Thermal analysis and reliability evaluation of cascaded H-bridge MLPVI for grid-connected applications

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Abstract: Nowadays, great progress has been made in the development of multilevel inverters (MLIs) in grid-connected photovoltaic (PV) energy systems, because of the advantages such as reduced voltage stress on the power semiconductor switches and having higher efficiency. Since the multilevel PV inverter (MLPVI) has been a critical part within the failures of a PV energy system, it becomes important to predict the lifetime of the components and the MLPVI system. In this paper, a five-level modular cascaded H-bridge MLI is analyzed based on grid-connected application for PV energy system, and an independent dc-link voltage controller for each H-bridge is implemented by taking the reference value generated from the maximum power point tracking algorithm of each PV module. The loss and thermal distributions of the different power devices in the MLPVI system are investigated and illustrated for various pulse-width modulation (PWM) controllers. The junction temperature and power losses of the components in the MLPVI are simulated by MATLAB/Simulink and Piecewise Linear Electrical Circuit Simulation (PLECS) blockset, which validate the theoretical analysis. The reliability of the MLPVI is evaluated using parts stress method. It shows that the MLPVI has a good reliability when phase shift PWM control technique is used.

1 Introduction

From the past few years, grid-connected photovoltaic (PV) systems are the higher developing solar energy applications due to ever increasing energy demand and the exhaustive nature of fossil fuel resources and its increased prices [1, 2]. Solar power is the most common and available renewable power source to meet our rapidly increasing energy requirements. In these systems, the use of all the available energy depends on the static converter topology that is used in it. On the basis of inverter topology and PV module configuration, the grid-connected PV systems are categorised into five groups: centralised, string, multistring, ac-module inverters and cascaded inverters [3].

The grid-connected multilevel PV inverter (MLPVI) systems are very popular, because multilevel inverters (MLIs) have some attractive features such as: high efficiency with low switching frequency control methods, draw input current with low distortion and possibility of eliminating the step-up transformer in the high-voltage grid-connected applications [4, 5]. Additionally, cascaded MLI need several dc sources which can be derived from PV string, and also making possible of the independent voltage control and the tracking of the maximum power point (MPP) in each string. This characteristic has some advantages. On the one hand, reduce the power loss of the PV system in case of mismatch in the strings due to unequal solar radiation. On the other hand, the energy harvested from PV panels can be maximised [6].

Since the MLPVI has been a critical part of the failures of a PV system, the power switching devices and dc-link capacitors are the main components present in the PV inverter. The main factors to get a failure of these devices are thermal stress, electrical stress, mechanical stress and so on. From these factors, the most frequently observed failure mechanisms are related to thermal stresses [7], so it is paramount important to evaluate the thermal performance of the power switching devices and consequently predict the decreased lifetime of the PV inverter and its components.

In this paper, thermal analysis and reliability evaluation of single-phase-cascaded H-bridge MLPVI has been conducted according to the procedure as shown in Fig. 1. A simple control scheme is presented based on the controller presented in [8, 9] for achieving an individual MPP tracking (MPPT) for each PV array. The perturb and observe algorithm is used [10] to obtain the MP from each PV panel. This is the most commonly used algorithm because of its simple implementation. To obtain the MP from each PV array, the dc voltages of each array are individually controlled. The control signals driving each H-bridge is designed using phase shift pulse-width modulation (PSPWM) and level SPWM (LSPWM) control techniques. Conduction and switching losses are evaluated, and also the thermal analysis has been done using the PLECS based on the information provided on the datasheet.

A reliability analysis of the MLPVI under PSPWM and LSPWM control techniques has been carried out using parts stress method [11].

2 Control of the single-phase grid-connected MLPVI

The single-phase MLI consists of \( n \) H-bridge converters connected in series, dc-link of each H-bridge is fed by a short string or PV panel. The inverter can produce the output voltage waveform with \( n \) levels, this enables the harmonic reduction in the generated current and reduces the filter size due to high-quality voltage waveform.

In this paper, the control scheme proposed in [9] is used for this application as shown in Fig. 1. The sum of two dc-link voltages \( v_{L1} \) and \( v_{L2} \) is controlled by a proportional–integral (PI) that determines the amplitude of the input current \( i_{S2} \). A suitable reference for the current loop is obtained by multiplying the output of this controller with a normalised sinusoidal signal in phase with the voltage grid. On the other hand, the PI current controller gives the sum of the continuous switching functions \( S_1 + S_2 \).
The control of the second H-bridge voltage $v_{c2}$ is made through another controller. This scheme sets the phase of $S_2$ equals to grid phase. The power factor can be controlled either by changing the magnitude and phase of the voltage of the first cell or the magnitude of the voltage of the second cell. The total control part is divided into three sections, namely total voltage loop, second cell voltage loop, and total current loop. The three control loops are shown in Fig. 2.

For the independent control of each dc-link voltage, two PI controllers are necessary. To design the controllers, a suitable transfer functions are obtained by the linearisation of the dynamic behaviour of the system. In this case, it will be considered that the system operates at a nominal radiation of 1 kW/m² and at 25°C. As a first step, the transfer function of the loop that considers the total dc-link voltage will be calculated. The derivation of this expression is documented in [8]. Adding these two equations yields

$$S_1i_s + S_2i_s = i_{PV1} + i_{PV2} - C_1\frac{dv_{C1}}{dt} - C_2\frac{dv_{C2}}{dt}$$  \hspace{1cm} (1)

By considering only the dc component of the term $S_1i_s + S_2i_s$, (1) becomes

$$S_1^*i_s + S_2^*i_s = i_{PV1} + i_{PV2} - C_1\frac{dv_{C1}}{dt} - C_2\frac{dv_{C2}}{dt}$$  \hspace{1cm} (2)

To simplify the transfer function, the currents of the PV panels $i_{PV1}$ and $i_{PV2}$ will be considered as disturbances and the term $S_1^* + S_2^*$ constant. Under this assumption, the following transfer function is obtained:

$$\frac{v_{C1}(s) + v_{C2}(s)}{I_s(s)} = -\frac{S_1^* + S_2^*}{2C_s}$$  \hspace{1cm} (3)

where $C_1=C_2=C$ and the term $S_1^* + S_2^*$ is defined for nominal operating conditions, as it is indicated in [9]. The second voltage control loop is necessary to adjust the voltage difference between both dc-link voltages. In this case, it will be assumed that the current magnitude delivered by the first control loop is constant; thus

$$\frac{v_{C2}(s)}{I_s(s)} = -\frac{I_s^*}{2C_s}$$  \hspace{1cm} (4)

where $I_s^*$ is defined for nominal steady-state conditions, as it is shown in [9]. The schemes for the two voltage loops are shown in Figs. 2a and b. The design of the three PI controllers has been carried out with root locus method in the Laplace domain [12]. The current controller was designed using a bandwidth of 400 rad/s, which is high enough to provide an adequate current

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**Fig. 1** Flowchart of the reliability analysis of MLPVI system

**Fig. 2** Voltage and current loops of the proposed control scheme

- a: Total voltage loop
- b: Second cell voltage loop
- c: Current loop

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The voltage controllers were designed considering the frequency of the MPPT algorithm.

3 Loss calculation and thermal model of an insulated gate bipolar transistor (IGBT) power module

3.1 Loss calculation

The thermal behaviour of the power module can be estimated through the power losses on the device. The total power losses of a power module ($P_{\text{total}}$) can be broadly divided into the switching and conduction losses of the IGBTs and diodes, and the total losses are given by their sum

$$P_{\text{total}} = (P_{Sc} + P_{Sd}) + (P_{Dc} + P_{Ds})$$  \hspace{1cm} (5)

where $P_{Sc}$ and $P_{Dc}$ are the instantaneous conduction losses of the IGBT and diode, respectively, $P_{Sd}$ and $P_{Ds}$ are the switching losses of the IGBT and diode, respectively.

The instantaneous conduction losses of the IGBT can be calculated by the following equation:

$$P_{Sc} = v_{ces} \times i_{c}(t) + r_{c} \times i_{c}^{2}(t)$$  \hspace{1cm} (6)

where $v_{ces}$ is the on-state zero-current controller emitter voltage, $i_{c}$ is the collector current, and $r_{c}$ is the collector-emitter on-state resistance, same like IGBT, the instantaneous conduction losses of the diode can be given by

$$P_{Dc} = v_{dss} \times i_{d}(t) + r_{d} \times i_{d}^{2}(t)$$  \hspace{1cm} (7)

where $v_{dss}$ is the on-state zero-current voltage across the diode, $i_{d}$ is the current through the diode, and $r_{d}$ is the diode on-state resistance. The values of $r_{c}$ and $r_{d}$ can be calculated from the information provided in the power module datasheet.

The sum of the switch on energy without taking the reverse recovery process into account ($E_{TW-on}$) and the switch on energy caused by the reverse recovery of the free-wheeling diode ($E_{Tr-on}$) can give the turn-on switching losses in IGBT ($E_{T-on}$):

$$E_{T-on} = E_{TW-on} + E_{Tr-on}$$  \hspace{1cm} (8)

In similar manner, the turn-off energy losses also can be calculated. The total switching losses of the IGBT ($P_{Sc}$) can be given by the sum of the turn-on switching energy loss ($E_{T-on}$) and turn-off switching energy loss ($E_{T-off}$):

$$P_{Sc} = E_{T-on} + E_{T-off}$$  \hspace{1cm} (9)

The turn-on energy losses of the diode ($E_{Tr-on}$) can be calculated from the reverse recovery energy losses during a small period. The switch-off losses in the diode are normally neglected and assumed to be zero. The total switching losses of the diode ($P_{Ds}$) can be given by

$$P_{Ds} = E_{D-on} \times f_{sw}$$  \hspace{1cm} (10)

where $f_{sw}$ is the switching frequency.

3.2 Thermal model

There are two types of commercialised power modules available in the market: (i) power module with base plate and (ii) power module without the base plate. In this paper, the power module consisting of a base plate is considered. The thermal model for the single power module is indicated in Fig. 3.

The junction temperature of the power module can be estimated using the loss information. Once the power losses are determined, they are conducted through their junction to case thermal impedance ($Z_{thj-ci}$) for both IGBT and diode.

From the manufactures datasheet, each of the thermal parameters can be found. The thermal impedance of the heat sink is not provided in the datasheet. This parameter can be selected in accordance with the desired dissipated heat. The foster network has been used here to model the thermal impedance from the junction to case. The instantaneous junction temperature of the power module can be expressed as

$$T_{j} = T_{c} + Z_{thj-ci} \times P_{\text{total}}$$  \hspace{1cm} (11)

where $T_{j}$ is the case temperature, which can be defined as

$$T_{c} = T_{a} + \left( Z_{thc-bh} + Z_{thh-ah} \right) \times P_{\text{total}}$$  \hspace{1cm} (12)

where $T_{a}$ is the ambient temperature, $Z_{thc-bh}$ and $Z_{thh-ah}$ gives the thermal impedance from case to heat sink and heat sink to ambient, respectively.

4 Methodology for reliability evaluation

There are two extensively used approaches for reliability evaluation: ‘part stress’ and ‘parts count’ methods [13]. Parts count method is used when there is no detailed system information. This method is used to predict the failure rate under reference conditions. Therefore, this method can be considered as an approximate method. Unlike the part count method, parts stress method requires various stresses on each part and accurate environment conditions are needed. A number of various parameters in this method lead to increase in the accuracy. Similar to parts count method, the total failure rate in the parts stress method can be calculated by summing all failure rates. In this paper, parts stress method is applied to have more reliable results based on the measurements.

The reliability $R$ can be calculated by the following equation:

$$R(t) = e^{-\lambda t}$$  \hspace{1cm} (13)

where $\lambda$ is the failure rate and $t$ is the time that should elapse until the first failure occurs, and it is defined as by equation

$$t = \frac{1}{\lambda}$$  \hspace{1cm} (14)
The mean time between the failure (MTBF) of a system is expressed by the following equation:

\[
MTBF = MTTF + MTTR = \frac{1}{\lambda}
\]  

(15)

where MTTF is the mean time to failure and MTTR is the mean time to repair.

The total rate of the system failure (\(\lambda_{\text{system}}\)) is expressed by the equation

\[
\lambda_{\text{system}} = \sum_{i=1}^{N} (\lambda_{\text{part}})_i
\]

(16)

The predicted failure rate of the switch, diode, capacitor, and inductor can be calculated using (17)–(20), respectively [14]

\[
\lambda_{\text{p, switch}} = \lambda_{b, \text{switch}} \times \pi_\text{Q} \times \pi_\text{A} \times \pi_\text{E} \times \pi_\text{T}
\]

(17)

\[
\lambda_{\text{p, diode}} = \lambda_{b, \text{diode}} \times \pi_\text{Q} \times \pi_\text{E} \times \pi_\text{C} \times \pi_\text{S} \times \pi_\text{T}
\]

(18)

\[
\lambda_{\text{p, capacitor}} = \lambda_{b, \text{capacitor}} \times \pi_\text{CV} \times \pi_\text{Q} \times \pi_\text{E}
\]

(19)

\[
\lambda_{\text{p, inductor}} = \lambda_{b, \text{inductor}} \times \pi_\text{C} \times \pi_\text{Q} \times \pi_\text{E}
\]

(20)

where \(\lambda_b\) is the base failure rate, which is 0.012 and 0.064 for switch and diode, respectively. The base failure rate of the inductor can be calculated using (17)

\[
\lambda_{b, \text{inductor}} = 0.00035 \times \exp\left(\frac{T_{\text{HS}} + 273}{329}\right)^{15.6}
\]

(21)

where \(T_{\text{HS}}\) is the temperature of the inductor hot spot, which can be calculated using (22)

\[
T_{\text{HS}} = T_\text{A} + 1.1 \times \Delta T
\]

(22)

where \(T_\text{A}\) is the device ambient operating temperature (in degree Celsius) and \(\Delta T\) is the average temperature rise above ambient. The following equation can be used to find the \(\lambda_b\) of the capacitor

\[
\lambda_{b, \text{capacitor}} = 0.00254 \times \left(\frac{S}{0.5}\right)^3 + 1 \times \exp\left[5.09 \times \left(\frac{T_\text{A} + 273}{378}\right)^5\right]
\]

(23)

where \(S\) is the ratio of operating voltage to the rated voltage. In \(\lambda_{\text{p, switch}}\) and \(\lambda_{\text{p, diode}}, \pi_T\) is the temperature factor and can be calculated as is shown in below equation:

\[
\pi_{\text{T, switch}} = \exp\left(-1925 \times \frac{1}{T_j + 273} - \frac{1}{298}\right)
\]

(24)

\[
\pi_{\text{T, diode}} = \exp\left(-1925 \times \frac{1}{T_j + 273} - \frac{1}{293}\right)
\]

(25)

where \(T_j\) is the junction temperature, and can be calculated using (11).

In (18), \(\pi_S\) is the stress factor and which can be obtained by

\[
\pi_S = V_s^{2.43}
\]

(26)

where \(V_s\) is the ratio of operating voltage to the nominal voltage.

From (19), \(\pi_{\text{CV}}\) is the capacitor factor and is calculated as follows:

\[
\pi_{\text{CV}} = 0.34 \times C^{0.12}
\]

(27)

where \(C\) is the capacitance value (in microfarads).

The environmental factor (\(\pi_{\text{E}}\)) and quality factor (\(\pi_{\text{Q}}\)) of different elements are presented in [15]. The application factor (\(\pi_A\)) and constant construction factor (\(\pi_C\)) has been considered to be equal to 10 and 1, respectively [15].

The reliability assessment mainly depends on the temperature factor, and it is related to the conduction and switching losses of semiconductor components. In this paper, the powerful simulation software PLECS is used to determine the accurate power losses and junction temperature of the devices present in the MLI.

5 Results and discussion

A 600 V/30 A IGBT (IGW30N60H3) module is selected for inverter power module. The electrical parameters of the system are shown in Table 1. The control method has been done in MATLAB/Simulink, and the power loss model is done by PLECS blockset in Simulink environment based on look-up tables provided by the manufacturers in the datasheet. The thermal model is implemented based on a thermal Foster model, which parameters are provided also in the datasheets.

For better analysis, the proposed control scheme that is the single-phase grid-connected MLPVI is simulated using two different PWM techniques. The first one is PSPWM and another one is LSPWM technique. In the inverter, each H-bridge has its own PV panel, and all PV panels are operated under the same irradiance of 1000 W/m² and temperature of 25°C. The output voltage of the single-phase MLI is shown in Fig. 4.

From (17)–(27), the \(\pi\) factors and failure rate of the IGBT module and capacitor for single-phase-cascaded H-bridge MLPVI. Here, the thermal performances for both IGBTs and diodes are done together because they have a complementary cycle of operation. By using PSPWM technique, the junction temperature of the IGBT power device is lower compared with using LSPWM technique, and the simulation results are shown in Fig. 5.

The results show that the MTBF in Case-I (Table 2) is higher than in Case-II (Table 3), because the failure rate of the device mainly depends on the temperature factor, and the temperature depends on the losses produced by the device. In PSPWM

**Table 1 System parameters**

| Parameter               | Value      |
|-------------------------|------------|
| dc-link voltage         | 121        |
| dc-link capacitor       | 4700 µF    |
| connection Inductor     | 1 mH       |
| grid resistor           | 0.1 Ω      |

![Fig 4 Single-phase five-level inverter output voltage waveform](image-url)
6 Conclusion

In this paper, an independent dc-link voltage controller is applied to grid-connected single-phase five-level-cascaded H-bridge MLPVI. The main advantage is it has the capacity to operate at a lower switching frequency than a two-level converter. To evaluate the reliability, the loss and thermal model of different power devices and also the total system have been done. The reliability has been calculated using the co-simulation platform of PLECS blockset in Simulink and parts stress method. For the better evaluation of thermal and reliability analyses, the controller is operated under PSPWM and LSPWM techniques. It shows the reliability of the MLPVI with PSPWM technique higher than the converter with LSPWM technique.

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8 References

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