An Online Transformerless Uninterruptible Power Supply (UPS) System With a Smaller Battery Bank for Low-Power Applications

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Abstract—Uninterruptible power supplies (UPS) are widely used to provide reliable and high-quality power to critical loads in all grid conditions. This paper proposes a nonisolated online UPS system. The proposed system consists of bridgeless PFC boost rectifier, battery charger/discharger, and an inverter. A new battery charger/discharger has been implemented which ensures the bidirectional flow of power between dc link and battery bank, reducing the battery bank voltage to only 24V, and regulates the dc-link voltage during the battery power mode. Operating batteries in parallel improves the battery performance and resolve the problems related to conventional battery banks that arrange batteries in series. A new control method, integrating slide mode and proportional-resonant control, for the inverter has been proposed which regulates the output voltage for both linear and nonlinear loads. The controller exhibits excellent performance during transients and step changes in load. The operating principle and experimental results of 1-kVA prototype have been presented for validation of the proposed system.

Index Terms—Battery charger/discharger, power factor correction, transformerless uninterruptible power supply (UPS).

I. INTRODUCTION

UNINTERRUPTIBLE power supplies (UPS) provide clean, conditioned, and reliable power to critical loads such as communication systems, network servers, medical equipment’s, etc., in all grid conditions [1], [2]. Typically, the UPS provides unity power factor, high efficiency, high reliability, low cost, and low transients response time from grid mode to battery mode and vice versa [3], [4].

UPS systems can be categorized as online, offline, and line interactive UPS systems [5]. Online UPS systems are most popular and common configuration among them, as it provides isolation to load from the grid and has negligible switching time. A conventional online UPS system consists of a rectifier for PFC, a battery bank, and an inverter connected to the load [6].

Grid frequency transformers are normally employed to reduce the battery bank voltage and provide isolation from the transients and spikes generated inside the grid. Since the transformer is operating at line frequency, thus increasing the size and weight of the system substantially. An online UPS system with high-frequency transformer isolation has been used to overcome the problem related to the grid frequency transformer UPS system [7]. Although the size has been reduced, the efficiency of the system decreases due to high number of active switches in these topologies.

In order to overcome the problems related to aforementioned topologies, the transformerless UPS system has been introduced. Transformerless UPS systems have comparatively high efficiency, small weight and volume of the system. The only disadvantage in the transformerless UPS system is their susceptibility toward the interference caused by the devices connected to the same grid. This makes the transformerless UPS more suitable for environments where the connected grid is less polluted [8].

Several transformerless topologies has been proposed in [9]–[12], focusing on efficiency improvement, volume and weight reduction, decreasing the number of switches, and capital cost of the system. But the size of the battery bank in all the proposed systems so far is enormously high. Generally, the batteries are connected in series to achieve the high battery bank voltage. But series battery arrangement has major drawbacks and limitations in charging and discharging. Small imbalance in voltages occurs across the battery cells during charging and discharging since battery cells are not equal. Hence, these cannot provide the same performance during operation. Overcharging will cause severe overheating, low performance, and even destruction [13]. Similarly, deep discharge may cause the battery cell to be damaged permanently [14]. Due to this reason, a small battery bank with batteries operating in parallel improves the performance of the battery bank significantly. The batteries operating in parallel have following advantages:

1) the number of batteries is not restricted to the dc-link voltage. The volume, weight, and backup time of the battery bank should be designed according to specific application;
2) cost reduction as no extra voltage balancing circuit is required;
3) damaged batteries can be isolated or replaced from the battery bank, thus, leaving the sensitive system operation uninterrupted. This is the prime function of the UPS system;
4) since discharging currents of the batteries can be profiled individually. Hence, the stored energy in the batteries can be utilized more efficiently.

Manuscript received October 14, 2015; revised January 13, 2016; accepted February 18, 2016. Date of publication March 8, 2016; date of current version September 16, 2016. This work was supported by the High Impact Research of University of Malaya—Ministry of Higher Education of Malaysia under Project UMC/HIR/MOE/ENG/24, FRGS-FP014–2014A, and by the Bright Spark Unit. Recommended for publication by Associate Editor D. Xu.

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Digital Object Identifier 10.1109/TPEL.2016.2537834

A UPS topology is proposed in [15] which employs only three-leg converter for both rectification and dc–ac conversion. But the battery bank of 360 V is connected to the dc link, which is enormously high and all the current drawn by the batteries is pulsating. This pulsating current affects the reliability of the battery bank. In [16], a transformerless UPS has been proposed with bidirectional battery charger and discharger connected to the battery bank. But still the battery bank is very high and is not suitable for low-power applications. Nonisolated topology of the UPS system has been proposed in [17]. It reduces the battery bank to nine batteries but it employs an autotransformer to achieve high dc-link voltage that increases the size and weight of the system.

In this paper, a novel transformerless online UPS has been proposed as shown in Fig. 1. The proposed UPS employs a high-gain bidirectional converter which operates between the dc-link voltage and the battery bank. Using bidirectional charger/discharger, the battery bank is reduced to only 24 V (single battery), thus, it eliminates the drawbacks related to large string of series-connected batteries. The bridgeless boost rectifier provides the regulated dc-link voltage to feed the inverter and maintains the power factor correction. A new controller combining slide mode and proportional-resonant (PR) control has been implemented for the inverter control which shows good performance with low total harmonics distortion (THD) and high stability for both nonlinear and impulsive loads. The size and cost of the proposed system is comparatively very low as no bulky transformer has been used, with small battery bank and high efficiency. Hence, the proposed UPS system is excellent choice for low-power application with low cost and weight of the system.

Experimental results based on 1-kVA laboratory prototype have been presented to validate the performance of the system. The proposed UPS system shows excellent steady state and dynamic performance. The advantages of the proposed system are as follows: 1) new battery charger and discharger has been introduced which reduce the size of the battery bank significantly, 2) high input power factor, 3) new robust inverter control scheme for nonlinear and impulse load, and 4) high efficiency and low cost of the system.

II. PROPOSED SYSTEM DESCRIPTION

The schematic of the proposed single-phase online UPS system is shown in Fig. 1. The proposed system consists of a bridgeless PFC boost rectifier, a bidirectional converter, and an H-bridge inverter. The boost rectifier provides the power factor correction and regulated dc-link voltage. The efficiency of the bridgeless rectifier is also high as compared to conventional rectifiers because it eliminates some devices from the power flow path and reduces the conduction losses considerably. Introducing a bidirectional converter for battery charging and discharging with high-voltage gain reduces the size of the battery bank significantly. The H-bridge inverter with new robust control scheme is proposed for regulating the nonlinear load and provides fast transient response during change of modes.

A. Modes of Operation

The operation of the UPS can be divided into two modes of operation. Grid mode and battery mode as shown in the Fig. 2.

**Grid Mode:** When the grid voltage is stable and there is no power failure, the UPS system operates in the grid mode. The rectifier provides the regulated dc-link voltage and feeds the inverter while the bidirectional converter keeps charging the battery bank.

**Battery Mode:** In case of power failure or voltage sag at input, the magnetic contactor (MC) is opened and the rectifier is disabled. Then, the power is supplied to the load by the battery which uses the battery discharger and the inverter. The value of the dc-link capacitor is kept high in order to provide sufficient energy to the inverter during the transition between the battery mode and the grid mode of operation.

A bypass switch has been added in the system to increase the reliability of the system. In case of internal fault in the system...
or overloading and overheating of the circuit, the bypass switch turns ON and provides a direct path for the power from the utility grid to the connected load [2].

**B. Bidirectional Converter**

A new nonisolated bidirectional dc–dc converter with a coupled inductor has been proposed which works as battery charger/discharger and operates between the battery bank and the dc link. The converter has following advantages:

1. high-voltage gain in both the buck and boost mode;
2. less number of passive components in the circuit;
3. only three active switches are used to perform bidirectional operation;
4. zero voltage switching (ZVS), synchronous rectification, and voltage clamping circuit are used that reduce the switching and conduction losses.

A coupled inductor has been used with $L_P$ as primary inductance and $L_S$ as the secondary inductance. The capacitor $C_{b2}$ inserted in the main power across the primary and secondary windings of the transformer gives high-voltage conversion ratio and reduces the peak current stress allowing the continuous current in the primary. Also, the voltage stress of the capacitor $C_{b2}$ is minimum at this position in the circuit.

1. **Battery Charging/Buck Operation:** The characteristic waveforms of the converter during the battery charging mode are shown in Fig. 3. $D_1$ is the duty cycle of $S_3$ and $S_{ax}$, while $D_4$ is the duty cycle of switch $S_4$. Both $D_1$ and $D_3$ are related to each other by a relationship $D_1 (= 1 - D_3)$. $L_m$ is the magnetizing inductance of the coupled inductor with turns ratio $N = N_2 / N_1$, where $N_1$ is the number of turns in primary winding and $N_2$ is the number of turns for secondary winding. The operation of the circuit during the battery charging mode in each interval is shown in Fig. 4.

**Interval 1 ($t_0 \sim t_1$):** The Switch $S_4$ remains ON while the switches $S_3$ and $S_{ax}$ are OFF during interval 1. The current $i_{LS}$ flows from dc link to the battery bank through the capacitor $C_{b2}$ and both the windings of the coupled inductor. Applying KVL, we get

$$V_d = V_{LS} + V_{Cb2} + V_{LP} + V_{Bat}$$  \hspace{1cm} (1)

$$V_d = V_{LP} (1 + N) + V_{Cb2} + V_{Bat}.$$  \hspace{1cm} (2)

The diode $D_{b3}$ is also conducting with the continuous inductor current $i_{Lb}$ into the battery bank. Hence, $V_{Bat}$ is the voltage across the inductor $L_b$.

**Interval 2 ($t_1 \sim t_2$):** At the start, the switch $S_4$ turns OFF. Due to the storage energy in the leakage inductor, the polarities are reversed across the primary and secondary windings ($L_S$ and $L_P$) of the coupled inductor. Switch $S_4$ is OFF in this mode, but the secondary current $i_{LS}$ is still conducting, so the switch $S_{ax}$ body diode is forward bias in order to keep the current $i_{LS}$ flowing. The diode $D_{b3}$ remains forward biased in this mode. The body diode of switch $S_4$ gets forward biased as the secondary current $i_{LS}$ decreases, however, the primary current $i_{LP}$ remains the same.

**Interval 3 ($t_2 \sim t_3$):** Both the switches $S_3$ and $S_{ax}$ turns ON following the ZVS condition. The capacitor $C_{b2}$ starts discharging across the battery bank through the switch $S_{ax}$ and the inductor $L_b$. Thus, the secondary current is induced in reverse by the discharging capacitor $C_{b2}$. The clamp capacitor $C_{b1}$ also discharges through the diode $D_{b2}$ by adding small current $i_b$ into the secondary current flowing into the battery bank.

Using the voltage second balance, $V_{Cb2}$ will be

$$V_{Cb2} = V_{Lb} + V_{Bat} + V_{LS}.$$  \hspace{1cm} (3)

The stored energy in the coupled inductor is released by the primary current through the switch $S_4$ into the battery bank. Using the voltage–second balance, the $V_{Lb}$ is given by

$$D_1 V_{Lb} = D_3 V_{Bat}.$$  \hspace{1cm} (4)

The primary winding voltage $V_{LP}$ can be obtained as

$$D_3 V_{LP} = D_1 V_{Bat}.$$  \hspace{1cm} (5)

Putting (4) and the values of $V_{Lb}$ and $V_{LP}$ into (2), the voltage gain during the buck mode of operation is given by

$$G_{buck} = V_{Bat} / V_d = [D_3 (1 - D_3)] / \left[ 2N (1 - D_3)^2 + 1 \right].$$  \hspace{1cm} (6)

**Interval 4 ($t_3 \sim t_4$):** Both the switches $S_3$ and $S_{ax}$ turn OFF at the start of this mode. The primary and secondary winding currents $i_{LP}$ and $i_{LS}$ will continue conduction due to the leakage inductance of the coupled inductor. The secondary current will charge the parasitic capacitance of the switches $S_3$ and $S_{ax}$, and discharge the parasitic capacitance of the switch $S_1$. When the voltage across the switch $S_{ax}$ equals to $V_{Bat}$, the body diode of the switch $S_4$ get forward biased. The primary current $i_{LP}$ starts decreasing unless it gets equal to the secondary current $i_{LS}$, then this mode finishes.

**Interval 5 ($t_4 \sim t_5$):** The switch $S_4$ turns ON under ZVS condition. The capacitor $C_{b1}$ is charged through the clamped diode $D_{b1}$. The primary and secondary current starts increasing.
At the end of this mode, the circuit starts repeating interval 1 of the next cycle.

2) Battery Discharging/Boost Operation: The characteristic waveform of the bidirectional converter during the battery discharge mode is shown in Fig. 5. The bidirectional converter steps up the low battery bank voltage to the high dc-link voltage. The switch \( S_{ax} \) remains OFF during battery discharging. The battery discharger operation during each interval is shown in Fig. 6.

**Interval 1 \((t_0 \sim t_1)\):** During interval 1, the switch \( S_3 \) is ON, while the switch \( S_4 \) was OFF. The low battery bank voltage is applied at the low-voltage side of the circuit. The capacitor \( C_{b1} \) remains charged before interval 1 and the magnetizing current \( i_{Lm} \) of the coupled inductor increases linearly as shown in Fig. 5. Applying KVL, we get
\[
V_{Bat} = V_{LP} = V_{LS}/N. \tag{7}
\]

The voltage across the primary winding may be derived using voltage second balance
\[
V_{LP} D_3 = V_{Bat} D_1. \tag{8}
\]

**Interval 2 \((t_1 \sim t_2)\):** The switch \( S_3 \) turns OFF in interval 2. The primary current \( i_{LP} \) charges the parasitic capacitance across the switch \( S_3 \) and the secondary current \( i_{LS} \) discharges the parasitic capacitance across switch \( S_4 \). When the voltage across switch \( S_3 \) becomes equal to the capacitor voltage \( V_{C1} \), this interval finishes.

**Interval 3 \((t_2 \sim t_3)\):** Since the switch \( S_3 \) is OFF, the leakage inductance causes the primary current \( i_{LP} \) to decrease while the secondary current \( i_{LS} \) increases. As a result, the body diode of switch \( S_4 \) gets forward biased. The capacitor \( C_{b1} \) starts charging through diode \( D_{b1} \) because the voltage across the switch \( S_3 \) gets higher than the capacitor \( C_{b1} \). This limits the voltage stress across the switch \( S_3 \). The voltage across the \( C_1 \) is given by
\[
V_{C1} = V_{bat} + V_{LP}. \tag{9}
\]

Using (7)
\[
V_{C1} = V_{Bat}/D_3. \tag{10}
\]

**Interval 4 \((t_3 \sim t_4)\):** The switch \( S_3 \) turns ON under the condition of ZVS. The primary and secondary windings of the coupled inductor and the capacitor \( C_{b2} \) are all now connected in series to transfer the energy to the dc link. \( i_{LP} \) starts increasing until it reaches \( i_{LS} \), then, it follows \( i_{LP} \) till the end of the interval 4. Thus, the energy stored in the primary and secondary windings discharges across the dc link. Both the diodes \( D_{b1} \) and \( D_{b2} \) remain reverse biased during this interval as shown in Fig. 6(d). Using voltage second balance, we get
\[
V_d = V_{Bat} + V_{LS} + V_{C2} + V_{LP}. \tag{11}
\]
\[
V_d = V_{Bat} + V_{C2} + (N + 1) V_{LP}. \tag{12}
\]

**Interval 5 \((t_4 \sim t_5)\):** During this interval, the switch \( S_4 \) turns OFF. The current \( i_{LS} \) charges the parasitic capacitance of the
Fig. 6. Topological stages in the boost mode. (a) Mode 1. (b) Mode 2. (c) Mode 3. (d) Mode 4. (e) Mode 5. (f) Mode 6.

Fig. 7. Voltage conversion ratio w.r.t. duty cycle $D_1$ and $D_3$.

By putting (8) and (13) into (12), the voltage gain of the circuit is

$$V_d = V_{bat} + V_{bat}/D_3 + (N + 1)D_1/D_3V_{bat}$$

$$G_{boost} = V_d/V_{bat} = (2 + ND_1)/(1 - D_1).$$

The body diode of the switch $S_3$ starts forward biased because of the polarities of the capacitor $C_{b2}$ and inductor $L_P$.

Interval 6 ($t_5 \sim t_6$): During interval 6, the switch $S_3$ turns ON under the condition of ZVS. Since $S_3$ is not deriving any current from the clamped circuit, thus, the switching losses remain low due to ZVS and the efficiency of the circuit increases. When both $V_{Cb1}$ and $V_{Cb2}$ get equal, then, the next switching cycle starts and repeats the operation in interval 1.

The turn ratio $N$ is selected as such to satisfy $G_{boost}$ and $G_{buck}$ gains for required dc link and battery bank voltage. Fig. 7 shows the voltage gain of buck and boost modes with respect to duty cycle $D_1$ and $D_3$, respectively, at different turn ratio. Turn ratio $N = 4$ satisfies the operation of the bidirectional converter between the required dc link and battery bank.

Table I shows the comparison of different bidirectional converters recently published. The voltage conversion ratio of the proposed converter shows more diversity as compared to [20] and [21], with less number of switches. In [18], the authors have shown high gain ratio, but with five switches, that increases the size and cost of the circuit. The size of the proposed circuit is considerably small with small heat sink for the given power rating, and only few passive auxiliary components are used. Since the battery voltage is very low and high current flows from the battery bank into the converter. Thus, it increases conduction.
losses. However, the switching losses are not significant as all the switches of the bidirectional dc–dc converter are following ZVS condition. The high current can increase the size and cost of the system, hence limits the operation of proposed topology for very high-power applications where the input current can be very high.

C. Rectifier

The rectifier performing the unity power factor consists of the bridgeless PFC boost rectifier. The bridgeless PFC boost rectifier does not use the full-wave bridge rectifier, reducing one semiconductor device in the main current path. Thus, the conduction losses are reduced, which increases the efficiency of the rectifier. The bridgeless PFC has the advantage of reducing the conduction loss by 30% [22]. This topology is suitable for applications where high power density and high efficiency are required. The bridgeless rectifier consists of two boost converters, each operating in the half cycle of the ac supply. By adding two slow diodes $D_a \sim D_b$, the common mode noise (EMI Losses) can be suppressed considerably and high efficiency can be achieved as compared to the conventional rectifier. Both the switches $S_1$ and $S_2$ of the rectifier are driven by the same gate signal, thus makes the control of the circuit quite easy. The inductors $L_{11}$ and $L_{12}$ of the boost rectifier can be wound in the same core in order to increase the utilization of the magnetic material [23].

The operation of the rectifier during only positive half cycle has been shown in Fig. 8. The switch $S_1$ turns ON, as the input voltage $V_{in}$ (ac) turns positive. The current flows from the input through the inductor $L_{11}$ and $L_{12}$, storing the energy in both the inductors. The change in the input current $i_{in}$ is same as the change in the inductor current, given by

$$\Delta i_{in} = \frac{1}{L_{11} + L_{12}} V_{in} DT_s. \quad (16)$$

When the switch $S_1$ turns OFF, the energy is released by the inductors. The current flows through the diode $D_1$ into the dc link $V_d$, returning through the body diode of the switch $S_2$ into the input supply. The input current in $D_1$ and $S_2$ is same as the inductor current given by

$$\Delta i_{in} = \frac{1}{L_{11} + L_{12}} (V_{in} - V_d)(1 - D)T_s. \quad (17)$$
Depending on the duty cycle $D$ of both the switches $S_1$ and $S_2$, the input current variation for one complete switching cycle $T_s$ is given by

$$\left(L_{11} + L_{12}\right)\frac{\Delta i_{in}}{T_s} = V_{in} DT_s + (V_{in} - V_d) (1 - D) T_s. \quad (18)$$

The EMI noise is generated due to high-frequency switching that is leaked through the parasitic capacitance from the converter to the ground. Therefore, the EMI noise suppression diodes $D_a$ and $D_b$ are used, as they provide the conducting path between the output bus and the input line during both the positive and negative cycles, thus, the voltage potential of the output bus is stabilized [24].

### III. Control Strategy

The control scheme for the inverter keeps operating in both the grid and battery mode. For the rectifier, the control scheme operates only in the grid mode, while the battery charger and discharger also switch during the change of modes. The control schemes for controlling different parts of the UPS in different modes of operation are shown in Fig. 9.

#### A. Inverter Control

A conventional full-bridge voltage source inverter has been used to perform dc to ac conversion. Different high performance control schemes have been presented for the inverter. Among these control schemes, deadbeat control [25] is the most popular control technique because of its fast dynamic response, as the tracking error settles to zero in finite sampling steps. However, the deadbeat control is very sensitive to model uncertainties, parameters mismatch, and noise in the high sampling frequency. The repetitive controller (RC) [26] also provides good regulation for the nonlinear loads with periodic distortions and excellent harmonics rejection. However, the dynamic response of the RC is relatively very slow, that is why another fast response control scheme is usually integrated with the RC to increase the response time of the controller. Moreover, poor tracking accuracy and large memory requirement are additional limitations in the RC controller. Model predictive control [27] predicts the behavior of the output voltage for each switching state at every sampling interval. A cost function is derived to predict the next switching state. Though this control imposes very small computational burden, it requires very high sampling rate and does not provide any analysis of stability and robustness.

For the nonlinear load, the slide mode control (SMC) strategy has gained special interest because of its effective performance and high-frequency switching control in power inverters against nonlinear system with uncertainties. A major feature of the SMC is its robustness, good dynamic response, stability against nonlinear loading conditions, and easy implementation. SMC provides good regulation in a wide range of operating conditions. The integral SMC method has been proposed for efficient ac tracking of the system in [28]. Though this system has reduced the harmonics contents in the output voltage but offers limited ability for high-order harmonics. SMC with the continuous time control method has been implemented in [29]. The hysteresis-type switching function has been introduced for each leg of the inverter which increases the hardware complexity. In rotating SMC [30], the time-varying slope based on the SMC method was proposed which rotates the sliding surface in order to get the faster response for the nonlinear conditions. This different value for the slope has been applied during the transient and steady-state operation, causing the surface to rotate according to load variation. Multiresonant SMC has been implemented for the grid-connected inverters [31]. It relatively reduces the tracking error and THD of the grid current. But this controller is preferred for the grid-connected inverter with low THD, and also, the system parameters determine the reference signal which may degrade the robustness of the controller.

In order to control the output voltage of the inverter, a cascaded control algorithm of SMC and PR control has been analyzed for the proposed UPS topology. It is a new control scheme for the single-phase bipolar voltage source inverter of the UPS system. The inner current loop is controlled by the SMC while the outer voltage loop is controlled by the PR control. The chattering phenomenon in the SMC is eliminated by using the smoothed control law in the narrow boundary layer. The smoothed control law applied to the pulse width modulator results in a fixed switching frequency operation of the inverter. Thus, the proposed controller adopted the characteristics of both SMC and PR control. The controller shows good response with low THD and high stability for nonlinear loads. The main advantages of the proposed controller are as follows:

1) very low THD for both linear and nonlinear load;
2) very robust in operation;
3) fast transient response;
4) easy implementation.

The circuit diagram of the single-phase inverter with the LC filter and proposed controller for the nonlinear load is shown in Fig. 10, where $V_d$ is the applied dc-link voltage, $V_{out}$ is the filter capacitor, and $C_f$ is the output voltage. $i_{L_f}$ is the inductor, $L_f$ is the current, and $i_o$ is the output current through the load $R$, given by $i_0 = V_{out}/R$. The state equations of the inverter are

![Block diagram of inverter control](image-url)
given as
\[
\frac{d}{dt} \begin{bmatrix} V_{out} \\ i_{Lf} \end{bmatrix} = \begin{bmatrix} 0 & 1/C_f \\ -1/L_f & 0 \end{bmatrix} \begin{bmatrix} V_{out} \\ i_{Lf} \end{bmatrix} + \begin{bmatrix} 0 \\ V_d/L_f \end{bmatrix} u + \begin{bmatrix} -i_o/C_f \\ 0 \end{bmatrix}
\]

where \( u = \text{Control input} = \{-1, 0, +1\}. \)

In order to implement the sliding mode control, the voltage error \( x_1 \), and its derivative \( x_2 = \dot{x}_1 \) need to be find
\[
x_1 = V_{out} - V_{ref}
\]
\[
x_2 = \dot{x}_1 = V_{out} - \dot{V}_{ref} = i_{Cf}/C_f - \dot{V}_{ref},
\]
where \( V_{ref} = V_m \sin(\omega t) \)
\[
\begin{bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{bmatrix} = \begin{bmatrix} 0 & 1/L_f C_f \\ -1/L_f C_f & -1/R C_f \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \begin{bmatrix} 0 \\ V_d/L_f C_f \end{bmatrix} u + \begin{bmatrix} -V_{ref}/L_f C_f \end{bmatrix}.
\]

Consider the slide surface equation
\[
S = \lambda x_1 + x_2.
\]

In order to ensure the stability of the sliding function, the Lyapunov function \( V(t) = S^2/2 \) has to be satisfied with the minimum condition \( \dot{V}(t) < \eta |s| \), keeping the scalar \( s \) at zero while \( \eta \) is strictly positive constant. Hence, the condition for stability will be \( V(t) < 0 \)
\[
\dot{V}(t) = SS
\]
\[
V(t) = S \left[ \lambda x_2 - \frac{1}{L_f C_f} x_1 + \frac{V_{DC}}{L_f C_f} u - \frac{V_{ref}}{L_f C_f} - \frac{x_2}{R C_f} \right] u.
\]

In order to satisfy the sliding condition (22), despite of the uncertainty on the dynamics of the nonlinear function, \( u \) is replaced by the “–sign(s)” function
\[
u(t) = -\text{sign}(s) = \begin{cases} +1, & \text{if } S(x) > 0 \\ -1, & \text{if } S(x) < 0 \end{cases}
\]
\[
\dot{V} = \left| S \left[ \lambda x_2 - \frac{1}{L_f C_f} x_1 - \frac{V_{DC}}{L_f C_f} \text{sign}(s) \right] - \frac{V_{ref}}{L_f C_f} - \frac{x_2}{R C_f} \right|
\]
\[
\dot{V} = |S| \left[ \text{sign}(s) \left[ \lambda x_2 - \frac{1}{L_f C_f} x_1 - \frac{V_{ref}}{L_f C_f} - \frac{x_2}{R C_f} \right] \right]
\]
\[
\text{sign}(s) \left[ \lambda x_2 - \frac{1}{L_f C_f} x_1 - \frac{V_{ref}}{L_f C_f} - \frac{x_2}{R C_f} \right] < \frac{V_{DC}}{L_f C_f}.
\]

Hence, it is clear that the stability condition is fulfilled when (28) is satisfied. Now to apply the sliding control law to the inverter, put the value of \( x_1 \) and \( x_2 \)
\[
S = \lambda (V_{out} - V_{ref}) + \frac{i_{Cf}}{C_f} - \dot{V}_{ref}
\]
\[
S = \lambda (V_{out} - V_{ref}) + \frac{1}{C_f} (i_{Cf} - i_{ref}).
\]

Since the sliding mode controller has the common inherent property of chattering phenomena, it causes low control accuracy and high losses in the circuit. In order to overcome the chattering phenomena, a smoothed SMC has been implemented. This can be achieved by smoothing out the control discontinuity in a thin boundary layer neighboring the sliding surface
\[
B(t) = \{ x, |S(x;t)| \leq \phi \} \phi > 0,
\]
where \( \phi \) is the boundary layer thickness and \( \varepsilon = \frac{\phi}{2} \) is the boundary layer width. Hence, \( B(t) \) is chosen as such that all the trajectories starting at \( B(t = 0) \) remain inside \( B(t) \) for all \( t > 0 \) as shown in Fig. 11(a). Hence, we interpolate \( S \) inside \( B(t) \) for instance, and replace \( S \) by an expression \( S/\phi \). Thus, (30) will be
\[
\frac{S(x)}{\phi} = \frac{\lambda}{\phi} (V_{out} - V_{ref}) + \frac{1}{C_f \phi} (i_{Cf} - i_{ref}).
\]

The smoothing control discontinuity assigns a low-pass filter structure to the local dynamics thus eliminates chattering. The smoothed control law applied to the pulse width modulator results in the fixed switching frequency of the inverter. The control law needs to be tuned very precisely in order to achieve a tradeoff between the tracking precision and robustness to the uncontrolled dynamics as shown in Fig. 11(b).

Conventionally, the PR controller provides a large gain at the fundamental frequency and strictly follows the sinusoidal reference, reducing the steady-state error and improving the stability of the system. The transfer function of the ideal PR controller is given by
\[
G_{PR} = K_P + \frac{2K_R s}{s^2 + \omega_0^2}
\]
where \( K_P \) is the proportional gain, \( \omega_0 \) is the resonant frequency, and \( K_R \) is the resonant gain.
The ideal PR controller gives the infinite gain at the resonant frequency but no gain and phase shift at other frequencies. Hence, more appropriate is nonideal PR control, given as

\[
G_{PR} = K_P + \frac{2K_R\omega_c s}{s^2 + 2\omega_c s + \omega_c^2}, \quad (34)
\]

Hence, selecting a suitable cutoff frequency \( \omega_c \) can widen the bandwidth, reducing the sensitivity toward the frequency variations. By combining the PR controller with the SMC, the performance of the inverter is improved, as the resonance controller provides better regulation of the output voltage and reduces the total harmonic distortion considerable.

Hence, the final equation for the control of the inverter can be derived by combining the PR control and SMC for the current loop

\[
\frac{S(x)}{\phi} = \frac{1}{C_f \phi} \left[ i_{C_f} - i_{ref} \right] + \frac{\lambda}{\phi} \left[ K_p (V_{out} - V_{ref}) \right] + k_i \left( \frac{2s}{s^2 + 2\omega_c s + \omega_c^2} \right) (V_{out} - V_{ref}) \right]. \quad (35)
\]

Thus, (35) shows the dynamic behavior of the system with both SMC and PR compensator. The error in the voltage loop is compensated by the appropriate PR parameters, thus, the output voltage is compelled to follow the reference ac voltage leading to the system stability while the SMC drives the system to the zero sliding surface with maximum stability. Since the capacitor error current contains the ripples from the inductor, the current peak may reach high values. So, \( \phi \) should be carefully assigned value in order to compensate the slope from the high-current ripple of the capacitor. Hence, the PR controller eliminates the steady-state error at resonant frequency or harmonic at that frequency.

The response time of the system \( \lambda \) determines the dynamics and robustness of the system. It is clear from (35), that smaller value of \( \lambda \) leads to slow response time, while higher \( \lambda \) values though increase the response time but take larger time to reach the sliding surface. Thus, the optimal value for \( \lambda \) is equal to the switching frequency of the inverter.

According to [32], the slope of the carrier wave is given as \( 4V_m \times f_s \), where \( V_m \) is the magnitude and \( f_s \) is the frequency of the carrier wave. The slope of the error signal to the modulator is given by \( V_{DC}/4LC\phi \). According to the limitation of the pulse width modulator

slope of error signal < slope of the carrier signal

\[
4V_m \times f_s < V_d/4LC\phi. \quad (36)
\]

Thus, the minimum value of \( \phi \) can be calculated by

\[
\phi \simeq \frac{10V_d}{16LCf_vm f_s}. \quad (37)
\]

In order to design the controller for the inverter as shown in Fig. 10, the value of \( \phi \) can be derived using (37) considering the circuit parameters from Table VIII. \( K_P \) and \( K_R \) are the proportional gain and resonant gain selected for stable response of the PR controller. The value of \( \alpha \) is the division factor to bring the output voltage compatible to reference and is selected considering the electronic circuit limitations. \( \lambda \) is the dynamic response of the inverter and it is equal to the switching frequency of the modulator. The magnitude of the carrier \( V_m \) is selected to realize the inequality (36). Final control parameters have been derived for the stable operation of the inverter and are presented in Table II.

Table III shows the comparison of the proposed control scheme with the SMC and other common controllers. The proposed controller shows an improvement in terms of reducing the THD and transient response with robust control of the inverter.

### B. Rectifier Control

The rectifier of the UPS system is controlled by well-known average current mode control as shown in Fig. 9. In this control scheme, the faster inner current loop regulates the inductor current so that its average value during each period follows the rectified input voltage. The slower outer voltage loop maintains the rectified output voltage close to the reference voltage and generates the control signal \( v_i \) for the current loop. The steady-state analysis of the rectifier shows stable performance during the grid mode. The state-space equations of the rectifier are derived as

\[
\frac{di_L}{dt} = \frac{V_{in}}{L_{11} + L_{12}} D + \frac{(V_{in} - V_d)}{L_{11} + L_{12}} (1 - D) \quad (38)
\]

\[
\frac{dv_d}{dt} = -\frac{V_d}{RC_d} D + \frac{i_L}{C_d} - \frac{V_d}{RC_d} (1 - D). \quad (39)
\]

Assuming that the current loop has high bandwidth as compared to voltage loop, and the output capacitor \( C_d \) is large enough to give approximately constant output voltage, i.e., \( dv_d/dt = 0 \). With \( V_{in} = 0 \), the small signal control \( \delta \) to input current \( i_L \) transfer function \( G_{i_L\delta}(s) \) of the inner current loop is
TABLE IV
BATTERY SPECIFICATIONS

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated Capacity</td>
<td>35 Ah</td>
</tr>
<tr>
<td>Nominal Voltage</td>
<td>24 V</td>
</tr>
<tr>
<td>Min. Voltage</td>
<td>16 V</td>
</tr>
<tr>
<td>Max. Charging Current limit</td>
<td>9.9 A</td>
</tr>
<tr>
<td>Max. Discharge Current</td>
<td>105 A</td>
</tr>
<tr>
<td>Initial SoC</td>
<td>70%</td>
</tr>
<tr>
<td>Internal Resistance</td>
<td>8 mΩ</td>
</tr>
</tbody>
</table>

TABLE V
SPECIFICATIONS OF THE PROPOSED UPS SYSTEM

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input voltage</td>
<td>( V_{in} )</td>
</tr>
<tr>
<td>Output voltage</td>
<td>( V_{out} )</td>
</tr>
<tr>
<td>Grid frequency</td>
<td>( f_g )</td>
</tr>
<tr>
<td>Output frequency</td>
<td>( f_o )</td>
</tr>
<tr>
<td>Number of batteries</td>
<td>( V_b )</td>
</tr>
<tr>
<td>Maximum output power</td>
<td>( P_{max} )</td>
</tr>
<tr>
<td>DC-link voltage</td>
<td>( V_d )</td>
</tr>
</tbody>
</table>

TABLE VI
DESIGN PARAMETERS OF THE RECTIFIER

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Symbol</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Inductor</td>
<td>( L_1, L_2 )</td>
<td>( 800 ) μH</td>
</tr>
<tr>
<td>Diodes</td>
<td>( D_1, D_2 )</td>
<td>BYC10–600</td>
</tr>
<tr>
<td>Switches</td>
<td>( S_1, S_2 )</td>
<td>SPP11N60C3</td>
</tr>
<tr>
<td>Slow Diodes</td>
<td>( D_{s1}, D_{s2} )</td>
<td>GBH1508</td>
</tr>
<tr>
<td>Switching frequency</td>
<td>( f_s )</td>
<td>30 000 Hz</td>
</tr>
</tbody>
</table>

TABLE VII
SPECIFICATION OF THE BATTERY CHARGER/DISCHARGER

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Symbol</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC-link voltage</td>
<td>( V_d )</td>
<td>360 V</td>
</tr>
<tr>
<td>Battery bank voltage</td>
<td>( V_b )</td>
<td>24 V</td>
</tr>
<tr>
<td>Switching frequency</td>
<td>( f_s )</td>
<td>30 000 Hz</td>
</tr>
<tr>
<td>Coupled inductor</td>
<td>( L_P, L_S )</td>
<td>Turns ratio ( N = 4; L_m = 107 ) μH; PQ-5050 core;</td>
</tr>
<tr>
<td>Inductor</td>
<td>( L_0 )</td>
<td>300 μH</td>
</tr>
<tr>
<td>Capacitor</td>
<td>( C_{11, 12, 2d} )</td>
<td>( C_{11, 12} = 2 \times 2 ) μF ceramic, ( C_d = 1900 ) μF</td>
</tr>
<tr>
<td>Slow Diodes</td>
<td>( D_{s1}, D_{s2}, D_{s3} )</td>
<td>IPW60R045CP MOSFET</td>
</tr>
<tr>
<td>Diodes</td>
<td>( D_{s1}, D_{s2}, D_{s3} )</td>
<td>Ultrafast Recovery diode UF5408</td>
</tr>
</tbody>
</table>

The Bode response of the rectifier. (a) Current loop gain. (b) Voltage loop gain.

\[
G_{i.d}(s) = \frac{V_d}{s(L_{11} + L_{12})}. \tag{40}
\]

The stability of the current loop depends on the current loop gain, hence, suitable proportional-integral (PI) controller, \( G_i(s) = k_{pi} + \frac{k_{ii}}{s} \), is used for compensating the current loop. The Bode plot of the current loop gain \( T_i = G_{i.d}(s)G_i(s) \) is obtained considering the circuit parameters shown in Table IV. The value of proportional gain \( K_{pi} \) and integral gain \( K_{ii} \) is selected as 2.3 and 1200, respectively, for the stable operation of the current loop. Fig. 12(a) presents the bode plot of the current loop gain with phase margin of 89° and stable operation of the rectifier. Same approach is used to compensate the voltage loop of the average current control scheme. \( \hat{v}_d \) is the reference current for the current loop. Assuming the constant input voltage, the small signal control \( \hat{v}_c \) to output transfer function \( G_{VdVc}(s) \) of the voltage loop is derived as

\[
G_{VdVc}(s) = \frac{\hat{v}_d}{\hat{v}_c} = \frac{V_{in}R}{2V_d(sCR + 2)}. \tag{41}
\]

In order to force the output voltage to follow the reference voltage \( V_{ref} \), a PI compensator has been employed. Combining the power stage with the PI controller \( G_c(s) = k_{pv} + \frac{k_{pv}}{s} \) provides the overall loop gain \( T_c = G_cG_{VdVc}(s) \) of the voltage loop. The value of \( K_{pv} \) and \( K_{iv} \) in voltage loop is selected as 1.2 and 13, respectively. The stability of the voltage loop can be analyzed using the Bode plot obtained by considering the parameters from Table VI, as shown in Fig. 12(b). The system shows good stability with positive phase margin.

C. Battery Charger and Discharger Control

The controller for the battery charger/discharger during both grid mode and battery mode has been shown in Fig. 9. During battery charging, the controller operates as constant current (CC) mode or constant voltage (CV) mode depending on the battery voltage, while in battery discharging, the controller regulates the dc-link voltage as well as the primary inductor current. It is
assumed that the primary inductor current $i_{LP}$ flows continuously. The steady-state analysis of the battery charger/discharger is performed using the average state variable method [35]. The state-space equations for the charger with the coupled inductor are

$$\frac{di_{LP}}{dt} = \frac{V_{Bat}}{L_{m}(N+1)}d + \frac{(V_{Bat} - V_d)(1 - D)}{L_{m}(N+1)} \tag{42}$$

$$\frac{dV_o}{dt} = \frac{i_p}{C(N+1)}(1 - D) - \frac{V_o}{RC} \tag{43}$$

Solving state-space equations gives the primary inductor to control transfer function $G_{i_{LP}d}(s)$ and output to control transfer function $G_{v_{o}d}(s)$ of the battery charger/discharger as

$$G_{i_{LP}d}(s) = \frac{\hat{i}_{LP}}{d} = \frac{NV_{Bat} + v_d}{(N+1)RCL_{m}} \times \left(\frac{s + \frac{i_p}{RC}}{s^2 + s \frac{1}{RC} + \frac{(1-D)^2}{(1+N)^2L_{m}C}}\right) \tag{44}$$

$$G_{v_{o}d}(s) = \frac{\hat{v}_{o}}{d} = \frac{\frac{-i_p}{C(N+1)}}{s^2 + s \frac{1}{RC} + \frac{(1-D)^2}{(1+N)^2L_{m}C}} \left(\frac{NV_{Bat} - v_d/L_{m}(N+1)}{(1-D)}\right) \tag{45}$$

The right-half plane zero in the control to output transfer function has been placed properly with suitable selecting the design components. Using current mode control and selecting optimum value of $L_{M}$, load current, and duty cycle $D$ of the converter, keeps the circuit operation under stable condition [36], [37].

Considering the gain due to the clamp capacitor $C_{b2}$, the transfer equation is given by

$$G_{v_{o}d}(s) = \frac{\hat{v}_{o}}{d} = \frac{s \left(\frac{-i_p}{C(N+1)}\right)}{s^2 + s \frac{1}{RC} + \frac{(1-D)^2}{(1+N)^2L_{m}C}} + \frac{(1-D)}{C(N+1)} \left(\frac{NV_{Bat} - v_d/L_{m}(N+1)}{(1-D)}\right) \tag{46}$$

The right-half plane zero in the control to output transfer function has been placed properly with suitable selecting the design components. Using current mode control and selecting optimum value of $L_{M}$, load current, and duty cycle $D$ of the converter, keeps the circuit operation under stable condition [36], [37].

In battery discharging control, the voltage loop with the PI compensator $G_v = k_p + k_i/s$ regulates the dc-link voltage $V_d$
and provides the reference current $i_{\text{ref}}$ for the current loop. Similarly, the PI compensator is added in the current loop to force the primary inductor current $i_P$ to follow the reference current $i_{\text{ref}}$ from the voltage loop. The Bode plot of the current loop gain and voltage loop gain has been generated considering the battery charger/discharger circuit parameters from Table VII. The values of $k_p$ and $k_i$ are 1.7 and 9, respectively, for the voltage loop, while 2.3 and 2300, respectively, for the current loop. The system shows good stability with the positive phase margin and has no right-half plane poles as shown in Fig. 13. It is easy to achieve higher cross over frequencies by adjusting a suitable gain of the compensators as the phase never reaches to –180.

For designing the battery charging controller, the equivalent electric circuit of the battery is presented in Fig. 14. The Thevenin battery model is most commonly used model [38], which consist of an ideal battery voltage $E_0$, internal resistance $R_i$, polarization capacitor $C_P$, and polarization resistance $R_P$. All the elements used in the model are functions of battery state of charge (SoC). $N_S$ and $N_P$ are the number of cells in series and parallel, respectively. The battery terminal voltage $V_{\text{Bat}}$ can be presented as

$$V_{\text{Bat}} = N_s (E_0 - I_{\text{Bat}}R_i - V_{\text{CP}}),$$

(47)

where $V_{\text{CP}}$ is the polarization voltage and $I_{\text{Bat}} = \frac{I_{\text{Load}}}{N_p}$ is the battery current. Model parameters have been identified for the battery as shown in Table IV.

In the charging mode, the controller operates as CC mode or CV mode depending on the battery voltage as shown in the Fig. 9. In the current loop, the battery input current $i_{\text{Bat}}$ is forced...
to follow the reference current $i_{\text{ref}}$ using the PI compensator in

$$i^* = K_p (i_{\text{ref}} - i_{\text{Bat}}) + K_i \int (i_{\text{ref}} - i_{\text{Bat}}) dt. \quad (48)$$

Similarly, the battery voltage is regulated by the voltage loop using the PI compensator that forces the output battery voltage $V_{\text{Bat}}$ to follow the reference voltage $V_{\text{ref}}$. The current limiter is introduced to limit the maximum charging current of the battery as specified in Table IV. If $i_{\text{ref}}$ is greater than $i_{\text{limit}}$, the battery is charged at the CC mode, in contrast if $i_{\text{ref}}$ is less than $i_{\text{limit}}$, the battery is charged at the CV mode.

IV. EXPERIMENTAL RESULTS

To verify the performance of the proposed UPS system, a laboratory prototype has been implemented with the specifications shown in Table V. The control scheme for inverter, rectifier, and battery charger/discharger has been implemented using DSP TMS320F28335. The design parameters of the rectifier, battery charger/discharger, and inverter are shown in Tables VI, VII, and VIII, respectively. The backup storage system consists of two batteries (each battery is 24 V/35 Ah), or parallel batteries depending upon the backup time for the connected load.

The utility input voltage and the current waveform in the grid mode of operation is shown in the Fig. 15. The input current waveform is very close to the sinusoidal and has almost unity power factor with THD 4.5%. Fig. 16 and 17 show drain to source voltage of the switches $S_3$ and $S_4$ of the bidirectional converter (battery charger) during both the buck and boost mode.
TABLE IX

<table>
<thead>
<tr>
<th>UPS Topology</th>
<th>Efficiency</th>
<th>Power Ratings</th>
<th>System Specification</th>
<th>Battery Bank</th>
<th>Size and Weight</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transformerless offline UPS system [39]</td>
<td>High</td>
<td>1 kVA</td>
<td>220 V</td>
<td>144 V</td>
<td>Medium</td>
</tr>
<tr>
<td>A reconfigurable UPS for multiple power quality [15]</td>
<td>High</td>
<td>1 kVA</td>
<td>110 V</td>
<td>300 V</td>
<td>-</td>
</tr>
<tr>
<td>Transformerless online UPS system [16]</td>
<td>96%</td>
<td>3 kVA</td>
<td>220 V</td>
<td>192 V</td>
<td>Smaller</td>
</tr>
<tr>
<td>Nonisolated UPS with 110/220V input-output voltage [17]</td>
<td>86%</td>
<td>2.6 kVA</td>
<td>110 V and 220 V</td>
<td>108 V</td>
<td>Medium</td>
</tr>
<tr>
<td>Z-source inverter based UPS system [40]</td>
<td>&gt; 90%</td>
<td>3 kVA</td>
<td>220 V</td>
<td>360 V</td>
<td>Smaller</td>
</tr>
<tr>
<td>Proposed UPS system</td>
<td>92%</td>
<td>1 kVA</td>
<td>220 V</td>
<td>24 V</td>
<td>Smallest</td>
</tr>
</tbody>
</table>

of operation, respectively. Both the switches are operating under the condition of ZVS. Similarly, the output voltage and current of the bidirectional converter during battery charging has been shown in Fig. 18. The output voltage and the current waveform during the linear load are shown in Fig. 19. The waveform is sinusoidal with THD less than 1%. Also, the system is connected with the nonlinear load designed according to the standard of IEC62040–3. The system shows good performance with THD of 1.25% for the nonlinear load as shown in the Fig. 20.

Figs. 21 and 22 show the output voltage and the current waveform during step change in load from 0 to 100% and from 100% to 0. The system shows good response to step changes and provides the regulated output voltage regardless of the load changes. When the grid power is interrupted and the system switches from grid mode to battery mode, the rectifier is no more in operation and the battery charger/discharger operates in discharging mode giving the regulated dc-link voltage. The transient effect in the output voltage is very small and the UPS system provides uninterruptible power to the load as shown in Fig. 23. Similarly, the transition from battery mode back to grid mode upon the restoration of the grid power is shown in Fig. 24. The waveform is 110/220 V input–output voltage [17] and reduces the battery bank significantly. A new control for the inverter provides the regulated sinusoidal output voltage with low THD for both linear and nonlinear load. Overall, the volume of the system is minimized by reducing the size, weight, and battery bank of the system. The experimental results show good dynamic and steady-state performance. It may be recommended to extend this proposed UPS system to the three-phase transformerless online UPS system.

A single-phase transformerless online UPS has been proposed in this paper. A bridgeless boost rectifier has been used with the average current control method that increases the efficiency of the system and provides the power factor correction. A new bidirectional converter for battery charging/discharging has been implemented which ensures transformerless operation and reduces the battery bank significantly. A new control for the inverter provides the regulated sinusoidal output voltage with low THD for both linear and nonlinear load. Overall, the volume of the system is minimized by reducing the size, weight, and battery bank of the system. The experimental results show good dynamic and steady-state performance. It may be recommended to extend this proposed UPS system to the three-phase transformerless online UPS system.

REFERENCES