Efficiency Optimization of Biological Photovoltaic Cells using Nano-Powered Electronic Circuits Technology

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Abstract: This study presents a new method for optimizing the power conversion efficiency of the Biological Photovoltaic cells (BPVs). In this method, the electronic circuits that operate within ultra-low power range are used rather than propose new strategies for facilitating the electrons transfer to the anode inside the cell through design new electrodes coupled with the use of different species of photoautotrophic organisms. The proposed method includes building a Nano-Powered Transconductance Amplifier (NPTA) using the Electrically Programmable Analog Devices (EPAD). The NPTA is designed with a special topology, through which the output voltage is almost equal to the open-circuit voltage of the BPV cell and the output current is boosted by the large transconductance of the NPTA. The new approach exhibits a large increase in the power conversion of the BPVs which can be utilized to power the electronic devices.

Key words: Design, organisms, NPTA, facilitating, electrodes, voltage

INTRODUCTION

Comparing to the Silicon-based Photovoltaic cells (SPVs), the design of the BPVs that based on the photosynthetic reaction is attracting substantial interest as a means for conversion of solar light energy into electrical power. This interest is due to the ease of design, low manufacturing cost and energy generation potential during the night as a result of being the electrons are stored inside the photoautotrophic organisms during daylight hours. The BPVs are characterized by a short-circuit current density with a low rate of the microamps and open-circuit voltage with an average value of 0.25 V (Kazemzadeh et al., 2017), making them less efficient in converting solar energy into electrical energy. Therefore, these cells suffer from a scant efficiency in the electrical power conversion comparing to the SPVs. Hence, the objective of this work is to produce commercial devices with low manufacturing cost and excellent energy conversion efficiency.

BPV systems are defined by the type of the biological materials that they employ such as aquatic plants, green algae and cyanobacteria and the mechanism of electrons transfer from the biological material to the anode. In principle, the process of the electrons generation is common to all of the BPV systems where the electrons are produced by the photoautotrophic organisms through light-driven oxidation of water. Based on the BPV system that utilize sub-cellular components or whole cell, the BPVs are divided into two sub-categories, sub-cellular BPVs and cellular BPVs (Bombelli et al., 2011).

Sub-cellular BPVs are those that use the Photosystem II (PSII) which is a biochemical mechanism through which the photoautotrophic organisms absorb light energy for photosynthesis (Badura et al., 2011). In this type of BPVs, some studies have utilized the thermophilic cyanobacterium in the light harvesting and energy generation as it is quite stable in its purified form (Kato et al., 2012). Using the technique of inserting a nanoelectrode into the chloroplast of the living cell, photosynthetic electrons (1.2 pA at 6 A m⁻²) were directly extracted (mediator-free) from the living algal cell (Ryu et al., 2010), although, scale-up this technique would be difficult. Thylakoid membrane was used for photocurrent generation via. Indirect Extracellular Electron Transfer (IEET) where the composite resulting from thylakoid-carbon nanotube exhibited high photo-electrochemical activity under illumination (Calkins et al., 2013). However, the practical applications of the sub-cellular BPVs are restricted by the stability of the light harvesting components (Raven, 2011; Bombelli et al., 2011).

In the cellular BPVs, the oxygenic photosynthetic organisms are used to produce electrons through the photolysis of water and supply them to the anode without the aid of heterotrophic species. Notably, the photosynthetic organisms in the cellular BPVs can generate electronic current in the dark through the aerobic breakdown of the internal carbon reserves accumulated during the light (McCormick et al., 2011, 2015). Mostly, the use of the prokaryotic cyanobacteria (blue-green algae) has been preferred in the cellular BPVs over the eukaryotes due to its simple physiology where the
electron transfer across the membrane requires fewer steps (Schultze et al., 2009). Some studies utilized various species of the cyanobacteria including filamentous strains and unicellular strains to generate light-dependent current (Tsujimura et al., 2001). Through utilizing lipid-soluble artificial electron mediators to extract the electrons from the photoautotrophic organisms, power outputs higher than a few mW have been achieved (Yagishita et al., 1997) but their addition can cause a significant reduction the organisms viability over time (Martens and Hall, 1994). Some researchers achieved a significant positive light response through coating the anodes of the photosynthetic microbial fuel cell with electrically conductive polymers (Zou et al., 2009). Other researchers achieved a maximum power output of 6 mW m⁻² where they exhibited that a positive light response could be preserve for several weeks (Pisciotta et al., 2010). Micro fluidic-based BPV showed the highest recorded power densities of 100 mW m⁻² in the light and 80 mW m⁻² in the dark (Bombelli et al., 2015). However, on basis of the data reported for all species of the filamentous cyanobacteria (phormidium), the current produced by the photoautotrophic organisms in BPVs that use H₂O as a substrate for electrons generation was calculated to be <0.5% of the total number of the electrons generated by water oxidation (Ochiai et al., 1983). This means that despite of the improvements made on the BPVs, these cells still suffer from scant efficiency in the electrical power conversion.

MATERIALS AND METHODS

Research method: In this research, due to the low operating voltages, the electrically programmed MOSFETs that having a threshold voltage of zero volt with a precision tolerances are used in design the proposed NPTA for optimize the power conversion efficiency of the BPVs. Electrical programming technique was designed especially for the MOSFETs to largely reduce the power dissipation and current leakage, through enabling precision control of a specific parameters such as the gate threshold voltage. This technique involves trimming the MOSFET gate with a floating gate of Polysilicon embedded in the MOSFET gate oxide (Anonymous, 2005). The trimmed gate is electrically programmed by injecting a charge of electrons with a sufficient energy (hot electrons) to enter the oxide layer through the floating gate in order to create an electrically programmable analog device. Once the electrical programming is done, the required voltage and current levels are stored indefinitely in the device. This type of the MOSFET which can be electrically programmed through the trimmed gate is called the EPAD MOSFET.

The EPAD MOSFETs are characterized as ultra low-powered normally-on devices with large trans conductance, even at very low supply voltages. These devices can be electrically programmed in matched pairs or individually within a dual and quad packages to enable independent control of each device. The EPAD MOSFETs were designed and built on a single monolithic chip by the Advanced Linear Devices, Inc. manufacturer in dual and quad packages to enable design a multiple cascaded stages operate at a very low bias voltage. Thereby, by using these devices, it is possible to built an ultra low-powered electronic circuits that operate at very low supply voltage. Consequently, the zero-threshold EPAD MOSFETs packages used in this work are the dual package ALD212900 and the quad package ALD310700. The dual package ALD212900 is a matched pair of N-channel zero-threshold EPAD MOSFETs characterized by the following features and applications (Anonymous, 2017a, b):

Features:

- Zero threshold $V_{TH} = 0.00V ± 0.01V$
- Offset voltage match to 2 mV/10 mV max
- Sub-threshold voltage (nano-power) operation
- Minimum operating voltage of less than 100 mV
- Minimum operating current of less than 1nA
- Minimum operating power of less than 1nW
- DC current gain of larger than 10⁸ at 25°C
- High transconductance and output conductance
- Low ON drain-to-source resistance, $R_{DS}$(ON) of 14Ω
- Output current of higher than 50 mA
- Matched and tracked temp-coefficient
- Tight lot-to-lot parametric control (structurally similar)
- Positive, zero and negative $V_{TH}$ temp-coefficient
- Low input capacitance and leakage currents

Applications:

- Low overhead current mirrors and current sources
- Zero power normally-on circuits
- Energy harvesting circuits
- Very low voltage analog and digital circuits
- Zero power fail-safe circuits
- Backup battery circuits and power failure detector
- Extremely low level voltage-clamps
- Extremely low level zero-crossing detector
- Matched source followers and buffers
- Precision current mirrors and current sources
- Matched capacitive probes and sensor interfaces
- Charge detectors and charge integrators
- High gain differential amplifier input stage
- Matched peak-detectors and level-shifters
- Multiple channel sample-and-hold switches
- Precision current multipliers
- Discrete matched analog switches and multiplexers
- Nano power discrete voltage comparators
The quad package ALD310700 is a matched pair of P-channel zero-threshold EPAD MOSFETs characterized by the following features and applications (Anonymous, 201a, b).

**Features:**
- Precision matched gate threshold voltages
- Precision offset voltages, 10 mV max
- Sub-threshold voltage operation
- Low min operating voltage of <0.2V
- Ultra low min operating current of <1nA
- Nano-power operation
- Wide dynamic operating current ranges
- Exponential operating current ranges
- Matched transconductance and output conductance
- Matched and tracked temperature characteristics
- Tight lot-to-lot parametric control (structurally similar)
- Positive, zero and negative $V_{TH}$ temp-coefficient bias currents
- Low input capacitance
- Low input and output leakage currents

**Applications:**
- Precision current mirrors and current sources
- Low temp-coefficient ($\leq 50$ ppm$^\circ$C) current mirrors and sources
- Energy harvesting circuits
- Very low voltage analog and digital circuits
- Backup battery circuits and power failure detectors
- Precision low-level voltage clamps
- Low-level zero crossing detector

Figure 1 and 2 depicts the pin configuration of ALD212900 and ALD310700 packages.
In general, for analysing the transconductance amplifier circuit, Fig. 2 shows the basic differential amplifier-based transconductance amplifier in the conventional MOSFETs structure where the conventional MOSFET equations also apply to the EPAD MOSFETs (Chao, 2015).

In the circuit of Fig. 1, practically it was observed that even if voltage $V_1$ and $V_2$ are equal, the output is not zero which reveals that the more realistic expression of $V_1$ and $V_2$ is:

\[
V_i = \frac{V_1 + V_2}{2} + \frac{V_1 - V_2}{2} = \frac{V_1 + V_2}{2} = \frac{V_m + V_d}{2} \tag{1}
\]

\[
V_2 = \frac{V_1 + V_2}{2} - \frac{V_1 - V_2}{2} = \frac{V_1 + V_2}{2} = \frac{V_m - V_d}{2} \tag{2}
\]

where $V_{cm} = (V_1 + V_2)/2$ which is common to both $V_1$ and $V_2$ is called common-mode voltage. Hence, the practical transconductance amplifier operates in two modes, common-mode and differential-mode. To explain the operating in these two modes from Eq. 1 and 2, Fig. 3 shows the equivalent circuits to the common-mode and differential-mode.

where, $V_{C1}$ and $V_{C2}$ are the output voltages due to the common-mode $R_{SS}$ which is of very large value is the shunt resistance of the open-circuit current $I_{ref}$; $V_{st}$ and $V_{o}$ are the output voltages due to the differential-mode and $I_{SS} = I_{st} + I_{st}$.

In the practical differential amplifier, the common-mode Voltage $V_{cm}$ which is a half sum of the input voltages is undesired signal because it causes a deviation from the ideality and the desired behaviour of the differential amplifier. This deviation is due to the amplification given to the $V_{cm}$ that appears in the output. In the ideal differential amplifier, the common-mode voltage gain should be zero, so that, the output is not affected by the common-mode voltage. Based on modelling the MOSFET by its equivalent circuit to the low frequencies, Fig. 4 shows the equivalent low-frequency circuit model to the common-mode for determining the ideality condition in the performance from the circuit of Fig. 4:

\[
V_{C1} = \left( g_{m1}V_{st} + g_{m2}V_{st} \right) \eta_{ns} + \eta_{ss}I_{ss} \tag{3}
\]

Or from the circuit of Fig. 3 a, since, $V_{gs3} = V_{cm} - R_{SS}I_{SS}$:

\[
V_{C1} = \left[ g_{m3}V_{C1} \right. \left. \frac{V_{gs3}}{r_{ns}} \left( V_{cm} - R_{SS}I_{ss} \right) \right] \eta_{ns} + \eta_{ss}I_{ss} \tag{4}
\]

Solving Eq. 4 for $V_{C1}$ gives:

\[
V_{C1} = \frac{R_{SS}I_{ss} \left( 1 + g_{m3} \eta_{ns} \right) g_{m3} \eta_{ss}V_{cm}}{1 - \eta_{ns} \left( g_{m3} + \frac{1}{r_{ns}} \right)} \tag{5}
\]

or since $(1/r_{ns})\ll g_{m3}$:

\[
V_{C1} \approx \frac{R_{SS}I_{ss} \left( 1 + g_{m3} \eta_{ns} \right) g_{m3} \eta_{ss}V_{cm}}{1 - \eta_{ns} g_{m3} \frac{1}{r_{ns}}} \tag{6}
\]

From the circuit of Fig. 4 where $V_{gs3} = V_{cm} - R_{SS}I_{SS}$.


\[ g_m(V_{cm} - R_{SS}I_{SS}) + \frac{V_{cm} - R_{SS}I_{SS}}{r_{ds}} = \frac{I_{SS}}{2} \]  

Since, \( R_{SS} \) is of very large value, then from the circuit of Fig. 4, could entirely drops on \( R_{SS} \) (\( R_{SS} I_{SS} = V_{c1} \)). Therefore, Eq. 7 can be approximated as:

\[ g_m(V_{cm} - R_{SS}I_{SS}) \approx \frac{I_{SS}}{2} \]  

Solving Eq. 6 for \( I_{SS} \) gives:

\[ I_{SS} \approx g_m V_{cm} \approx \frac{V_{cm}}{1 + g_m R_{ss}} \]  

By substituting Eq. 9 in Eq. 6 and solving for \( V_{c1} \) yields:

\[ V_{c1} \approx \frac{V_{cm}(1 + g_m R_{ss}) g_m r_{ds} V_{cm}}{1 - g_m r_{ds} g_m} \approx -\frac{V_{cm}}{r_{ds} g_m} \]  

From the circuit of Fig. 4:

\[ V_{c1} = I_{c2} R_L \]  

\[ I_{c2} = g_m V_{c1} - \frac{V_{cm} - R_{SS}I_{SS}}{r_{ds}} \]  

Substituting the value of \( I_{SS} \) from Eq. 9 and the value of \( V_{c1} \) from Eq. 10 in Eq. 12 where from the circuit of Fig. 3 a \( V_{gs4} = V_{cm} - R_{SS}I_{SS} \) gives:

\[ I_{c2} \approx \left( \frac{1}{r_{ds}} g_m V_{gs4} \right) \frac{V_{cm} - V_{c1}}{r_{ds} g_m} \]  

Thus, by substituting Eq. 13 in Eq. 11 and solving for yields:

\[ V_{c2} \approx \frac{(1/r_{ds})[g_m/g_m r_{ds}]}{(1/r_{ds})[1/(r_{ds} g_m)]} V_{cm} \]  

Hence, Eq. 14 states that the condition of suppressing \( V_{c2} \) which is the undesired output voltage resulting from the common-mode voltage is that both of M1, M2 and M3, M4 should be matching. This condition is achieved through utilizing the packages ALD212900 and ALD310700, since, they structurally contain of EPAD MOSFETs with a matched transconductance and output conductance. In case of that M1, M2 and M3, M4 are matching, differential amplifier-based transconductance amplifier will operate in the differential-mode only in order to achieve the ideality and the desired behaviour. Consequently, Fig. 5 shows the equivalent low-frequency circuit model to the differential-mode.

\[ V_{di} = [g_m V_{di} - (g_m V_{di}^2/2)](r_{ds} g_m) \]  

Solving Eq. 15 for \( V_{di} \) gives:

\[ V_{di} = \frac{\frac{g_m r_{ds} g_m}{2} V_{di}^2}{2[1 + g_m r_{ds} g_m]} \]  

\[ V_i = i_{di} = \frac{g_m}{2} V_{di} \]  

Substituting the value of \( V_{di} \) from Eq. 16 in Eq. 18 gives:

\[ I_i \approx \frac{g_m}{2} V_{di} \left[ V_i / (r_{ds} g_m) \right] \]  

By substituting Eq. 19 in Eq. 17 and solving for yields:

\[ V_i \approx \frac{g_m R_L}{1 + [R_i / (r_{ds} g_m)]} V_i \approx g_m R_i V_{di} \]  

Since, all devices should be in the saturation region (operating region), the saturation current (\( D_{SAT} \)) of M4 by neglecting the channel length modulation parameter (\( \lambda \)) is given by:

\[ D_{SAT} = \mu_e c W / L V - V^2 \]  

Thus, the transconductance, \( g_m \) of M4 is:

\[ g_m = \frac{D_{m}}{V_{GSS} V_{TH}} \]  

or from Eq. 21, since:

\[ V_{GSS} V_{TH} = \frac{2D_{m}}{\mu_e c W / L} \]  

Thus, the transconductance, \( g_m \) of M4 is:

\[ g_m = \frac{D_{m}}{V_{GSS} V_{TH}} \]  

\[ g_m = \frac{2D_{m}}{\mu_e c W / L} \]
Fig. 6: The proposed topology of the NPTA

Hence, since, M3 and M4 are matching, then $I_{Dsat3} = I_{Dsat4}$. This means that $2I_{Dsat4} = I_{Dsat3} + I_{Dsat4} = I_{ref}$. Therefore, from Eq. 24, the transconductance $g_{m4}$ of the transconductance amplifier is:

$$g_{m4} = \sqrt{\frac{\mu c v}{L}} \left(\frac{W}{L}\right) I_{ref}$$  \hspace{1cm} (25)

where $\left(\frac{W}{L}\right) M4 = \left(\frac{W}{L}\right) M3$. In general, the output voltage, $V_o$ of the transconductance amplifier is:

$$V_o = g_{m2} R_L V_{in}$$  \hspace{1cm} (26)

Hence, by substituting Eq. 25 in Eq. 26 yields:

$$V_o = \sqrt{\frac{\mu c v}{L}} \left(\frac{W}{L}\right) I_{ref} R_L V_{in}$$  \hspace{1cm} (27)

Thus, the output current, $I_o$ of the transconductance amplifier in case of the differential-mode is:

$$I_o = \frac{V_o}{R_L} = \sqrt{\frac{\mu c v}{L}} \left(\frac{W}{L}\right) I_{ref} V_{in} = g_{m2}(V_1-V_2)$$  \hspace{1cm} (28)

From Eq. 25 and 28, it is observed that $I_o$ of the transconductance amplifier that having matching MOSFETs can be boosted by boosting the $I_{ref}$ and through using MOSFETs having large transconductance such as the EPAD MOSFETs. Consequently, in this research, the packages ALD212900 and ALD310700 that contain of matching N and P-type EPAD MOSFETs with a large are used in design the NPTA and the reference current $I_{ref}$ is boosted through the proposed the topology which is shown in Fig. 6.

In this topology, the current is mirrored by the current $I_{M1}$ mirror $M1$ and $M2$ on the drain of $M2$ ($I_{M2}$) and then, $I_{M2}$ which is equal to $I_{M1}$ is mirrored by the current mirror $M5$ and $M9$ on the drain of $M9$ ($I_{M9}$) to achieve that $I_{M9} = I_{M1}$ equals. The current $I_{M6}$ which is the difference between $I_{M6}$ and is mirrored by the current mirror $M6$ and $M7$ on the drain of $M7$ to achieve that $I_{M7}$ equals $I_{M6}$. Thus:

$$I_o = I_{ref} - I_{M6} = I_{M7} - I_{M1}$$  \hspace{1cm} (29)

Hence, since, $M1$, $M6$ and $M3$, $M8$ are matching(ALD310700 package and ALD212900 package, respectively), then, when no input voltages ($V_1$ and $V_2$) are applied $I_{M6}$ is equal to $I_{M1}$ or $I_{M7}$ is equal to $I_{M1}$. Therefore, the undesired output current due to the bias requirements is zero even if the output is short circuited. Thus, there is no error ratio in the output and the load voltage $V_o$ is closely equal to the bias voltage $VDD$ even with a very low load resistance. Through the proposed topology, the overall transconductance $g_{m0}$ of the NPTA can be increased through amplifying the reference current $I_{ref}$ as long as from Eq. 25 $g_{m0}$ is directly proportional to $I_{ref}$. $I_{ref}$ can be amplified through setting the drain-to-source
Voltage ($V_{DS}$) of the transistor M4 which is used as a current source to a large enough value that the drain to source current ($I_{DS}$) of M4 which is the $I_{ref}$ is saturated at the value set by the gate Voltage ($V_{GS}$). In this case, the drain-to-source voltage of M4 is much larger than the thermal Voltage ($V_T$) ($V_{DS} \gg V_T$) and the saturated $I_{ref}$ can be modelled in the saturation region as:

$$I_{ref} = I_{Ds} = I_{Ds} \approx I_{Ds} e^{\frac{\eta V_T}{V_T}}$$

Formulation of Eq. 30 in the saturation region, comes from the drain-to-source current ($I_{DS}$) relationship with the gate-to-source Voltage ($V_{GS}$) and drain-to-source voltage $V_{DS}$ in the sub-threshold region (weak inversion) which is given by the exponential relationship (Arora, 1993):

$$I_{DS} = I_{D0} e^{\frac{\eta V_T}{V_T}}$$

where:

$$I_{D0} = \frac{\mu_e V^2 T}{L} \left( \frac{qN_A \epsilon_{Si}}{2} \left( \frac{V_{GS} - V_{th}}{\eta} \right)^2 - V_T \right)$$

Where:
- $\mu_e$: The electron mobility [cm$^2$ (V.s)]
- $V_T$: The thermal voltage where $k = 1.38 \times 10^{-23}$ J/K is the Boltzmann’s constant
- $T$: The temperature in Kelvin and $q = 1.6 \times 10^{-19}$ Coulombs is the electron charge
- $W$: The gate Width
- $L$: The gate Length
- $N_A$: The holes concentration
- $V_p$: The drop voltage across the gate-oxide region and

$$\eta = 1 + \frac{1}{c_{ox} q N_A} \left( \frac{\epsilon_{Si} q N_A}{2} \right)$$

where $c_{ox}$ is the gate oxide capacitance per unit area and $\epsilon_{Si}$ is the silicon permittivity to the electric field. Thus, it is observed from Eq. 31 in case of that $V_{DS} \gg V_T$ the dependence of $I_{Ds}$ on $V_{DS}$ can be neglected but not completely because there is a residual sensitivity of the $I_{Ds}$ to $V_{DS}$. This case is similar to the saturation region for strong inversion. Therefore, on the condition that $V_{DS} \gg I_{DS}$ in the saturation region is given by Eq. 30. Hence, from the circuit of Fig. 6, since, the gate-to-source Voltage ($V_{GS}$) of M4 is:

$$V_{GS} = V_{DS} - V_{SS}$$

then by substituting Eq. 34 in Eq. 30, yields:

$$I_{ref} = I_{Ds} e^{\frac{\eta V_{GS} - V_{SS}}{V_T}}$$

or since, the MOSFETs used in this work are matching EPAD MOSFETs of zero threshold Voltage ($V_{TH} = 0$), the reference current in the saturation region can be approximately modelled as:

$$I_{ref} \approx I_{Ds} e^{\frac{\eta V_{GS}}{V_T}}$$

Consequently, since all EPAD MOSFETs are in the saturation region ($V_{DS} \gg V_T$), the transconductance $g_m$ of the proposed NPTA can be derived as following:

$$I_{M1} \approx I_{Ds} e^{\frac{V_{-V_{th}}}{\eta V_T}} = I_{Ds} e^{\alpha V_T}$$

$$I_{M6} \approx I_{Ds} e^{\frac{V_{-V_{th}}}{\eta V_T}} = I_{Ds} e^{\alpha V_T}$$

$$I_{ref} = I_{M1} + I_{M6} = I_{Ds} e^{-\alpha V_T} \left( e^{\alpha V_T} + e^{\alpha V_T} \right)$$

where, $\alpha = 1/($ $\eta V_T$ $)$ = $q/($ $\eta kT$ $)$. Hence, from Eq. 39:

$$e^{\alpha V_T} = \frac{I_{ref}}{I_{Ds}}$$

Substituting Eq. 40 in Eq. 37 and 38, respectively gives:

$$I_{M1} \approx I_{ref} e^{\alpha}$

$$I_{M6} \approx I_{ref} e^{\alpha}$$

Thus, since, $I_{M1} = I_{M2}$ and $I_{M6} = I_{M7}$

$$I_e = I_{M1} - I_{M2} = -I_{ref} e^{\alpha} \left( e^{\alpha V_T} + e^{\alpha V_T} \right)$$

Multiplying both the nominator and denominator of Eq. 43 by $e^{-V_{1+V_{2}/2}}$ gives:

$$I_e = I_{ref} e^{\alpha V_T} \left( e^{-\alpha V_T} - e^{-\alpha V_T} \right) = -I_{ref} \tanh \left( \frac{\alpha V_T}{2} \right)$$
Fig. 7: The proposed NPTA using the open-circuit voltage of the BPV cell

Hence, the transconductance \( g_m \) of the proposed NPTA is the slope of the hyperbolic tan in Eq. 44 at the origin, i.e:

\[
g_m = \frac{\partial I_o}{\partial (V_1-V_2)}|_{V_1=0, V_2=0} = \frac{-\alpha_{m}}{2} \text{sech} \left[ \frac{\alpha (V_1-V_2)}{2} \right] |_{V_1=V_2=0}
\]

or since, sech\(^2\) \( [\alpha (V_1-V_2)/2] \)|\( V_1=V_2=0 \) = 1

\[
g_m = \frac{-\alpha_{m}}{2} = \frac{I_o}{2kT/q}
\]

Equation 46 states that the overall \( g_m \) of the NPTA can be increased by boosting the reference current \( I_{ref} \) through the gate Voltage, \( V_{G4} \) of the transistor M4. Thus, the output current of the proposed NPTA is:

\[
I_o = g_m (V_1-V_2) = \frac{I_{ref}}{2kT/q} (V_1-V_2)
\]

Hence, by utilizing the open-circuit Voltage \( (V_{BPV}) \)of the BPV cell if \( V_1 = -V_{BPV}/2 \), \( V_2 = +V_{BPV}/2 \), \( VDD = +V_{BPV}/2 \), \( VSS = -V_{BPV}/2 \) and \( V_{G4} = +V_{BPV}/2 \) then from Eq. 47, the output current of the NPTA is expressed as:

\[
I_o = \frac{I_{ref}}{2kT/q} V_{BPV}
\]

Equation 48 states that the output current of the proposed NPTA depends on both the overall transconductance \( g_m \) and the open-circuit voltage of the BPV cell. Accordingly, Fig. 7 shows the proposed NPTA using the open-circuit Voltage \( V_{BPV} \) of the BPV cell.

RESULTS AND DISCUSSION

The electricity/area produced to be easily measured, the BPV cell used in this work was constructed using the Acrylic sheets by two cylindrical chambers separated by a Proton Exchange Membrane (PEM) of Nafion 117 the anodic chamber (diameter 10 cm and height 6 cm) and cathodic chamber (diameter 10 cm and height 2 cm). The anodic chamber was provided with one sealable opening which is necessary for performing the auxiliary operative activities while the cathodic chamber is open to the air to guarantee the presence of oxygen required for the cathodic reaction. The anode material chosen is Indium Tin Oxide (ITO) which is an effective material for electrons harvesting while the cathode material chosen is copper. In this cell, the living organism which is used as a harvesting material of the light is the cyanobacteria as it is capable of self-assembly and self-repair inside the cell.

The BPV cell performance was evaluated at 300°K, without using the proposed NPTA by means of polarization curve through varying the external Resistance \( (R_L) \) from 2 M\( \Omega \) to 5 \( \Omega \) where the voltages decrease from 150 to 10 mV while the current densities increase from 0-1.25 mA/m\(^2\) as reported in Fig. 8.

To evaluate the performance of the BPV cell in the power conversion without using the NPTA, the power curve was reported as shown in Fig. 9 where basing on the maximum power transfer theorem, the cell internal resistance of 120 \( \Omega \) was calculated.
Fig. 8: The polarization curve of the BPV cell without using the proposed NPTA

Fig. 9: The power curve of the BPV cell without using the proposed NPTA

Fig. 10: The polarization curve of the BPV cell using the proposed NPTA

In comparison, the polarization curve and the power curve of the BPV cell using the proposed NPTA are reported as shown in Fig. 10 and 11, respectively through varying the external resistance from 2 MΩ to 5 Ω at 300°K.

Fig. 11: The power curve of the BPV cell using the proposed NPTA

It is observed from Fig. 9 that without using the proposed NPTA, a Maximum Power Point (MPP) of 0.3 μW was recorded for the BPV cell under illumination, corresponding to a current density of 0.5 mA/m² and a voltage of 60 mV. While from Fig. 11 by using the proposed NPTA, the MPP recorded is 5.84 μW corresponding to a current density of 7.3 mA/m² and a voltage of 80 mV.

CONCLUSION

In this study, a new approach to optimize electrical power conversion potential of the BPVs was proposed through utilizing the nano-powered electronic circuit rather than using different strategies for develop the cell design. The proposed method exhibited a large increase in the maximum power conversion (from 0.3-5.84 μW). Which means that an increase of 20 times in the output power has been achieved from the BPV cell by using the nano-powered electronic circuit technology. Thus an effective power can be obtained from the BPVs to power the electronic devices through using array of cells for increasing the output maximum power point.

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