N_p I_o\left[\exp\left\{\frac{q\left(V_{PV} + \frac{N_s R_s I_{PV}}{N_p}\right)}{nkT_c N_s}\right\} - 1\right] - \frac{N_p V_{PV} + N_s I_{PV} R_s}{N_s R_{SH}}. \tag{5}$$

The linearized PV model at the MPP can be obtained by taking the derivative of the I–V curve (5) with respect to V_{PV} as in Villalva and Ruppert (2009) and is given as

$$\begin{aligned}\frac{dI_{PV}}{dV_{PV}} &= f(I_m, V_m)\\ &= -\frac{qI_oN_p}{nkT_cN_s}\left(1 - \frac{N_sR_s}{N_p}\frac{I_m}{V_m}\right)\left[\exp\left\{\frac{q\left(V_m + \frac{N_sR_sI_m}{N_p}\right)}{nkT_cN_s}\right\} - 1\right] - \frac{N_p}{N_sR_{SH}}\left(1 - \frac{N_sR_s}{N_p}\frac{I_m}{V_m}\right),\end{aligned} \tag{6}$$

where I_m [A] is the PV current at MPP and V_m [V] is the PV voltage at MPP. The equivalent model of the PV panel at MPP (Fig. 3b) (Kim et al., 2016) can be linearized at MPP as shown in the input part of Fig. 2 and is given as

$$V_{in} = \frac{1}{f}I_{in} + \left(V_m - \frac{1}{f}I_m\right), \tag{7}$$

where V_{in} [V] is the equivalent voltage of the linearized model and I_{in} [A] is the equivalent current of the linearized model and $1/f$ is denoted as the input resistance R_{in} [Ω].