Design of a High Performance Resonant Controller for Improved Stability and Robustness of Islanded Three-Phase Microgrids

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ABSTRACT Islanded microgrids face difficulties due to the presence of nonlinearity, asynchronous load, and unknown load dynamics. Moreover, conventional control schemes in the islanded microgrids show slow dynamic response, significant voltage-current oscillations, frequency change, low output power quality, and less reference tracking capability. In this regard, a robust and high performance controller is required against the instability issues related to various load conditions and sudden load changes in the solar photovoltaic (PV)-based solar photovoltaic (PV) based islanded microgrids. This paper presents the design and implementation of a second-order high performance resonant controller for robustness and improving the stability of a solar PV based three-phase islanded microgrid (TPMG) under varying system conditions. The design of the proposed controller is based on a backstepping scheme where control Lyapunov functions are used to find transfer functions. The transfer functions that are obtained by this approach are the transfer functions of a resonant controller with proportional-integral controllers. The performance of the proposed controller is investigated in MATLAB/Simulink. The simulation results demonstrate the robustness of the proposed controller in terms of stability, dynamic responses, voltage-current oscillations, total harmonic distortion, and reference tracking of the TPMG. Moreover, the performance of the proposed controller is illustrated against various load dynamics and sudden load changes. A laboratory-scale experiment verifies the simulation results of the proposed controller.

INDEX TERMS Islanded microgrid, resonant controller, PI controller, total harmonic distortion.

I. INTRODUCTION

Global climate change, the exponential rise in global power demand from electrification, and the need to provide power to remote locations have drawn the attention of scientists and environmentalists toward increased use of renewable energy sources (RESs) [1]. Deployment of RESs, such as photovoltaics (PV), wind turbines, biomass, and hydropower-fueled sources reduces the requirement for traditional fossil-fueled resources, e.g., oil, coal, and natural gas to meet increased energy demand [2]. Notably, the microgrid concept has paved the way to integrate distributed RESs with local energy storage (ES) systems, power converters, local loads, and the main grid [3]. Moreover, microgrids can operate in either islanded or grid-connected mode, improving the economic and sustainability-driven benefits of the demand-side resources in the grid while improving reliability and resilience [4]. The optimal control schemes in these modes provide high performance in terms of stability, voltage-current oscillations, dynamic responses, reference tracking,
and total harmonic distortion (THD) [5]. Due to initial support from the main grid, the design of control schemes in the grid-connected mode is less challenging than in the islanded mode [6].

The use of solar PV is expanding day by day among all RESs due to increased investment in manufacturing technology which has yielded reduced costs and cell performance improvement while continuing to deliver value through zero emissions, fewer infrastructure challenges, and zero fuel costs with the static nature of PV resources [7]. Besides PV, ES systems are also critical components of microgrids that ensure stable operation and offset the variability of RESs [8]. Optimal microgrid control schemes are generally sought as a design feature requirement to offset microgrid challenges, including various load and generation conditions and sudden voltage changes in the PV plant, requiring microsource modeling, ES system modeling, inverter modeling, and load modeling [9]. The modeling of solar PV, ES systems, and local loads are intentionally ignored to reduce complexity [10]. However, the design of a controller requires inverter modeling to regulate active and reactive power injection [11], [12]. Nevertheless, several controllers have been implemented in the islanded three-phase microgrids (TPMGs), such as droop controller, proportional-integral-derivative (PID) controller, model predictive controller, fuzzy controller, proportional-resonant (PR) controller, sliding mode controller, linear-quadratic-regulator (LQR), backstepping controller, $H_{\infty}$ controller, and neural network controller [13].

Furthermore, two-stage power conversion systems have limitations over single-stage power conversion systems in the stability analysis [14]. Therefore, the dc-dc converter is avoided as the analysis becomes more complex regarding system parameter adjustment. However, the dc-link capacitor is used to remove voltage fluctuation of the inverter input voltage from the solar PV [15]. Nevertheless, maximum active power or minimum reactive power sharing is not the primary goal of the islanded inverter control [16]. The key issue is to fulfill the power demand of the local loads [17]. Moreover, the presence of non-linear load, balance-unbalance load, asynchronous load, and unknown load dynamics in the TPMG creates difficulty in controlling THD, voltage-current oscillation, less reference tracking, lower stability, and lower dynamic responses [18]. Several investigations have been carried out regarding these controlling issues in the recent past.

The droop control scheme is an islanded microgrid’s well-known voltage-frequency control strategy [19]. The droop control based on line impedance can be implemented in three ways [20]. Since the third approach requires a compensated structure and the common bus voltage information is difficult to obtain in practice, this approach is challenging to implement. [21]. The second method employs a signal injection scheme. However, it results in voltage distortion [22]. The first method involves virtual impedance, which results in the degradation of the system voltage quality [23]. However, load dynamics are not integrated into traditional droop control schemes directly. Therefore, rapid load changes can cause voltage-frequency instability [24]. Moreover, the droop control method has the disadvantages of slow dynamic response, compromising power-sharing reliability, and dependency on filter impedance [25].

For any islanded microgrid, the PID controller is the most conventional controller to regulate voltage or frequency. However, a PID controller is constrained by its low bandwidth and lack of robustness regarding voltage-current oscillations, reference tracking, and system dynamics [26]. Another control technique reported in [27] is hierarchical control for voltage-frequency regulation to achieve high bandwidth and steady-state output. This technique has demonstrated success. However, if a feature of the control fails, active-reactive power cannot be controlled [28]. The LQR method uses linearization techniques to achieve rapid and effective voltage-frequency regulation in the islanded microgrids. Nevertheless, the function of the LQR controller depends on the plant dynamics, which may change due to parameter variation or uncertainty. The robustness of the LQR controller may be limited [29]. The $H_{\infty}$ controller is an advanced control technique for the voltage–frequency control of the islanded microgrid. As this controller is highly dependent on the system dynamics, the higher order system includes the higher order controller, which is very complex and challenging to implement in a digital signal processor [30].

A resonant controller (RC) has a high gain around the resonant frequency. It can eliminate steady-state error and is best suited for TPMG operations. In [31], a second-order RC has been proposed for the single-phase and three-phase islanded microgrids to control grid voltages under different load conditions. An improved proportional P+RC was proposed in [32]. The reference tracking operation of these controllers against the grid voltage instability was ineffective. Furthermore, the effects of change in load dynamics and the THD of the grid voltage during the islanded mode have not been studied extensively in those papers. A synchronous reference frame proportional-integral (PI) regulator to control the current of three-phase converters was investigated in [33]. In [34], a fractional P+RC has been introduced to reduce the voltage-current harmonics for better system performance under non-linear load conditions. However, the paper did not show the operations under different load conditions.

In terms of controlling voltage frequency and minimizing grid voltage THD, an adaptive PR controller for synchronizing inverters connected to the grid has excellent performance [35]. However, the actions for the islanded mode have not been covered in this article. Applying a fuzzy-logic inference system considering the system frequency regulation requirements and battery energy storage system (BESS) state of charge (SoC), a degree of participation (DP) based secondary frequency control (SFC) for the BESS cluster has been proposed in [36]. The distribution of BESS in SFC has been achieved to minimize the operation cost of the BESS cluster by quadratic programming algorithm. However, this
The article does not discuss various load operations for the BESS. The use of a fractional-order integral proportional derivative with filter based on an imperialist competitive algorithm, also known as an integral tilt derivative with filter controller (ITDF), has been suggested in [37] for frequency control in two areas interconnected by MG (isolation mode) and renewable penetration. The transient responses during load perturbations, wind speed fluctuation, solar irradiance changes, system parametric, broad variations in renewable source unpredictability, and load disturbances exhibit the excellent performance of the ITDF control approach.

In [38], a dual-stage fractional order PID controller has been proposed to enhance the primary frequency regulation of interconnected multi-microgrids in the standalone mode. It reduces the total inertia of a single-area microgrid system pursued by a two-area microgrid to integrate a tidal power plant. A Fuzzy tilt integral derivative (FTID) with a filter plus double integral (FTIDF-II) control strategy for frequency control in a hybrid system has been proposed in [39]. Various transient response parameters, such as settling time, peak overshoots, and undershoots have been investigated to analyze the hybrid system’s dynamic performance evaluation. However, these control techniques do not report the grid voltage performances and THDs. A power-sharing control method based on the linear quadratic regulator with optimal reference tracking (LQR-ORT) has been proposed in [40] for three-phase inverter-based generators using LCL filters islanded mode. LQR-ORT controller for islanded and grid-connected microgrids improves transient response, accuracy on power sharing, and voltage and frequency regulation.

It has not, however, been exposed to non-linear load conditions. An adaptive nonlinear backstepping control scheme has been proposed in [41] to control a three-phase grid-connected inverter. Extracting maximum power from the PV, controlling active power, injecting fewer harmonic components into the grid, and excellent performances against different atmospheric conditions are the key features of the controller. However, this control technique increases mathematical complexity.

This paper proposes a high performance resonant controller using the backstepping technique where control Lyapunov functions (CLFs) are used to find the transfer functions. The relevant aspects of the proposed approach can be listed as,

- Regulation of active and reactive power by adjusting the phase angle and the amplitude of the output voltage;
- The total reduction of voltage oscillation and perfect reference tracking property;
- Reduction of the THD to a great extent ability;
- Robust performances against various load dynamics and sudden load change conditions; and
- High stability and fast dynamic response in the islanded microgrid operation.

This paper is organized as follows. Section II discusses the circuit topology and mathematical modeling of the TPMG are discussed. Section III explains the proposed controller’s design features and the necessary reference signal calculations. Section IV investigates the proposed controller’s performances in the TPMG system under various electrical and load conditions. An experimental investigation is carried out in section V. Finally, section VI concludes the paper.
The mathematical modeling of TPMG has been carried out at the point of common coupling (PCC).

Each distributed generation (DG) unit consists of solar PV, the circuit diagram of the islanded TPMG with various load parameters used in TPMG.

The system dynamics can be described as \[[25],[31]\], where the state matrix is \(A\), the input matrix is \(B\), the output matrix is \(C\), the uncertainty matrix is \(U\), the state vector is \(x\), the input vector is \(u\), the output vector is \(y\), and the uncertainty vector is \(d\). The s-domain representation of the system and uncertainty can be given as \([25]\):

\[
S_0(s) = C(sI - A)^{-1}B
\]

The detailed transfer function of the system is given in the appendix. \(\theta_i\) is the phase angle between the DG units and the common bus, and the common bus’s phase angle is assumed 0°. If \(P\) is the average active power and \(Q\) is the average reactive power, then the average active and reactive power feed into the microgrid can be expressed as \([42]\):

\[
P = 1.5(V_{pcc,d}I_{g,d} + V_{pcc,q}I_{g,q})
\]

\[
Q = 1.5(V_{pcc,d}I_{g,q} - V_{pcc,q}I_{g,d})
\]

### TABLE 1. Parameters Used in TPMG.

| Parameter at dc-link | Symbol | Values |
|----------------------|--------|--------|
| Voltage              | \(V_{dc}\) | 700 V  |
| Filter inductance    | \(L_f\) | 2 mH   |
| Filter capacitance   | \(C_f\) | 15 \(\mu\)F |
| Line impedance       | \(L_L\) | 0.45 mH |
| Resonance frequency  | \(w_r\) | 5.7 kHz |
| Carrier frequency    | \(w_c\) | 15 kHz  |
| Grid frequency       | \(f_g\) | 50 Hz   |
| Damping ratio        | \(\zeta\) | 0.7    |
| Positive constant    | \(\gamma\) | 100    |

### II. MODELING OF THE THREE-PHASE MICROGRID

#### A. CIRCUIT TOPOLOGY

The circuit diagram of the islanded TPMG with various load dynamics is depicted in Fig. 1. Solar PV is implemented to meet the power demand of the local loads and the microgrid. The closed-loop system with the proposed controller provides gate pulses for the three-phase inverter. The inverter and PV are connected through the dc-link. An LC filter eliminates the converters’ high frequency harmonics and nonlinearity. Each distributed generation (DG) unit consists of solar PV, an inverter, a controller, and a filter. Local loads are connected with the DG units. The DG units are connected through a point of common coupling (PCC).

#### B. MATHEMATICAL MODELING

The mathematical modeling of TPMG has been carried out in this sub-section. If \(V_{dc}\) is the inverter input voltage, \(V_{in}\) is the inverter output voltage, \(I_{L}\) is the current through the inductor, \(I_{g}\) is the grid current, \(V_{g}\) is the grid voltage, \(V_{pcc}\) is the common bus voltage, \(L_f\) is the filter inductance, \(C_f\) is the filter capacitance, \(f_g\) is the grid frequency, and \(L_i\) is the impedance between DG units and the common bus, then the system dynamics can be described as \([25],[31]\):

\[
\begin{align*}
I_{L,d} &= L_f^{-1}(V_{in,d} - V_{g,d}) + \omega_0 I_{L,q} \\
I_{L,q} &= L_f^{-1}(V_{in,q} - V_{g,q}) - \omega_0 I_{L,d} \\
\dot{V}_{g,d} &= C_f^{-1}(L_f^{-1} - I_{g,d}) + \omega_0 V_{g,d} \\
\dot{V}_{g,q} &= C_f^{-1}(L_f^{-1} - I_{g,q}) - \omega_0 V_{g,d} \\
\dot{I}_{g,d} &= L_f^{-1}(V_{g,d} - V_{pcc,d}) + \omega_0 I_{g,d} \\
\dot{I}_{g,q} &= L_f^{-1}(V_{g,q} - V_{pcc,q}) - \omega_0 I_{g,q} \\
V_{in,d,q} &= V_{dc}\delta_{dq}
\end{align*}
\]

where \(\omega_0 = 2\pi f_g\) and \(\delta_{dq}\) is the duty cycle. TPMG design parameters are listed in Table 1. The state-space modeling of the system can be written from (1)–(3) as:

\[
A = \begin{bmatrix} 0 & \omega_0 & -L_f^{-1} & 0 & 0 & 0 \\ -\omega_0 & 0 & 0 & -L_f^{-1} & 0 & 0 \\ 0 & \omega_0 & -C_f^{-1} & 0 & 0 & \omega_0 \\ 0 & C_f^{-1} \omega_0 & 0 & 0 & -C_f^{-1} \\ 0 & 0 & L_f^{-1} & 0 & 0 & \omega_0 \\ 0 & 0 & 0 & L_f^{-1} & -\omega_0 & 0 \end{bmatrix}
\]

#### III. MODELLING OF THE PROPOSED CONTROLLER

The tracking errors \(e_1\) and \(e_2\) correspond to the inverter-side inductor current. Similarly, errors \(e_3\) and \(e_4\) correspond to the inverter capacitor voltage, and \(e_5\) and \(e_6\) correspond to the grid-side inductor current. These errors are required to design the proposed controller for the inverter. These tracking errors can be expressed as \(e_1 = I_{g,d} - I_{g,d}^c\), \(e_2 = I_{g,q} - I_{g,q}^c\), \(e_3 = V_{g,d} - V_{g,d}^c\), \(e_4 = V_{g,q} - V_{g,q}^c\), \(e_5 = I_{L,d} - I_{L,d}^c\), and \(e_6 = I_{L,q} - I_{L,q}^c\). The derivative of \(e_1\) and \(e_2\) can be written from (1) as \([43]\):

\[
\begin{align*}
\dot{e}_1 &= \dot{I}_{g,d} - \dot{I}_{g,d}^c = L_f^{-1}(V_{g,d} - V_{pcc,d}) + \omega_0 I_{g,q} - \dot{I}_{g,q}^c \\
\dot{e}_2 &= \dot{I}_{g,q} - \dot{I}_{g,q}^c = L_f^{-1}(V_{g,q} - V_{pcc,q}) - \omega_0 I_{g,d} - \dot{I}_{g,d}^c
\end{align*}
\]

According to the Lyapunov stability criteria, the derivatives of the CLFs of errors \(e_1\) and \(e_2\), then \(W_1\) and the derivative of \(W_1\) can be written as \([43]\):

\[
W_1 = 0.5(e_1^2 + e_2^2)
\]

\[
\dot{W}_1 = e_1(I_{g,d}^{-1}(V_{g,d} - V_{pcc,d}) + \omega_0 I_{g,q} - \dot{I}_{g,q}^c) \\
+ e_2(I_{g,q}^{-1}(V_{g,q} - V_{pcc,q}) - \omega_0 I_{g,d} - \dot{I}_{g,d}^c)
\]
where $I_{g_a}^c$ and $I_{g_q}^c$ can be evaluated from (6) as,

$$I_{g_a}^c = (P + Q)/(3V_{pcc,d}), \quad I_{g_q}^c = (P - Q)/(3V_{pcc,q}) \quad (10)$$

To ensure stability, the derivative of $W_1$ must be negatively defined. For this purpose, $V_{g_a}^c$ and $V_{g_q}^c$ can be selected from (9) as [43],

$$V_{g_a}^c = V_{pcc,d} + L_d(I_{g_a}^c - \omega_0I_{g_a} - \gamma e_1)$$
$$V_{g_q}^c = V_{pcc,q} + L_d(I_{g_q}^c + \omega_0I_{g_d} - \gamma e_2) \quad (11)$$

where $\gamma$ is a positive constant and $\gamma = 100$. $V_{g_a}^c$ and $V_{g_q}^c$ are the virtual control variables (VCVs) of errors $\dot{e}_1$ and $\dot{e}_2$. Furthermore, the derivative of the errors $e_3$ and $e_4$ can be written from (2) as [43],

$$\dot{e}_3 = \dot{V}_{g_a} - \dot{V}_{g_a}^c = C_f^{-1}(I_{g_a}^c - I_{g_a,d}) + \omega_0V_{g_a} - V_{g_a}^c$$
$$\dot{e}_4 = \dot{V}_{g_q} - \dot{V}_{g_q}^c = C_f^{-1}(I_{g_q}^c - I_{g_q,d}) - \omega_0V_{g_q} - V_{g_q}^c \quad (12)$$

If $W_2$ is the CLF of the errors $\dot{e}_3$ and $\dot{e}_4$, then $W_2$ and the derivative of $W_2$ can be written as [43],

$$W_2 = 0.5(e_3^2 + e_4^2) $$
$$W_2 = e_3(C_f^{-1}(I_{g_a} - I_{g_a,d}) + \omega_0V_{g_a} - \dot{V}_{g_a}^c)$$
$$+ e_4(C_f^{-1}(I_{g_q} - I_{g_q,d}) - \omega_0V_{g_q} - \dot{V}_{g_q}^c) \quad (13)$$

To ensure stability, the derivative of $W_2$ must be negatively defined. For this purpose, $I_{g_a}^c$ and $I_{g_q}^c$ can be selected from (14) as [43],

$$I_{g_a}^c = I_{g_a,d} + C_f(\dot{V}_{g_a} - \omega_0V_{g_a} - \gamma e_3)$$
$$I_{g_q}^c = I_{g_q,d} + C_f(\dot{V}_{g_q} - \omega_0V_{g_q} - \gamma e_4) \quad (15)$$

$I_{g_a}^c$ and $I_{g_q}^c$ are the VCV of the errors $\dot{e}_3$ and $\dot{e}_4$. Finally, the derivative of the errors $e_5$ and $e_6$ can be written from (3) as [43],

$$\dot{e}_5 = \dot{I}_{g_a,d} - I_{g_a}^c = L_f^{-1}(V_{in,a} - V_{g_a,d}) + \omega_0I_{g_a} - I_{g_a}^c$$
$$\dot{e}_6 = \dot{I}_{g_q,d} - I_{g_q}^c = L_f^{-1}(V_{in,q} - V_{g_q,d}) - \omega_0I_{g_q} - I_{g_q}^c \quad (16)$$

If $W_3$ is the CLF of the errors $\dot{e}_5$ and $\dot{e}_6$, then $W_3$ and the derivative of $W_3$ can be written as [43],

$$W_3 = 0.5(e_5^2 + e_6^2) $$
$$W_3 = e_5(L_f^{-1}(V_{in,a} - V_{g_a,d}) + \omega_0I_{g_a} - I_{g_a}^c)$$
$$+ e_6(L_f^{-1}(V_{in,q} - V_{g_q,d}) - \omega_0I_{g_q} - I_{g_q}^c) \quad (17)$$

To ensure stability, the derivative of $W_3$ must be negatively defined. For this purpose, $V_{in,a}^c$ and $V_{in,q}^c$ can be selected from (18) as [43],

$$V_{in,a}^c = V_{g,a} + L_d(I_{g_a}^c - \omega_0I_{g_a} - \gamma e_5)$$
$$V_{in,q}^c = V_{g,q} + L_d(I_{g_a}^c + \omega_0I_{g_d} - \gamma e_6) \quad (19)$$

$V_{in,a}^c$ and $V_{in,q}^c$ are the VCV of $e_5$ and $e_6$. Furthermore, the gate pulses for the insulated-gate bipolar transistors (IGBTs) in the inverter are achieved through sinusoidal pulse width modulation (SPWM). The magnitude and angle for SPWM are achieved from $S_{dq}$, where $S_{dq} = V_{in,a}^c/V_{dc}$. Consequently, in order to obtain $S_q$ and $S_a$, it is essential to achieve the reference signals $I_{g_a}^c$, $I_{g_q}^c$, $V_{g_a}^c$, $V_{g_q}^c$, $I_{g_a}^c$, and $I_{g_q}^c$ from the proposed controller’s transfer functions. Therefore, the proposed controller’s transfer functions will be generated by implementing (10), (11), (15), and (19) equations. Moreover, the equations (11), (15), and (19) must be represented as [43],

$$I_{g_a} = L_f^{-1}(V_{g_a} - V_{pcc,a}) + \omega_0I_{g_a} + \gamma e_1$$
$$I_{g_q} = L_f^{-1}(V_{g_q} - V_{pcc,q}) - \omega_0I_{g_q} + \gamma e_2$$
$$V_{g_a} = C_f^{-1}(I_{g_a,d} - I_{g_a}) + \omega_0V_{g_a} + \gamma e_3$$
$$V_{g_q} = C_f^{-1}(I_{g_q,d} - I_{g_q}) - \omega_0V_{g_q} + \gamma e_4$$
$$I_{g_a} = L_f^{-1}(V_{in,a} - V_{g_a,d}) + \omega_0I_{g_a} + \gamma e_5$$
$$I_{g_q} = L_f^{-1}(V_{in,q} - V_{g_q,d}) - \omega_0I_{g_q} + \gamma e_6 \quad (20)$$

The proposed controller’s overall transfer function can be represented from (20) as [25],

$$S_T = C_T \left( sI - A_T \right)^{-1} B_T + D_T \quad (21)$$

where the state matrix is $A_T$, the input matrix is $B_T$, the output matrix is $C_T$, the feedforward matrix is $D_T$, the state vector is $x_T$, and the input vector is $u_T$. These vectors and matrices can be expressed as shown in the equation at the bottom of the next page [25].

Nevertheless, the reference signals $I_{g_a}^c$, $I_{g_q}^c$, $V_{g_a}^c$, $V_{g_q}^c$, $I_{g_a}^c$, $I_{g_q}^c$, $V_{in,a}^c$, and $V_{in,q}^c$ are achieved through the proper selection of $C_T$ and $D_T$. To obtain $V_{g_a}$ and $C_T$ and $D_T$ can be selected as $C_T = [0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0]$, $D_T = [-\gamma L_4 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0]$. The respected part of the proposed controller can be selected using (21) as,

$$S_{T_{V_{g_a}}} = (k_{r1}) \left( s^2 + 2\gamma k_{a1} s + b k_{c1}^2 \right) \quad (22)$$

where $k_{r1} = 0.81$, $k_{a1} = -5.82 \times 10^4$, $b = -1$, $k_{c1} = 2.65 \times 10^5$, and $k_{w} = 2\pi \times 50$. Similarly, $C_T$ and $D_T$ can be selected as $C_T = [0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0]$, $D_T = [0 \ 0 \ -\gamma C_T \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0]$. The respected part of $I_{g_a}$ is achieved from (21) as,

$$S_{T_{I_{g_a}}} = (k_{r2}) \left( s^2 + 2\gamma k_{a2} s + b k_{c2}^2 \right) \quad (23)$$

where $k_{r2} = 1.1$, $k_{a2} = -4.34 \times 10^4$, and $k_{c2} = 6.58 \times 10^3$. However, $C_T$ and $D_T$ can be selected as $C_T = [0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0]$, $D_T = [0 \ 0 \ -\gamma C_T \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0]$. The respected part of the proposed controller can be selected using (21) as,

$$S_{T_{I_{g_a}}} = k_{p1} + \frac{k_{i1}}{s} \quad (24)$$

where $k_{p1} = 0.99$ and $k_{i1} = -85.84$. To obtain $I_{g_a}$, $C_T$ and $D_T$ can be selected as $C_T = [0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0]$, $D_T = [0 \ 0 \ -\gamma C_T \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 0]$. The respected part of the
proposed controller can be selected using (21) as,

\[ S_{T,V_{in,q}} = k_{p2} + \frac{k_{q2}}{s} \]  (25)

where \( k_{p2} = 1 \) and \( k_{q2} = -714.2 \). To obtain \( V_{C_{in,q}} \), \( C_T \) and \( D_T \) can be selected as \( C_T = [0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 1] \) and \( D_T = [0 \ 0 \ 0 \ 0 \ -\gamma L_f \ 0 \ 0 \ 0 \ 0 \ -\omega_o L_f \ 0] \). The respected part of the proposed controller can be selected using (21) as,

\[ S_{T,V_{in,q}} = k_{p3} \]  (26)

where \( k_{p3} = 0.17 \). Furthermore, \( C_T \) and \( D_T \) are selected as \( C_T = [0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 1] \) and \( D_T = [0 \ 0 \ 0 \ 0 \ -\gamma L_f \ 0 \ 0 \ 0 \ 0 \ -\omega_o L_f \ 0] \) to obtain \( V_{in,q}^{c} \). The proposed part of the proposed controller is selected using (21) as,

\[ S_{T,V_{in,q}} = k_{p4} \]  (27)

where \( k_{p4} = 1.43 \). It is evident from equations (22)–(27) that the proposed controller is a combination of second-order resonant controllers (SORCs), proportional-integral (PI) controllers, and gains. Moreover, the proposed controller’s design procedure includes uncertainty in the system. Briefly, the proposed control scheme can be illustrated as follow. Active power \( P \), reactive power \( Q \), and the common bus voltage \( V_{pcc} \) are used to achieve the grid current references \( I_{g_{dq}}^{c} \). The errors \( e_1 = I_{g_{d}} - I_{g_{d}}^{c} \) and \( e_2 = I_{g_{q}} - I_{g_{q}}^{c} \) are fed into the SORC to achieve the grid voltage references \( V_{g_{dq}}^{c} \). Grid current reference tracking is feasible with this control loop. The errors \( e_3 = V_{g_{d}} - V_{g_{d}}^{c} \) and \( e_4 = V_{g_{q}} - V_{g_{q}}^{c} \) are fed into the PI controller to achieve the inductor current references \( I_{L_{dq}}^{c} \). Grid voltage reference tracking is feasible with this control loop. The errors \( e_5 = I_{L_{d}} - I_{L_{d}}^{c} \), and \( e_6 = I_{L_{q}} - I_{L_{q}}^{c} \) are fed into the gains to achieve the inverter output voltage references \( V_{C_{in,q}}^{c} \). This control loop is utilized to attain fast dynamic responses, high bandwidth, and precise duty cycle. It is evident from (4) that the accurate duty cycle \( S_{dq} \) is achieved through the ratio of the inverter output voltage references \( V_{in,dq}^{c} \) and the inverter input voltage \( V_{dc} \). Therefore, the proposed controller handles uncertainty and follows Lyapunov stability criteria. Finally, the SPWM is achieved from \( S_{dq} \) as (25),

\[ S_{dq} = \frac{S_{d} = V_{dc}^{-1}(S_{T,V_{g_{dq}}} + S_{T,I_{L,d}} + S_{T,V_{in,q}})}{S_{q} = V_{dc}^{-1}(S_{T,V_{g_{dq}}} + S_{T,I_{L,q}} + S_{T,V_{in,q}})} \]  (28)

**IV. PERFORMANCE EVALUATION**

The performance of the controller is evaluated in MATLAB/Simulink. The performances of the PID [26], RC [31], and the proposed controller are investigated in terms of stability, grid voltage, and current oscillations against various load dynamics, sudden load change conditions, reference tracking capability, transient response, and THD.

**A. STABILITY**

Lyapunov stability criteria have ensured the stability of the proposed controller in sections III through (7)–(19). The proposed control scheme is depicted in Fig. 2 using (22)–(27). Fig. 3 illustrates the frequency characteristics of the PID, the RC, and the proposed controller. The PID controller has +3 dB gain until 112 Hz and a negative gain after 112 Hz. Thus, it contains all harmonics before 112 Hz and has a small gain at 50 Hz. However, The RC controller has a negative gain until 42 Hz and +3 dB gain after 42 Hz. Thus, it also contains harmonics and has a small gain at 54.3 Hz. Nevertheless, the
The block diagram of the proposed control scheme (a) in detail and (b) briefly. 

The bode diagram of the different controllers' transfer functions. 

The bode diagram of the different controllers' loop gain. 

The proposed controller has a negative gain from 0.5 Hz to 42 Hz and after 63 Hz, although it has a positive gain before 0.5 Hz and from 42 Hz to 63 Hz. Thus, the proposed controller has fewer harmonics and a very high gain at the fundamental frequency. Negative Imaginary theory provides one of the most common strategies for determining the stability of a system [44]. According to this theory, the negative feedback of two positive-real transfer functions is stable with loop gain phase variation from 0 to $-180^\circ$ [44]. The $S_o(s)$ system phase varies from 0° to $-180^\circ$ for $\omega \geq 0$. The proposed controller $S_T(s)$ phase ranges from 180° to 0° for $\omega \geq 0$, asserting them as positive-real transfer functions depicted in Fig. 3. 

Fig. 4 depicts the loop gain of different controllers. The loop gain phase of the proposed controller varies from 0° to $-180^\circ$ for $\omega \geq 0$. Therefore, the negative feedback of $S_o(s)$ and the proposed controller is stable and does not intersect the negative real axis. However, it has a higher gain around the resonance frequency $5.8 \times 10^3$ Hz. 

The frequency responses of the closed-loop gains of different controllers, $S_{cl} = S_o/(1-S_o S_T)$, are depicted in Fig. 5. 

### B. PERFORMANCE OF TMPG WHERE LOAD DYNAMICS PERFORMED IN PARALLEL

The microgrid system with distributed generations (DGs) shown in Fig. 1 employs an asynchronous load, a balanced load, a non-linear load, and an unknown load. The asynchronous load is modeled as a 4.5 HP (4 kW) three-phase squirrel-cage induction motor operating at 50 Hz, 1430 RPM, and 400 V line-line voltage. Two parallel-integrated six-pulse diode bridge rectifiers are modeled as the non-linear load dynamics. The three-phase unknown load dynamics represent a 12 µF capacitor parallel with a serially connected 28 Ω
resistor and a 247 mH inductor. The balanced load dynamics is modeled as a 2 kW, 1 kVar, three-phase load. Torque, $T_{sh/2} = (4000 \times 60)/(2\pi \times 1430 \times 2) = 13.36$ Nm, is applied initially in the three-phase induction motor. Figs. 68 show the grid voltages with PID, RC, and the proposed controllers, respectively. Figs. 9–11 depict the grid currents with PID, RC, and the proposed controllers, respectively. The PID controller’s inclusion in the microgrid system’s feedback results in extremely high oscillations in the grid’s voltage and current, as shown in Figs. 6 and 9, respectively. On the other hand, the integration of the RC generates medium and high oscillations in the grid voltage and current, which are depicted in Figs. 7 and 10, respectively. However, Figs. 8 and 11 exhibit zero oscillations in the grid voltage and current by implementing the proposed controller in the islanded microgrid. DGs effectively distribute power among them during parallel operations.

C. SUDDEN LOAD CHANGE EFFECT WHERE LOAD DYNAMICS PERFORMED SEPARATELY

1) Performance against the three-phase unknown load conditions: A 123 mH inductor, a 76 resistor, a 228 resistor, and a 15 F capacitor are connected as shown in Fig. 1 to raise the three-phase unknown load dynamics in the existing unknown load state while other load dynamics in the microgrid are disconnected from the microgrid from $t = 0.2-0.22$ s. When the

FIGURE 5. The bode diagram of the different controllers’ closed loop.

FIGURE 6. The grid voltage with THD for the PID controller when all load dynamics are performed together.

FIGURE 7. The grid voltage with THD for the RC when all load dynamics are performed together.

FIGURE 8. The grid voltage with THD for the proposed controller when all load dynamics are performed together.

FIGURE 9. The grid current with THD for the PID controller when all load dynamics are performed together.
The grid current with THD for the RC when all load dynamics are performed together.

FIGURE 10. The grid current with THD for the RC when all load dynamics are performed together.

The grid current with THD for the proposed controller when all load dynamics are performed together.

FIGURE 11. The grid current with THD for the proposed controller when all load dynamics are performed together.

load is suddenly raised, the three-phase grid current increases and develops strong oscillations with the integration of the PID controller. The current returns to the initial state at $t = 0.31$ s even when the increased load condition is eliminated at $t = 22$ s, as shown in Fig. 12. However, as illustrated in Fig. 13, the three-phase grid current integrates the RC controller. It returns to the initial state at $t = 0.27$ s. Additionally, it exhibits significant grid current oscillations. On the other hand, the grid current of the proposed controller recovers to the initial state faster than existing controllers because of its large bandwidth, high phase margin, and short settling time. When the torque is suddenly increased, the three-phase stator current rises and develops high oscillations with the integration of the PID controller. Even though the high torque situation is eliminated at $t = 22$ s, the current takes a very long time to restore to the initial state at $t = 0.33$ s, as seen in Fig. 15. However, as shown in Fig. 16, the integration of the RC controller causes the three-phase stator current to recover to the initial state at $t = 0.31$ s. On the other hand, as illustrated in Fig. 17, the enhanced three-phase stator current at $t = 0.2$ s returns to the initial state after a reasonable amount of time at $t = 0.26$ s under the proposed controller. The stator current oscillations are also zero under the proposed controller.

D. THE REFERENCE TRACKING CAPABILITY OF VARIOUS CONTROLLERS

The grid voltage reference tracking capability of the closed-loop system with the PID, RC, and proposed controller for
various load conditions are compared in Figs. 18–21. It is observed that the grid voltages do not accurately track the reference signal with the PID controller and contain harmonics, oscillations, and phase errors. The grid voltages barely follow the reference signal with the RC controller and include phase inaccuracy from oscillations. With the proposed controller, there is almost no phase error and distortion in the grid voltages as they follow the reference.

E. THD EVALUATION FOR VARIOUS CONTROLLERS
Table 2 displays the THDs of the grid voltages and currents for the PID, RC, and proposed controller for “phase-a” under various load conditions. With the proposed controller, the lowest THD of 0.12% is achieved for grid voltage under asynchronous load conditions, outperforming all previous controllers (e.g., 8.19% for the PID and 2.82% for the RC). The proposed system controls grid current under asynchronous load conditions with a minimal THD of 0.47%, which is superior to all other controllers (e.g., 4.78% for the PID and 0.98% for the RC). The proposed system outperforms all previous controllers (e.g., 2.19% for the PID and 1.25% for the RC), achieving a low THD of 0.19% for grid voltage under balanced load conditions. The proposed controller performs grid current under balanced load conditions with the lowest THD of 3.28%, which is superior to all other controllers (e.g., 9.21% for the PID and 4.22% for the RC, for example). With the lowest THD of 0.12% is achieved for grid voltage under asynchronous load conditions, outperforming all previous controllers (e.g., 8.19% for the PID and 2.82% for the RC). The proposed controller provides grid current under all integrated load conditions with a minimum THD of 2.19%, which is superior to all other controllers (e.g., 9.47% for the PID and 6.78% for the RC). With the proposed controller, the

FIGURE 14. The effect of sudden load increment in the grid current for the unknown load dynamics performed separately in the microgrid with the proposed controller.

FIGURE 15. The effect of sudden torque increment in the stator current for the asynchronous load dynamics performed separately in the microgrid with PID controller.

FIGURE 16. The effect of sudden torque increment in the stator current for the asynchronous load dynamics performed separately in the microgrid with RC.

FIGURE 17. The effect of sudden torque increment in the stator current for the asynchronous load dynamics performed separately in the microgrid with the proposed controller.
TABLE 2. Comparative Study of THD for different controllers.

| Load types                                                                 | Voltage | PID [26] | RC [31] | Proposed (SORC+PI) |
|----------------------------------------------------------------------------|---------|----------|---------|--------------------|
| All loads integrated (asynchronous load + balanced load + non-linear load + unknown load) | Voltage | 5.12%    | 1.86%   | 0.17%              |
|                                                                             | Current | 9.47%    | 6.78%   | 2.19%              |
| Asynchronous load (4.5 HP (4 kW) 50 Hz 1430 RPM 400 V line-line voltage three-phase squirrel-cage induction motor) | Voltage | 8.19%    | 2.82%   | 0.12%              |
|                                                                             | Current | 4.78%    | 0.98%   | 0.47%              |
| Balanced load (2 kW 1 kVAR three-phase load)                                | Voltage | 2.19%    | 1.25%   | 0.19%              |
|                                                                             | Current | 3.73%    | 2.98%   | 1.12%              |
| Non-linear load (six-pulse diode bridge)                                    | Voltage | 12.25%   | 2.89%   | 1.51%              |
|                                                                             | Current | 9.21%    | 4.22%   | 3.28%              |
| Unknown load (A 12 μF capacitor in parallel with serially connected a 28 Ω resistor and a 247 mH inductor) | Voltage | 7.12%    | 4.63%   | 2.03%              |
|                                                                             | Current | 36.26%   | 12.89%  | 4.69%              |

proposed controller, the lowest THD of 2.03% is attained for grid voltage under unknown load situations, outperforming all existing controllers (e.g., the PID’s 7.12% and the RC’s 4.63%). A minimum THD of 4.69% is performed for grid current under unknown load conditions with the proposed controller, which is better than all other controllers (e.g., 36.26% for the PID and 12.89% for the RC).

F. DYNAMIC RESPONSE EVALUATION FOR VARIOUS CONTROLLERS

Table 3 compares the dynamic responses using the proposed controller, the PID [26], and the RC [31]. The proposed controller has a 2.0 μs settling time, which is faster than all other controllers (e.g., 10.7 μs for the PID and 5.4 μs for the RC). The proposed controller has a rise time of 0.233 μs, which is longer than the other controllers (0.016 μs for the PID and 0.183 μs for the RC). The proposed controller has a worse peak time of 0.598 μs for the system than the other controllers (e.g., 0.156 μs for the PID and 0.524 μs for the RC). The system’s overshoot under the proposed controller is 31.4%, compared to 98.48% under the PID and 72.7% under the RC.

V. EXPERIMENTAL RESULTS

A 2.5 kW, 230 V laboratory test platform is set up for the experimental validation of the proposed controller, as shown in Fig. 22. The AMETEK TerraSAS Photovoltaic Simulator
TABLE 3. Comparative transient response of different controllers.

| Name of the Controller | Overshoot (%) | Settling Time (μs) | Rise Time (μs) | Peak Time (μs) |
|------------------------|---------------|--------------------|----------------|---------------|
| Proportional-integral-derivative controller (PID) [26] | 98.48 | 10.7 | 0.016 | 0.156 |
| Resonant controller (RC) [31] | 72.7 | 5.4 | 0.183 | 0.524 |
| Proposed second-order resonant controller with proportional-integral controller (SORC+PI) | 31.4 | 2.0 | 0.233 | 0.598 |

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FIGURE 23. The three-phase load voltages at the start-up under non-linear load with the (a) traditional PID controller, (b) RC, and (c) proposed controller. The three-phase currents under non-linear load under load unbalance conditions with the (d) traditional PID controller, (e) RC, and (f) proposed controller.

FIGURE 24. The three-phase load voltages with the (a) traditional PID controller, (b) RC, and (c) proposed controller under unbalanced load conditions. The three-phase currents with the (d) traditional PID controller, (e) RC, and (f) proposed controller under unbalanced load conditions.

supplies 250 V dc to the three-phase inverter module. The three-phase inverter module uses SKM50GB12T4 Si IGBTs (50 A, 1200 V). The dSPACE 1104 controller board generates the switching control signals, and a switching frequency of 5 kHz is chosen. The three-phase voltages and currents are filtered using a second-order LC filter (0.5 mH, 15 F). The 4.5 kW, 350 V Chroma three-phase programmable ac/dc electronic load, receives the generated power. The voltage and current signals are gathered using KEYSIGHT N2791A differential voltage probes and Agilent N2781A differential current probes, respectively. KEYSIGHT N2791A differential voltage probes and Agilent N2781A differential current probes are used to collect the voltage and the current signals, respectively. The voltage and current signals are observed using a KEYSIGHT InfiniVision DSOX4024A digital oscilloscope. Figure 23 compares the performance of the proposed controller with that of the conventional PID controller, the RC, under rapid load change circumstances. An identical sudden load change condition, as mentioned in section IV.C is adopted. After lowering the load, the performance of a controller improves with decreasing stabilizing time. The voltage and current signals stabilize with the PID controller and the RC after 122 ms and 110 ms, respectively, as illustrated in Figs. 23(a), 23(b), 23(d), and 23(e), due to the load decrease impact. However, the proposed controller’s voltage and current signals balance out after 82 ms, as seen in 23(c) and 23(f). Compared to the PID controller and the RC, the proposed controller’s response time under non-linear load conditions is reduced by 32.78% and 25.45%, respectively.

When the TPMG is feeding power to the non-linear load, as illustrated in Fig. 24, the performance of the proposed controller with that of the conventional PID controller, the RC, under rapid load change circumstances. An identical sudden load change condition, as mentioned in section IV.C is adopted. After lowering the load, the performance of a controller improves with decreasing stabilizing time. The voltage and current signals stabilize with the PID controller and the RC after 122 ms and 110 ms, respectively, as illustrated in Figs. 23(a), 23(b), 23(d), and 23(e), due to the load decrease impact. However, the proposed controller’s voltage and current signals balance out after 82 ms, as seen in 23(c) and 23(f). Compared to the PID controller and the RC, the proposed controller’s response time under non-linear load conditions is reduced by 32.78% and 25.45%, respectively.

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controller is also tested and compared in terms of the start-up voltage transient and unbalanced load situations. Unstable grid voltage with the PID controller is depicted in Fig. 24(a). Additionally, the voltages have a sizable harmonic content. Figure 24(b) shows how the RC performed in operation. The three-phase voltages are unbalanced and unstable even though the harmonic contents are significantly reduced. The performance of the proposed controller under the non-linear load is shown in Fig. 24(c). As can be seen, the proposed controller can ensure stable grid voltages with a THD of 0.12%. The currents are distorted with the PID controller under non-linear load, as seen in Fig. 24(d). Additionally, it takes the PID controller 80 ms to steady the signals with abrupt changes in load. As shown in Fig. 24(e), the response time may be shortened and current signal distortions can be minimized with the RC. The time it takes to stabilize is still 50 ms. Finally, Fig. 24(f) shows how well the proposed controller performs. The proposed controller effectively lowers current distortions and under sudden load change conditions, the response time to regain stability further decreases to 30 ms. Due to high bandwidth, high phase margin, and less settling time, the grid current of the proposed controller returns to the previous state faster than others.

VI. CONCLUSION

This paper proposes a high-performance resonant controller for a solar PV-based three-phase islanded microgrid to improve stability and robustness under various loads and for a solar PV-based three-phase islanded microgrid to improve stability and robustness under various loads and sudden load change conditions. The design of the proposed controller is based on a backstepping scheme where control Lyapunov functions are used to find transfer functions and resonant controllers with proportional-integral controllers resulting from these transfer functions. The simulation results demonstrate the robustness and high performances of the proposed controller in terms of stability, voltage-current oscillations, THD, reference tracking ability, and dynamic responses in solar PV-based TPG applications. With the proposed controller, an asynchronous load condition’s grid voltage can have a minimal THD of 0.12%. Grid current under asynchronous load situations is performed with the proposed controller at the lowest THD of 0.47%. A settling time of 2 µs is achieved through the proposed controller. Moreover, the high performances of the proposed controller are observed against various load dynamics and sudden load change conditions. The superior performance of the proposed controller is validated by extensive simulation and laboratory scale experiments. In conclusion, the proposed controller can be a good candidate for implementing the modern three-phase microgrid system. In the future, it will be essential to consider the transient responses to changes in solar irradiation, wide variations in the unpredictability of renewable sources, and load disturbances.

APPENDIX

The transfer function of the system from $V_{in,d}$ to $V_{g,d}$ can be written as

$$S_0, V_{in,d}, V_{g,d} = \frac{1.1 \times 10^8 s^4 + 2 \times 10^{16}s^2 + 2 \times 10^{21}}{s^6 + 1.4 \times 10^{-12}s^5 + 3.6 \times 10^{8}s^4 + 4.3 \times 10^{-4}s^3 + 3.2 \times 10^{16}s^2 + 3 \times 10^{4}s + 3.2 \times 10^{21}}$$

The transfer function of the system from $V_{in,q}$ to $V_{g,d}$ can be written as

$$S_0, V_{in,q}, V_{g,d} = \frac{7.2 \times 10^{10}s^3 + 2 \times 10^{-3}s^2 + 7.1 \times 10^{15}s + 0.5}{s^6 + 1.4 \times 10^{-12}s^5 + 3.6 \times 10^{8}s^4 - 1 \times 10^{-3}s^3 + 3.2 \times 10^{16}s^2 - 5.8 \times 10^{4}s + 3.2 \times 10^{21}}$$

The transfer function of the system from $V_{in,q}$ to $V_{g,q}$ can be written as

$$S_0, V_{in,q}, V_{g,d} = \frac{1.1 \times 10^8 s^4 + 1.2 \times 10^{-7}s^3 + 2.1 \times 10^{16}s^2 + 18.5s + 2.1 \times 10^{21}}{s^6 - 4.2 \times 10^{-12}s^5 + 3.6 \times 10^{8}s^4 - 1 \times 10^{-3}s^3 + 3.2 \times 10^{16}s^2 - 5.8 \times 10^{4}s + 3.2 \times 10^{21}}$$

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