Enhanced Differential Power Processor for PV Systems: Resonant Switched-Capacitor Gyrator Converter with Local MPPT

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Abstract—This paper introduces an enhanced differential power processor topology and principle of operation for photovoltaic systems (PV) that is based on switched-capacitor technology, featuring local maximum power point tracking (MPPT) capability, zero current switching (ZCS), high efficiency over wide operation range, and reduced size. The new converter operates as a voltage-dependent current-source and is regulated by dead-time or frequency control. Local MPPT on the individual PV elements is realized and the operation is demonstrated by simulation and experiments. Differential power processing operation is verified on 100W prototypes, demonstrating ultimate improvement in the power harvesting capability of above 90% and up to 99% out of the available power in the string, under different insolation levels.

I. INTRODUCTION

Full or partial shading of a serially connected PV string severely impacts the power that can be extracted from it [1], [2]. Generally, two groups of power processing solutions have been proposed to remedy the shading problem. One group is based on a dedicated converter/inverter per element [3], [4], whereas the other group of solutions keeps the series connection of the elements intact, and process mismatched currents due to the shaded unit(s) using parallel circuitry [5]-[7]. The latter has the advantage of processing only the power differences between the PV elements, thus minimizes conversion losses and improves reliability. This architecture, referred as differential power processing [8], is the subject of this study.

A unique advantage of the differential power processing architecture is that Maximum Power Point Tracking (MPPT) can be obtained locally, in the PV element level, by processing only the necessary amount of power needed to achieve Maximum Power Point (MPP). Several converter realizations have been proposed as candidates as Differential Power Processors (DPP), mainly derived from battery management applications [9], [10]. A Switched Capacitor Converter (SCC) proposed in [11], [12], as a voltage equalizer with simple open-loop control, relying on the assumption that MPP voltage deviation due to change in irradiance level is negligibly small. This approach stands out in its simplicity, high self-efficiency and lower cost. However, it lacks MPPT capability without introducing losses. A buck-boost topology has also been proposed, acting as an equalizer [7], and further developed in [8] to obtain local MPPT by differential processing to keep all PV units in MPP. However, compared to SCC technology of the same power level, it is bulkier in volume due to the large magnetic element required.

This paper introduces an enhanced DPP topology that incorporates the virtues of both worlds in a newly developed current sourcing converter that is based on resonant SCC technology (Fig. 1). The new converter is of low-volume, features high-efficiency with extended operation range, and is capable to perform local MPPT. This paper covers a brief introduction to DPP architecture, and presents a new converter topology with its principle of operation. Later, differential power processing and local MPPT are described by simulations and experiments.

![DPP Conceptual hardware setup.](image-url)
II. DIFFERENTIAL POWER PROCESSING – OPERATION AND REALIZATION

The main goal of differential power processing in PV systems is to maximize the power conversion efficiency by processing only a small portion of the power being produced. The concept has been initially introduced in [13], [14] and further developed and demonstrated in [8]. The architecture is conceptually shown in Fig. 2. It consists of \( N \) serially connected PV elements and \( N-1 \) current sourcing converters connected in parallel with two adjacent PV elements. The differential power passes along the string in a so-called bucket-brigade pattern. Each converter performs local MPPT to one of its two connected PVs by sinking or sourcing current to/from the neighboring PV element. A central grid-connected inverter is still used to interact with the grid and to track the global MPPT. Effectively, it also performs the local MPPT for the \( N^{th} \) element.

The differential converters are needed only in case of mismatch in the MPPs, i.e. only a portion of the power is processed. The MPP is maintained for each of the PV elements thanks to differential current provided by the DPP. The amount of power processed by each converter in a string of \( N \) elements to bring the mismatched element (PV\( x \), Fig. 2) to its MPP can be expressed as:

\[
P_{\text{DPP}} = \begin{cases} 
\frac{P_x - P_{x-1}}{N}, & j < x \\
\frac{P_x - P_{x+1}}{N}, & j \geq x 
\end{cases},
\]

where the index \( j \) represents the location of a converter in the string, \( P_x \) is the maximum power of the shaded element and \( P_{x-1} \) is the power of a non-shaded element at its MPP. It can be observed from (1) that adjacent converters to the shaded element are required to process most of the power whereas others located farther in the chain contribute a smaller portion of the power, linearly proportional to their location with respect to the mismatched PV element. A current-sourcing converter that operates as a DPP is thus required to be bidirectional and capable of step-up and step-down operation. In the context of this study, the harvesting factor, \( \xi \), is defined as the relationship between the total extracted power out of a serially connected chain, \( P_{\text{out}} \), and the sum of the absolute maximum power that can be harvested from each individual element, \( P_{i,MPP} \), that is:

\[
\xi = \frac{P_{\text{out}}}{\sum_{i=1}^{N} P_{i,MPP}}.
\]

Considering (1) and (2), it would be highly advantageous that the efficiency of the converter will be independent of the operating conditions. Converter topologies such as switched-capacitor [10] or buck-boost [9], originally proposed for battery equalization and realized for PV applications have many merits, however, their efficiency range is limited. SCC lacks MPPT capability [11]. In a buck-boost configuration, the efficiency range is somewhat limited around the nominal power level, and there is a trade-off between the size and performance of the converter.

To overcome these challenges, this study introduces a new DPP topology. It combines the benefits of reduced size SCC and current sourcing properties with high efficiency over a wide range [15].

III. RESONANT SWITCHED-CAPACITOR GYRATOR CONVERTER

The converter configuration as a DPP is described in Fig. 1. Similar to the architecture of a conventional inverting resonant SCC [16]-[19], the DPP includes four switches and a resonant tank. Two PV elements connect as input and output sources.

As opposed to the operation of a classical SCC that includes a charge state and a discharge state, here, an additional switching phase is introduced that breaks the rigid connection of the input/output voltage gain and the efficiency of the converter. Controlling the sequence of the switches governs the power flow direction, hence bidirectional step up/down operation.

![Fig. 2. Example of the power flow for a \( N \) PV panel string connected to \( N-1 \) gyrator DPPs, containing a shaded PV at the \( x \) location.](image-url)
The operation of the Resonant Switched-Capacitor Gyrator Converter (RSCGC), shown in Fig. 1, is described for one steady-state charge/discharge/balance cycle and is assisted by Fig. 3 that illustrates the capacitor voltage, \( V_C \), and the resonant tank current, \( I_C \), for a case of a MPP mismatch that requires a non-unity step-up conversion. By turning \( Q_1, Q_2 \) on, a charge state (S1) is commenced, which resonantly charges the flying capacitor from \( PV_1 \). At zero current, \( Q_1 \), \( Q_3 \) are turned on and \( Q_2 \), \( Q_4 \) are turned on (state S2). At this point, the flying capacitor resonantly discharges onto \( PV_2 \). If \( PV_1 \) voltage, \( V_{PV1} \), and \( PV_2 \) voltage, \( V_{PV2} \), are of different values, only a portion of the charge is delivered to the output and results in \( V_C \) that is different than its voltage at the starting point of S1. The amount of this voltage difference (neglecting \( R_s \) - parasitic resistances in each loop) equals to twice the residual voltage of the flying capacitor. By turning \( Q_2, Q_3 \) on (S3), the resonant tank is short-circuited. This creates the required charge-balance and reverses the flying capacitor voltage polarity such that the voltage at the end of S3 equals to the voltage at the beginning of S1.

The addition of a third, charge balancing state, to the switching sequence transforms the resonant SCC into a voltage dependent current-sourcing converter that, neglecting losses, is capable of accommodating any input to output voltage gain (larger and smaller than unity). It should be noted that the order of the switching sequence will govern the power flow direction.

To facilitate regulation of the amount of charge that is transferred to the output, a Pulse Density Modulation (PDM) is employed. A time-delay is introduced between the charge and the discharge states. The average currents \( I_{D}, I_{D}^{'} \) and voltages \( V_{PV1}, V_{PV2} \) of the converter can be defined as a gyrator relationship [15], [20], [21]:

\[
I_{D} = 2/jC \, V_{PV2} \quad ; \quad V_{PV1} = \frac{1}{2/jC} \, I_{D}^{'} 
\]

where \( f \) is the frequency of a cycle that includes the three states and the time-delay. Maximum differential current is passed when no additional time-delay is added. It can be observed from (3), that the differential current is inversely proportional to the time-delay, but linearly proportional with \( f \). This implies that controlling the differential current by the frequency as the correction signal is preferred, since variations in the time-delay would result in a nonlinear behavior. Furthermore, the time resolution of a variable frequency signal, generated by a local oscillator, is not constant and strongly depends on the operating point [22] which introduces additional nonlinearity. By modifying the correction signal to frequency, incremented by fixed steps, \( df \) these nonlinearities are eliminated, and the power converter can be treated as a constant gain block from frequency to current as prescribed in (3).

The relationship between the maximum current, \( I_{D\text{max}} \), passed from \( PV_1 \), and the voltage on an adjacent element, \( PV_2 \), is a function of the capacitive and inductive components of the resonant tank \( (L, C) \), and can be expressed as:

\[
I_{D\text{max}} = \frac{2}{3\pi Z} \, V_{PV2} \quad ; \quad Z = \sqrt{L/C}.
\]  

Assuming that identical parasitic resistances \( R_s \) exist in all three sub-circuits (Fig. 1), the expected efficiency of the converter can be estimated by:

\[
\eta = \left[1 + \frac{\pi R_s}{2Z} (A + A^{-1} - 1)^{-1}\right]^{-1} \quad ; \quad A = \frac{V_{PV2}}{V_{PV1}}.
\]  

As can be seen from (5), maximum efficiency is obtained at unity gain \( (A = 1) \), and it is a function of the ratio between resistance \( R_s \) to the resonant network characteristics. Ideally, assuming negligibly small parasitic resistances, the efficiency of the converter would be 100% for any finite conversion ratio. A unique feature associated hitherto only with switched-inductor converters, is now made available to resonant SCC as well.

IV. POWER FLOW CONTROL AND LOCAL MPPT

In case of MPP mismatch in more than one element in the string, the DPPs are required not only to operate at conversion ratios higher or smaller than unity to transfer the required current difference, but also to control the direction of the current toward the ‘weaker’ PV element from both sides of the string. The introduced RSCGC, shown in Fig. 1, operates as a current-source and thus is capable of stepping the voltage up or down as required by the load. To deliver power from \( PV_1 \) to \( PV_2 \) the sequence will be (S1, S2, S3). That is, charge from \( PV_1 \), discharge on \( PV_2 \), and reverse the flying capacitor polarity. In the case of power to be delivered from \( PV_2 \) to \( PV_1 \) the sequence will be reversed to (S2, S1, S3).

A. Zero Current Switching (ZCS)

To fully utilize the benefits of the converter, ZCS for the entire operation range is essential. Since the resonant characteristics may vary significantly with the operating conditions, the simplistic approach of a comparator-based zero current detector is not suitable here. Other direct ZCS methods such as [23] or [24] have demonstrated wide operation range, however, they rely on high performance processors with fast execution rate for proper operation.

![Fig. 3. Typical waveforms (obtained from simulation) of the flying capacitor voltage and current. Circuit parameters are: \( V_{PV1} = 20V \), \( V_{PV2} = 31V \), \( R_s = 0.15Ω \), \( L = 5.2μH \), \( C = 0.25μF \).](image)
To overcome the variations in the operating conditions and rely on limited processing capabilities, an adaptive ZCS control scheme has been developed, based on an accumulative calibration of the switching time. The control scheme applies an analysis of the rectified L-C tank current on areas in vicinity to the transition between switching states (estimated location of the zero crossing point). Instead of a continuous sampling that is required by other techniques, here the current is discretely sampled, forming a vector of the current shape in the area of the commutation. This is allowed assuming that the voltages at the DPP terminals are virtually constant during a calibration cycle which is much shorter than an MPPT iteration time.

For a calibrated ZCS transition, the rectified current signal should have a ‘V’ shape with a minimum at zero, centered at the exact switching time. Non-calibrated switching action results in one out of four possible scenarios, as can be observed in Fig. 4 and summarized in TABLE I, indicating whether the transition has occurred early or late with respect to the actual zero point. Based on this information, and on the distance between the minimum and transition time, the switching time may be increased or decreased per switching state.

**TABLE I. Criteria for distinguishing between early or late switching**

<table>
<thead>
<tr>
<th>Reference in Fig. 4</th>
<th>Shape</th>
<th>Minimum location</th>
<th>Minimum value</th>
<th>Conclusion</th>
</tr>
</thead>
<tbody>
<tr>
<td>(a)</td>
<td>‘V’</td>
<td>After</td>
<td>Zero</td>
<td>Early</td>
</tr>
<tr>
<td>(b)</td>
<td>‘V’</td>
<td>On time</td>
<td>Positive</td>
<td>Early</td>
</tr>
<tr>
<td>(c)</td>
<td>‘V’</td>
<td>Before</td>
<td>Zero</td>
<td>Late</td>
</tr>
<tr>
<td>(d)</td>
<td>‘W’</td>
<td>After</td>
<td>Zero</td>
<td>Late</td>
</tr>
</tbody>
</table>

*In relation to the transition time

**B. MPPT implementation**

Among the wide diversity of possible MPPT algorithms that can be realized using local information [25], the well-known Hill-Climbing (HC) scheme is chosen [26]. HC MPPT features a simple realization based on readily available information; it does not require continuous perturbations and its convergence rate to the MPP can be enhanced [26]. As described earlier, each DPP performs local HC MPPT on one PV element in the chain. In this study, it is set to be carried out on the lower element. A conceptual hardware setup is depicted in Fig. 1, showing the sensing points for the current and voltage of the PV element. The controller calculates \( \frac{dp_{pv}}{dv_{pv}} \) \((v_{pv} \text{ is the sensed PV voltage, } p_{pv}=v_{pv}i_{pv} \text{ is the PV power, and } i_{pv} \text{ is the sensed PV current})\) as the error signal and generates frequency command \( f \) such that the differential current, \( I_{PV} \), transferred by the DPP into or out of the controlled PV element. The procedure is repeated until \( \frac{dp_{pv}}{dv_{pv}} \) has stabilized in the vicinity of zero, as prescribed by the HC MPPT algorithm. The correction signal \( \frac{dp_{pv}}{dv_{pv}} \) can now be expressed as:

\[
\frac{dp_{pv}}{dv_{pv}} = i_{pv} + \frac{di_{pv}}{dv_{pv}}v_{pv}.
\]  

(6)

It can be observed from (6) that different insolation levels (which directly affect \( i_{pv} \)) result in different \( \frac{dp_{pv}}{dv_{pv}} \) curves. To facilitate uniform control, the correction signal can be adjusted to the current insolation level, i.e. by normalizing it to \( i_{pv} \). The normalized signal, \( dp_n \), can now be expressed as:

\[
dp_n = \frac{dp_{pv}}{i_{pv}} = 1 + \frac{di_{pv}}{dv_{pv}}v_{pv}.
\]  

(7)

Fig. 5 shows typical curves of (7) for a several insolation levels. It can be observed that when \( v_{pv} < V_{MPP} \), \( dp_n \) holds similar values and is bounded to 1. Over the MPP peak, when \( v_{pv} > V_{MPP} \), \( dp_n \) is negative and is not bounded. To assure convergence of the MPPT algorithm, this value can be limited to (-1), as well. The modified HC MPPT is thus designed to generate a variable step in \( f \), proportional to \( dp_n \) according to:

\[
f [n] = f [n-1] + \Delta f_{max} dp_n ,
\]  

(8)

where \( f [n] \) is the current frequency command, \( f [n-1] \) is the frequency generated in the previous MPPT iteration and \( \Delta f_{max} \) sets the maximum frequency step that is allowed. The maximum frequency of \( f \) is bounded by the RSCGC natural frequency, \((3\pi\sqrt{LC})^{-1}\).
The MPPT algorithm convergence time and pattern is primarily dependent on the step size $\Delta f_{\text{max}}$, which can be viewed as the integral gain factor of (8). For small frequency steps the convergence would be smooth but slow, whereas large frequency steps would result in rapid convergence but may introduce an oscillating error around the MPP. In this work, $\Delta f_{\text{max}}$ has been manually selected to produce the fastest convergence while still maintaining a first order pattern (i.e., without overshoot).

**C. Resolution effect on Limit cycle oscillations**

Convergence of $\Delta p_n$ to zero is impractical due to the discrete nature of $f$ (generated by the digital hardware), and as a consequence $\Delta p_n$ has discrete steps. The resolutions of $\Delta p_n$ and $f$ have to be carefully selected to assure stability around MPP, i.e. to avoid limit cycle oscillations [27]. Since $\Delta p_n$ is not sensed directly, a practical way to eliminate such oscillations is to decrease the resolution of the calculated $\Delta p_n$ around zero, i.e., by introducing a Zero Error Bin (ZEB) of size $\Delta$. The ZEB is defined such that an absolute value of the calculated $\Delta p_n$ less than $\Delta$ is considered as zero error, indicating the system has converge into the MPP. The size of $\Delta$ should be set such that the ZEB contains at least one value of $\Delta p_n$.

The smallest frequency step (1 bit resolution), $\Delta f_{\text{DCO}}$ that can be generated out of the DCO that generates the switching signal is a function of the time base of the local oscillator, $TB$, and the running frequency, $f_{\text{DCO}}$. According to [27], $\Delta f_{\text{DCO}}$ can be expressed as:

$$\Delta f_{\text{DCO}} = \frac{1}{N_{\text{TB}}} - \frac{1}{(N_{\text{TB}}-1)} TB = TB f_{\text{DCO}}^2 , \quad (9)$$

where $N_{\text{TB}}$ is the number of TBs in one period.

Around the MPP, the no limit cycles criterion can be expressed as:

$$A_{\text{qf}} \times G_{\text{f}} \left( V_{\text{pv}} \right) \times \Delta f_{\text{DCO}} < \Delta ; \quad \begin{cases} A_{\text{qf}} = \frac{d dp_n}{di} \\ G_{\text{f}} = \frac{di}{df} \end{cases} , \quad (10)$$

where $A_{\text{qf}}$ is the PV element small-signal gain between the normalized gradient ($dp_n$) and the differential current ($i_\text{pv}$) which can be obtained by taking the derivative of (7) with respect to $i_\text{pv}$ and $G_{\text{f}} (V_{\text{pv}})$ is the RSCGC current-to-frequency small signal gain which is obtained by derivation of $d$ with respect to $f$. The block diagram of Fig. 6 depicts the control loop of a single RSCGC DPP and the effect of $\Delta f_{\text{DCO}}$ step on $dp_n$.

The derivatives of $A_{\text{qf}}$ and $G_{\text{f}}$ around the maximum power point, which is of interest for limit cycles calculation, can be expressed as a function of the system parameters as:

$$A_{\text{qf}} \bigg|_{\text{MPP}} = \frac{V_{\text{1,MPP}}}{I_{\text{1,MPP}}} ; \quad G_{\text{f}} \bigg|_{\text{MPP}} = 2CV_{2,MPP} \cdot \quad (11)$$

where $V_{1,MPP}$ and $I_{1,MPP}$ are the controlled PV panel’s MPP voltage and current, respectively, and $V_{2,MPP}$ is the adjacent PV panel’s MPP voltage. It should be noted that the derivative of $A_{\text{qf}}$ has been obtained using a numerical calculation.

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**Fig. 6.** Block diagram of the DPP digital control system.

**Fig. 7.** Flowchart of the modified HC MPPT algorithm.
Having all the relationships of (10), and after some manipulations, the no limit cycles criterion can be expressed as:

$$\frac{V_{LMP}}{I_{LMP}} \times 2CV_{2MPP} \times TBf_{DCO}^2 < \Delta \quad (12)$$

The result of (12) forms the selection trade-off between the minimum frequency step that is generated and the deviation $$\Delta$$ from the true MPP. It implies that in a similar way as in other switch-mode applications, the existence of limit cycle oscillations depends on the command signal generator as well as on the local gain of the converter. However, as opposed to conventional applications where the source has no effect on limit cycles, here the characteristics of the PV element also contribute a gain factor to the loop. The tool formed in (12), given the system parameters, may assist in the selection of $$\Delta$$ and $$\Delta_{max}$$ in (8) that will allow rapid convergence to the MPP as well as a tolerable steady-state error.

D. Verification by Simulation

To demonstrate the operation of the new DPP and the advanced HC MPPT algorithm, a PSIM (PowerSim Inc., ver. 9.0) simulation test bench has been constructed as illustrated in Fig. 2, with $$N=3$$. The HC MPPT algorithm, depicted in Fig. 7, has been implemented using a C block element in the PSIM environment. A load resistance value $$R_{Load}$$ mimics the operation of the central inverter, provides the global MPP according to:

$$R_{Load} = \frac{\left(\sum V_{i,MPP}\right)^2}{\sum P_i,MPP},$$

where $$V_{LMP}$$ and $$P_{LMP}$$ are the PV elements’ corresponding MPP voltage and power. The resonant power stage elements’ values were $$C=1\mu F$$, $$L=1\mu H$$ and $$R=16m\Omega$$ The ZEB size was set at $$\Delta=0.05$$. The results shown in Fig. 8 reveals an MPP DC error of 0.55W (less than 0.5% of its $$P_{i,MPP}$$) without limit cycle oscillations and convergence within 23 iterations.

V. EXPERIMENTAL SYSTEM AND RESULTS

The operation of the differential power architecture was verified experimentally on a chain of three 180W SHARP PV panels (NU-180, E1) with two 100W RSCGC-based DPP prototypes, realized using an inverting bridge configuration as shown in Fig. 1. PMOS transistors were used for $$Q_1$$ and $$Q_2$$, and NMOS transistors for $$Q_3$$ and $$Q_4$$. The resonant tank parameters were $$C=1\mu F$$ ($10\times0.1\mu F$ ceramic), $$L=0.5\mu H$$. The bus capacitor was $$C_{BP}=50\mu F$$ ($5\times10\mu F$ ceramic). Adaptive ZCS and HC MPPT algorithms were implemented digitally on a dsPIC33FJ16GS502. Dead time between the switches was set to a constant 100ns. To eliminate limit cycle oscillations, an error window $$\Delta=0.14$$ around the MPP has been applied. This value is selected based on the criterion formed in (12) with some error margin due to the tolerance of the practical parameters.

Fig. 9 shows the inductor’s current and the gating signals for one DPP transferring differential current of 1.25A, for a case of an MPP difference of 38% between two adjacent panels. It also shows the added time delay that controls the average current and the effectiveness of the adaptive ZCS calibration. Fig. 10 demonstrates the convergence of the lower panel towards MPP, within a three PV element string containing two DPPs. The transient was created by switching on its corresponding DPP while the second DPP was connected and running, keeping its panel in MPP. TABLE II. summarizes the results of several non-uniform shading conditions and the harvesting factor $$\xi$$ for each experiment. The summary compares the operation with the DPPs to the operation of a chain with central inverter. Fig. 11 shows a picture of the testing environment.

### TABLE II. HARVEST IMPROVEMENT DUE TO ADDITION OF THE DPPS

<table>
<thead>
<tr>
<th>$$P_{LMP}$$</th>
<th>$$P_{LMP}$$</th>
<th>$$P_{P_{max}}$$</th>
<th>$$\xi_{inverter}$$</th>
<th>$$\xi_{DPP}$$</th>
<th>Improv.</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.5</td>
<td>0.95</td>
<td>1</td>
<td>340</td>
<td>0.78</td>
<td>0.94</td>
</tr>
<tr>
<td>0.4</td>
<td>0.95</td>
<td>1</td>
<td>327</td>
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<td>0.9</td>
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<tr>
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<td>0.74</td>
<td>1</td>
<td>237</td>
<td>0.68</td>
<td>0.96</td>
</tr>
<tr>
<td>0.86</td>
<td>1</td>
<td>233</td>
<td>0.94</td>
<td>0.99</td>
<td>5%</td>
</tr>
<tr>
<td>0.65</td>
<td>1</td>
<td>200</td>
<td>0.82</td>
<td>0.97</td>
<td>18%</td>
</tr>
<tr>
<td>0.54</td>
<td>1</td>
<td>203</td>
<td>0.81</td>
<td>0.95</td>
<td>17%</td>
</tr>
<tr>
<td>0.38</td>
<td>1</td>
<td>193</td>
<td>0.68</td>
<td>0.91</td>
<td>33%</td>
</tr>
</tbody>
</table>

a. Normalized to the power of the strongest panel

b. $$\xi$$ with central inverter only

c. $$\xi$$ with DPP and a central inverter
VI. CONCLUSIONS

An enhanced differential power processing converter for PV systems was presented. The new power converter combines virtues of both switched-capacitor and switched-inductor technologies, and features low-volume and high conversion efficiency over wide range. Although based on SCC architecture, the power module is fully capable of performing local MPPT such that only the mismatch power between PV elements is processed while converging each of the elements into their MPP.

A digital differential control scheme was developed and validated. The algorithm design and analysis takes into account the discrete nature of digital control and eliminates limit cycle oscillations.

A theoretical analysis on the existence of limit cycle oscillations has been carried out. It revealed that deviation from the MPP can be the result of steady-state oscillations around the maximum point due to the finite resolution of the digital hardware, and therefore less energy is extracted. It was also found that the contributors to the oscillations are the frequency resolution of the DCO and the DPP gain around the MPP, and more importantly that the source, the PV element, also has a dominant factor in the appearance of such oscillations. This is in contrary to other, conventional, switch-mode applications where the source is assumed as constant.

The experimental results demonstrate significant improvement of power harvesting capability up to 33% (translates into 43W) for a mini string of only two PV panels, and up to 42% for a three PV string (translates into 65W). Sub-panel utilization would increase power harvest, covering more shading scenarios [28], [29]. The power processor introduced in this study, is considered an extremely attractive candidate for sub-panel and even cell-level implementation due to its low-volume requirements.

REFERENCES


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