Control of Second Life Hybrid Battery Energy Storage System Based on Modular Boost-Multilevel Buck Converter

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Abstract—To fully utilise second life batteries on the grid system a hybrid battery scheme needs to be considered for several reasons; the uncertainty over using a single source supply chain for second life batteries, the differences in evolving battery chemistry and battery configuration by different suppliers to strive for greater power levels and the uncertainty of degradation within a second life battery. Therefore, these hybrid battery systems could have widely different module voltage, capacity, initial SOC and state-of-health (SOH). In order to suitably integrate and control these widely different batteries, a suitable multi-modular converter topology and associated control structure are required. This paper addresses these issues proposing a modular boost-multilevel buck converter based topology to integrate these hybrid second life batteries to a grid-tie inverter. Thereafter, a suitable module based distributed control architecture is introduced to independently utilise each converter module according to its characteristics. The proposed converter and control architecture are found to be flexible enough to integrate widely different batteries to an inverter dc-link. Modelling, analysis and experimental validation are performed on a single phase modular hybrid battery energy storage system prototype to understand the operation of the control strategy with different hybrid battery configurations.

Index Terms— Second life hybrid batteries, modular boost-multilevel buck converter, distributed control, boost-buck mode

 NOMENCLATURE

\[ V_{\text{batt},i} \quad \text{Battery terminal voltage of } i^{th} \text{ module} \quad \text{(V)} \]
\[ i_{\text{batt},i} \quad \text{Instantaneous battery current of } i^{th} \text{ module} \quad \text{(A)} \]
\[ I_{\text{batt},i} \quad \text{Average battery current of } i^{th} \text{ module} \quad \text{(A)} \]
\[ \alpha_i \quad \text{Desired weighting factor for } i^{th} \text{ module current} \quad \text{(A)} \]
\[ i_{\text{dc},i} \quad \text{Instantaneous dc-link current of } i^{th} \text{ module} \quad \text{(A)} \]
\[ I_{\text{dc},i} \quad \text{Average dc-link current of } i^{th} \text{ module} \quad \text{(A)} \]
\[ V_{\text{dc},i} \quad \text{Instantaneous dc-link capacitor voltage} \quad \text{(V)} \]
\[ V_{\text{dc}} \quad \text{Desired central DC-link capacitor voltage} \quad \text{(V)} \]
\[ I_{\text{dc}} \quad \text{Average DC-link current} \quad \text{(A)} \]
\[ i_{\text{dc}} \quad \text{Instantaneous value of DC-link current} \quad \text{(A)} \]
\[ d_i \quad \text{Instantaneous duty cycle of switches } S_i, S_{ii} \quad \text{(A)} \]
\[ d_{\text{av}} \quad \text{Overall duty cycle of multilevel buck converter} \quad \text{(A)} \]
\[ d_{\text{av}} \quad \text{Instantaneous duty cycle of switches } T_{i}, T_{ii} \quad \text{(A)} \]
\[ D_i \quad \text{Average duty cycle of switches } S_i, S_{ii} \quad \text{(A)} \]
\[ D_{\text{av}} \quad \text{Average duty cycle of switches } T_{i}, T_{ii} \quad \text{(A)} \]

Significant research has been carried out on battery energy storage systems (BESS) using a single type of battery system where a single dc-dc with a dc-ac or a direct dc-ac converter interface with the battery bank is used both in grid-tie and in microgrid applications [1] – [4]. These systems predominantly use new batteries where differences between the cells are fairly minimal. Therefore, active or passive balancing circuit is typically employed to overcome any imbalances (in terms of voltage or state-of-charge) among the cells [5] – [6]. This paper is concerned with hybrid second life battery systems e.g. re-using EV/HEV batteries in grid support applications because there is a significant interest in using these transportation batteries to help support the new smart grid functionalities [7] – [8]. The main advantage of these batteries are the supposed lower cost compared to new batteries and the chance to delay the development of the second life battery recycling chain which is in its infancy due to changing battery chemistry and the impact on the recycling process cost. These second life batteries are likely to trickle through the battery recyclers (at a module level) and therefore to get a sufficiently large system for grid support will require the use of different manufacturer’s batteries. Each battery in the system could have a different chemistry (e.g. Li-ion, lead-acid, NiMH etc.), voltage, capacity, initial state of charge (SOC) and state-of-health (SOH). As a battery fails it would be desirable to hot-swap it for one that works resulting in a hybrid mix encompassing everything from new batteries to batteries close to failure at any moment in time.

To meet the requirements for integrating together hybrid energy storage systems, multi-modular power converters (cascaded/parallel) are preferred [9] – [10]. Previous research on such hybrid energy systems with batteries has mainly focused on generation-storage hybridization for example; batteries with super-capacitor [11], batteries with PV/wind or fuel cells [12] – [14]. These researches also use the same type of battery system and focus on power sharing between the battery and other sources to increase battery useful life or...
smoothing the mismatch between the generation and demand.
In some cases, multi-modular converters are also used for power sources of the same type but under different operating conditions. For example, they can be used for PV systems under partial shading [15] – [17] using multi-modular dc-dc converters with a single inverter to deal with the heterogeneous nature of PV panels under MPPT conditions. Several authors have compared and recommended cascaded dc-dc converters over a parallel structure due to increased efficiency, reduced size and cost [15], [16]. In that research, boost type modules were preferred for low module voltages where the total dc-side voltage was less than the inverter dc-link voltage, whereas buck type modules were suggested for the case of higher module voltages. From a control point of view, a cascaded boost converter neither can deliver the peak power for a module under all inhomogeneous radiation conditions, nor can it provide the fault-tolerance without extra protection, whereas a cascaded buck structure is capable of handling both situations. In order to address the shortfall of a cascaded boost structure, a separate string inverter was proposed in [17] which could solve power mismatch problems but not the problem of integrating widely uneven module voltages together within a grid-tie inverter. Therefore, these three problems: a) integrating widely different voltage modules to an inverter dc-link b) fault-tolerance and hot-swapping and c) distributed or independent utilisation of different converter modules have been looked at in different contexts but not together in a hybrid energy system.

These are important and relevant issues in second life battery integration as the vehicle batteries, for example, come in the range of 12V, 24V up to 600V with ratings of 0.5kWh up to 50kWh. The smallest accessible module may be anything from a cell to a complete system. As a result, depending on the battery availability, the sum of battery/dc side voltages can be greater or less than the desired dc-link voltage of the line side inverter under normal operation which makes it difficult to integrate them to a common inverter dc-link using either a boost or a buck type modular converter. Moreover, these hybrid batteries tend to react in significantly different way based on their characteristics and SOH where the conventional voltage/SOC balancing strategies as reported in [11], [18] are not applicable. In order to control these batteries according to their characteristics each module may need to be controlled at significantly different current levels within the converter. As a result, the integration of such hybrid batteries is challenging.

There are no research reports to date about the power converter interface and control issues to integrate widely different battery types (new/second life) within a grid-tie energy storage system. This paper addresses these shortfalls in steps by proposing: a) a modular boost-multilevel buck dc-dc converter topology based on a cascaded structure to integrate hybrid second life battery modules to an inverter dc-link irrespective of their voltage levels and characteristics, b) a power sharing strategy based on weighting functions to utilise different battery modules depending on their relative characteristics within a set of hybrid batteries and c) a flexible distributed control structure based on a boost-buck operational mode which allows control of each converter module in a wide range while providing the grid support. The paper is structured as follows: converter topology is presented in section II, dynamic modeling, power sharing and distributed control structure are given in section III, section IV and section V respectively. The experimental validation is provided in section VI and section VII concludes the paper.

II. CONVERTER TOPOLOGY

This paper describes an H-bridge cascaded boost-multilevel dc-dc converter based topology with an inverter as shown in Fig. 1. This topology offers a good compromise between the cost, efficiency and reliability while maintaining the flexibility to deal with different battery voltages and power levels. By choosing an H-Bridge module, the converter can be operated in any of boost, buck and boost-buck mode by independently controlling the two legs of an H-bridge to enable different battery voltages to be dealt with for a required dc bus voltage with a narrow voltage range for grid inversion.

A boost type module is generally considered when all the dc-sources, be they, generation or batteries have low voltages which meets the condition $\sum V_{batt,i} < V_{dc}$ [15], [16]. However, this structure is not fault-tolerant and has an inherent current limitation ($i_{batt,i} \geq i_{dc}$) which makes it unsuitable when module currents demand to be lower the dc-link current.

On the other hand, a Buck type module could be used only when all the dc-sources meet the condition $\sum V_{batt,i} > V_{dc}$. The main advantages of a buck type module are: a) the ability to utilise a module better than a boost module because it is possible to make $i_{batt,i} < i_{dc}$, b) fault-tolerant in nature as it can bypass a faulty module. However, there are some limitations of this mode also which includes the need for a large number of modules if only low voltage batteries are available and the input current is discontinuous which adds the requirement for a high input capacitor across each battery module which increases the overall system size and cost.

A buck-boost module configuration has previously been compared to a boost and buck configuration for systems with mixed voltages [15]. The author concluded that there was poor switch utilization, low converter efficiency, large capacitors on both inputs and outputs which possibly increase the module size/cost and reduces the reliability of a converter module. Moreover, the requirement of different switch rating/module makes the module selection and design difficult.

In order to overcome these drawbacks, a modular boost-multilevel buck configuration is used with a single multi-switch buck converter (see Fig 1). The proposed configuration uses a boost type module with each battery bank and then adds a half-bridge in parallel with each module capacitor to make it fault-tolerant in boost mode and provide an opportunity to switch in the buck mode. Thereafter, these half-bridges are connected in cascade through a inductor to integrate all the modules to a common dc-link voltage as shown in Fig 1(a). Due to having a boost module at the input of each module, it can avoid a high input capacitor due to the continuous current which reduces size, cost of a module and makes the design easier because of the uniform voltage rating for all the devices.
A. Principle of operation

Each input converter \((S_i, S_j)\) can work in PWM mode to form the module capacitor voltages \(V_{\text{dc},1}, V_{\text{dc},2} \ldots V_{\text{dc},n}\) and then the cascaded half-bridge switches \((T_i, T_j)\) can be operated as a multilevel buck converter to similar [19] in order to maintain the desired central dc-link voltage \((V_{\text{dc}})\). In this way, the converter behaves as a two-stage dc-dc converter as indicated in Fig 1(a). The proposed topology configuration offers a large degree of flexibility, using an appropriate combination of converter module switches \((S_i, S_j, T_i, T_j)\) where each module may be operated in boost (buck module bypassed, \(T_i, T_j\) in idle), in buck mode (with the boost module bypassed, \(S_i, S_j\) in idle) or in boost-buck mode \((S_i, S_j, T_i, T_j\) in PWM). As a result of this, the overall converter also achieves a wide operational flexibility as shown in Table I.

![Diagram](image)

**Table I**

<table>
<thead>
<tr>
<th>Possible Modes</th>
<th>Switching of (S_i, S_j)</th>
<th>Switching (T_i, T_j)</th>
<th>Overall operation and range of control</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>(S_i, S_j) in PWM mode ((\forall i = 1 \ldots n))</td>
<td>(T_i) and (T_j) in idle mode ((T_i) is ON, (T_j) is OFF ((\forall i = 1 \ldots n))</td>
<td>Cascaded Boost Range (- (i_{\text{batt},i} \geq i_{\text{dc}}))</td>
</tr>
<tr>
<td>2</td>
<td>(S_i, S_j) in idle mode ((S_i, S_j = \text{OFF}) \ (\forall i = 1 \ldots n))</td>
<td>(T_i) and (T_j) in PWM mode ((\forall i = 1 \ldots n))</td>
<td>Multilevel buck Range (- (i_{\text{batt},i} \leq i_{\text{dc}}))</td>
</tr>
<tr>
<td>3</td>
<td>(S_i, S_j) in PWM mode ((\forall i = 1 \ldots n))</td>
<td>(T_i) and (T_j) in PWM mode ((\forall i = 1 \ldots n))</td>
<td>Boost - multilevel buck Range (- (i_{\text{batt},i} \leq i_{\text{dc}}))</td>
</tr>
<tr>
<td>4</td>
<td>(S_i, S_j) in idle</td>
<td>(T_i) is OFF and (T_j) is ON (\text{(bypass mode)})</td>
<td>Bypass (i^\text{th}) module</td>
</tr>
</tbody>
</table>

The application range expressed as \(\sum V_{\text{batt},i}\) against the control flexibility expressed as a range of \(i_{\text{batt},i}\) has been shown on a plane in Fig 1(b) to understand the operational envelope of the proposed converter in different modes. It can be seen from Fig 1(b) that the boost and the buck mode offer a narrow operating envelope on the plane. The boost mode suffers from a current limitation problem \((i_{\text{batt},i} \geq i_{\text{dc}})\) while a buck mode has an inherent voltage limitation issue \((\sum V_{\text{batt},i} > V_{\text{dc}})\). The boost-buck mode provides a wide control flexibility \((i_{\text{batt},i} \geq i_{\text{dc}} \text{ and } i_{\text{batt},i} \leq i_{\text{dc}})\) over any application range \((\sum V_{\text{batt},i} > V_{\text{dc}} \text{ and } \sum V_{\text{batt},i} \leq V_{\text{dc}})\) within a pre-defined design envelope. Therefore, this mode is more suitable to integrate widely different characteristics batteries to an inverter dc-link. Within each battery module the cells have balancing circuits similar to [5] – [6] in order to uniformly utilise each cell.

B. Module Design

The design of an H-bridge module, inductor and capacitor is challenging because of the uneven module voltages and potential variation of module current. There are two approaches: a) designing different module switch ratings and inductor/capacitor values according to individual module battery parameters or b) designing a uniform switch rating and inductor/capacitor by pre-defining a maximum battery voltage and current envelope. The latter approach has been followed in this work because different switch rating and passive component ratings in different modules complicates the design of the modular converter and prevents hot-swapping of a faulty battery module with any available battery module. A battery size envelope was used to choose the appropriate switch rating. The boost inductor was chosen to give a maximum of 5% current ripple as calculated in (1) where \(D_{\text{max}}\) is the maximum duty cycle of \(S_i, S_j, \Delta V_{\text{batt},i}\) is the magnitude of the ripple current, \(T_s\) is the switching time and \(V_{\text{sw}}\) is the maximum switch voltage rating of an H-bridge which is chosen to be 5 the times maximum battery voltage to account for a reasonable boost ratio and a margin of around 20%.

\[
L = D_{\text{max}} \times \frac{\max (V_{\text{batt},i})}{\Delta V_{\text{batt},i}} \times T_s = D_{\text{max}} \times \frac{\max (V_{\text{batt},i})}{0.05 \times \max (i_{\text{batt},i})} \times T_s \tag{1}
\]

The design of the module capacitor was set to correspond to a maximum allowable voltage ripple, \(\Delta V\), on the module (e.g.
1% of the capacitor voltage) as shown in (2). \( I_{dc,i} \) acts as a load to each module (Fig 1) and it depends on duty ratio of \( T_{ii}, T_{ii} (= D_{ii}) \) which is taken as unity to design the maximum value of \( C \).

\[
C = D_{\text{max}} \frac{(l_{dc})}{\Delta V} T_s \quad \text{or} \quad C = D_{\text{max}} \frac{(l_{dc})}{\Delta V} T_s
\]

Where \( \Delta V = 0.01 \times \max(V_{dc,i}) \) \( \quad (2) \)

### III. CONVERTER MODELLING AND CONTROL RANGE

The dynamic modelling of the converter in Fig 1(a) is composed of two stages: a) modelling input boost converter, b) modelling of modular multilevel buck converter.

#### A. Input side boost converter

The dynamic equations for the input side boost converter can be written as follows considering \( n \) converter modules:

\[
L \frac{d}{dt} i_{\text{batt},i} + R i_{\text{batt},i} + (1 - d_i)V_{dc,i} = V_{\text{batt},i} \quad \forall \ i = 1 \ldots n
\]

\[
C \frac{d}{dt} V_{dc} - (1 - d_i)b_{\text{batt},i} = -i_{\text{batt},i} \quad \forall \ i = 1 \ldots n
\]

\[
(V_{dc,i})_{\text{av}} = \frac{1}{1 - d_i} V_{\text{batt},i} \quad \forall \ i = 1 \ldots n
\]

It is necessary to investigate the model to accurately predict the dynamics of the control system. The required small model can be expressed by the state-space equations as shown in (6) employing two state-variables \( V_{dc,i} \) and \( i_{\text{batt},i} \) per module.

\[
\dot{X} = \begin{pmatrix} \frac{-R}{L} & \frac{1}{L} \\ \frac{-1}{c} & \frac{-1}{c} \end{pmatrix} X + \begin{pmatrix} \frac{V_{dc,i}}{L} \\ 0 \end{pmatrix} U \quad \forall \ i = 1 \ldots n
\]

Where, \( X = \begin{pmatrix} i_{\text{batt},i} \\ V_{dc,i} \end{pmatrix} \), \( U = \begin{pmatrix} V_{\text{batt},i} \\ \frac{d_{\text{inv}}}{i_{\text{inv}}} \end{pmatrix} \) \( \quad (6) \)

#### B. Multilevel Buck Converter

The cascaded switches \( T_{ii}, T_{ii} \forall i = 1 \ldots n \) at the output of each boost converter work as a combined buck converter along with the dc-link inductor as highlighted in Fig 1(a). Each of these switches can have an individual duty ratio \( d_{ii} \) and an overall average duty ratio \( d_{av} \) of the buck converter. This \( d_{ii} \) controls each module current \( b_{\text{batt},i} \) by controlling \( i_{\text{dc,i}} \) while \( d_{av} \) maintains the central dc-link voltage (\( V_{dc} \)) irrespective of \( d_{dc} \). The dynamic equations can be written as follows:

\[
L \frac{d}{dt} i_{\text{dc},i} + R i_{\text{dc},i} + V_{dc} = d_{av} \sum V_{dc,i} \quad \forall \ i = 1 \ldots n
\]

\[
d_{av} = \frac{V_{dc}}{\sum V_{dc,i}} = \frac{\sum V_{dc,i} i_{\text{dc,i}}}{\sum V_{dc,i}} \quad \forall \ i = 1 \ldots n
\]

\[
l_{\text{dc,i}} = D_{ii} l_{\text{dc}} \quad \forall \ i = 1 \ldots n
\]

\[
C \frac{d}{dt} V_{dc} = i_{\text{dc}} - i_{\text{inv}}
\]

The state-space model can be written as following:

\[
(\dot{X}) = \left( \begin{matrix} \frac{-R_{dc}}{L} & \frac{1}{L} \\ \frac{-1}{c_{dc}} & \frac{-1}{c_{dc}} \end{matrix} \right) (X) + \left( \begin{matrix} \frac{d_{av}}{L} & \frac{\sum V_{dc,i}}{L} \\ 0 & 0 \end{matrix} \right) (U) \forall \ i = 1 \ldots n
\]

Where \( X = \begin{pmatrix} V_{dc} \\ \frac{d_{av}}{i_{\text{inv}}} \end{pmatrix} \), \( U = \begin{pmatrix} \sum V_{dc,i} \end{pmatrix} \) \( \quad (11) \)

### C. Control Range of the Converter

The control range of the converter is essentially the relationship between \( b_{\text{batt},i} \) and \( l_{\text{dc}} \) because it provides an information about the converter operating range as shown in Fig 1(b). The power balance equation can be used to find this relationship using (12) where \( \eta \) is the efficiency of the module boost converter which is approximated to unity.

\[
V_{dc,i} l_{\text{dc},i} = \eta V_{\text{batt},i} b_{\text{batt},i} \forall \ i = 1 \ldots n
\]

\[
D_{ii} l_{\text{dc}} = \frac{V_{\text{batt},i}}{V_{dc}} b_{\text{batt},i} = (1 - D_{ii}) l_{\text{batt},i} \forall \ i = 1 \ldots n
\]

\[
b_{\text{batt},i} = \frac{D_{ii}}{(1 - D_{ii})} l_{\text{dc}} \forall \ i = 1 \ldots n
\]

It can be seen from (14) that a module current can be controlled to more or less than the dc-link current \( (l_{dc}) \) using an appropriate combination of \( D_{ii} \) and \( D_{ii} \). If \( b_{\text{batt},i} > l_{dc} \) then the condition \( (D_{ii} + D_{ii}) > 1 \) needs to be satisfied and \( b_{\text{batt},i} < l_{dc} \) the condition \( (D_{ii} + D_{ii}) < 1 \) needs to be fulfilled.

### IV. POWER SHARING STRATEGY FOR HYBRID BATTERIES

The control/power sharing of modular energy storage systems has been done previously using a single type of batteries as described earlier concentrating on the balancing control [11], [18], [20]. A distributed power sharing was performed in mainly PV system where the modular dc–dc converters were controlled according to the distributed MPPT [21] – [22]. However, the major differences between PV systems and a hybrid battery system are: a) control architecture in PV system is mainly unidirectional whereas in BESS it has to be bidirectional, b) the criteria for distributing the power between modules is dependent on multiple battery parameters such as, voltage, capacity etc. unlike in PV system which is solely dependent on different radiation conditions. The distributed sharing proposed in the paper is based on weighting factors \( \omega_i \) which represent the status (“goodness” or “badness”) of each battery module on an instantaneous basis. Since each of these hybrid module types will charge/discharge at different rates and have different maximum/minimum voltages and capacities, this strategy ensures that the charging/discharging trajectory of the hybrid modules during a charging/discharging cycle will all arrive at their maximum/minimum values at the same time using a current sharing strategy shown in (15). The derivation of the current sharing strategy and the desired weighting factor are provided in the Appendix.

\[
\omega_1 = \omega_2 = \ldots = \omega_n = f(V_{\text{batt},i}, SOC_{0,ii}, Q_{max,i})
\]
V. DISTRIBUTED CONTROL ARCHITECTURE

The relative weighting factors (i.e. $\omega_i; \omega_j$) can be significantly different in this hybrid battery integration because of differences in capacity ($Q_{\text{max},i}$) and initial state-of-charge ($SOC_{0,i}$) within a set of batteries. Therefore, the proposed control structure employs the boost-buck control mode to achieve a wide operational envelope as explained in Fig 1(b).

The main objectives of the control architecture are a) to control the central dc-bus $V_{dc}$ to a fixed value irrespective of the set of batteries present to allow the line side inverter to respond according to the desired grid side power demand, b) to control the hybrid battery modules according to the desired battery weighting factors ($\omega_i$) to optimally utilise them.

The principle concept of the proposed distributed control is to control the input side boost converters ($S_i, S_j$) to form equal module dc-link voltages ($V_{dc,m}$) irrespective of their input voltages and then utilise a concept of distributed duty ratio ($d_i \forall i = 1...n$) of the buck converter switches ($T_{b,p}, T_{b,q}$) as a function of battery weighting factors ($\omega_i$). This control operation makes the converter behave as a multilevel converter as indicated by the effect of the different module duty ratio’s in Fig 1(c). There could be two possible cases depending on the set of batteries present: a) a similar range of battery voltages and b) a widely different battery voltages. Therefore, two control techniques have been proposed.

A. Case 1 – All modules in boost-buck mode (dc-side control)

This control is used where all the modules are operated in boost-buck mode, e.g. all modules have similar voltages 12V with 24V etc. The module input voltages are boosted to $V_{dc,i}, V_{dc,j}, \ldots, V_{dc,n}$ using the input side boost converters ($S_i, S_j \forall i = 1...n$) and then the overall voltage ($\sum V_{dc,i}$) is bucked using the multilevel buck converter ($T_{b,p}, T_{b,q}$) to maintain the central inverter dc-link voltage constant at the time. All the module boost converters are controlled to a same voltage reference $V_{dc,m}$ independent of the weighting factors where the upper limit is limited by the maximum switch rating $V_{sw}$ of a module as shown in (16). The desired module independent control is achieved using the duty ratios $d_i \forall i = 1...n$. The proposed control structure is shown in Fig 2. The derivation is presented in (17) – (19) assuming $\eta_i \approx 1$.

$$nV_{dc,m} > V_{dc} \rightarrow \frac{V_{dc}}{n} < V_{dc,m} \leq V_{sw}$$

From (12), $V_{dc,i} = V_{dc,n}(d_i)l_{dc} \approx V_{batt,i}i_{batt,i}$ therefore,

$$d_i \propto V_{batt,i}i_{batt,i} \forall i = 1...n$$

According to (8), $d_{av} = \frac{\sum V_{dc,i}d_i}{\sum V_{dc,i}}$

After putting $V_{dc,i} = V_{dc,m} \forall i = 1...n$ in (8)

$$d_{av} = \frac{\sum V_{dc,m}d_i}{nV_{dc,m}} \rightarrow \sum d_i = nd_{av}$$

Therefore, $d_i = nd_{av} \frac{\omega_iV_{batt,i}}{\sum \omega_iV_{batt,i}} \forall i = 1...n$ (19)

B. Case 2 – Boost-k out of n modules only in buck mode (dc-side control)

This dc-side control is different, where not all the modules can operate in boost-buck mode. For example, where the battery module voltages are considered to be substantially different e.g. 24V with 220V etc. or when the switch rating is not sufficiently high enough to allow the boost operation of a higher voltage connected battery. A control strategy to deal with this is: a) to operate the higher input voltage module only in buck mode with the corresponding boost converters ($k$) in idle mode which means $S_i, S_{\text{bk}}$ in idle ($V_{dc,k} = V_{batt,k}$) and $T_b, T_{\text{bk}}$ in PWM, b) to operate the remaining modules ($n-k$) in boost-
buck mode as previously described which means $S_i, S_{ii}$ in PWM and $T_i, T_{ii}$ in PWM $i \neq k$ as shown in Fig 2. The module voltage reference ($V_{dc,m}$) is calculated in this case using (20). The duty ratio distribution is derived in (21) – (24).

$$\frac{(V_{dc,i} - \sum_{i \neq k} V_{batt,k})}{(n-k)} < V_{dc,m} < V_{sw} \forall i \neq k \quad (20)$$

From (12), $V_{dc,i} = V_{dc,m}, d_{ii} \propto \omega_i V_{batt,i} \forall i \neq k = 1 \ldots n$ For the $(k)$ modules in buck mode, from (12);

$$V_{dc,k} = V_{batt,k} V_{batt,k} d_{kk} = V_{batt,k} i_{batt,k} \rightarrow d_{kk} \propto \omega_k \quad (21)$$

Now with the help of (8), following expressions are derived:

$$V_{dc,m} \sum_{i \neq k} d_{ii} + \sum_k V_{batt,k} d_{kk} = d_{av} \left( (n-k)V_{dc,m} + \sum_k V_{batt,k} \right) \quad (22)$$

Equating, $V_{dc,m} \sum_{i \neq k} d_{ii} = d_{av}(n-k)V_{dc,m}$ and

$$\sum_k V_{batt,k} d_{kk} = d_{av} \sum_k V_{batt,k}$$

the following can be written,

$$\sum_{i \neq k} d_{ii} = (n-k) d_{av}$$

and

$$d_{av} = \frac{\sum_k V_{batt,k} d_{kk}}{\sum_k V_{batt,k}} \quad (23)$$

Now using (18) and (21) the following expressions can be derived

$$d_{ii,(i \neq k)} = (n-k) d_{av}$$

and

$$d_{kk} = d_{av} \frac{\sum_k V_{batt,k} d_{kk}}{\sum_k V_{batt,k}}$$

Where $d_{kk} = d_{av}$ for $k=1 \quad (24)$

C. Overall Control Structure

The detailed control structures for both the dc-side and the ac-side are shown in Fig 2 and Fig 3 respectively. The dc-side control in Fig 2 is composed of two-stages: a) control of the module boost converter and b) control of the overall multilevel buck converter. The module boost converters are controlled in voltage mode using an outer PI-voltage loop and an inner P-current loop. The selection of the voltage reference is done using (16) or (20). The proposed control structure of the multilevel buck converter is shown in Fig 2(b). It consists of two-stages: a) battery parameter estimation and weighting factor generation and b) closed loop control. The control employs a central dc-bus voltage loop and the output of that controller provides the overall duty ratio ($d_{av}$) command through an inner dc-link current controller. Thereafter, this $d_{av}$ is split into module duty ratios ($d_{ii}$) using (19) or (24) which maintains $V_{dc}$ as well as control the module currents in the desired manner. The grid side control in Fig 3 depends on the type of application of the energy storage system e.g. voltage control or the frequency control. The output of the voltage or frequency controller provides the reference for the inner $q$-axis (active power axis) current loop. The $d$-axis is taken as the reactive power axis. This $i_{sq}^*$ plays an important role because the weighting factor for charging and discharging is different on the dc-side. The sign of $i_{sq}^*$ indicates the phase of line side current with respect the line voltage. Therefore, this dynamic change-over is performed using the sign of $i_{sq}^*$ though an edge detector in the proposed control. The initial OCV is updated at the end of charging and discharging cycle. This is performed through a sample and hold (S/H) logic and an edge detector.

VI. EXPERIMENTAL INVESTIGATION

A four-module or five-level hybrid battery energy storage system has been built as shown in Fig 4 and tested in a grid connected condition to validate the proposed control. Different converter modules contain different batteries in terms of chemistry/type, voltage, capacity as shown in Table II. Initial characterizations such as pulse load tests [23] have been performed to find the initial battery parameters such as internal impedance ($Z$) before putting it into a converter module as shown in Table II. Thereafter, an online estimation method is employed to track these parameters during operation as described in Appendix. The inverter was controlled to meet a fixed grid side power demand.

![Fig 3 Line side inverter control structure for grid support application](image1)

![Fig 4 Multi-modular hybrid prototype: a) overall set-up, b) converter module](image2)
A. Converter and Multilevel Operation

In order to show the multilevel operation of the converter, the voltage $V_{AB}$ in Fig 1(a) is measured. Since all the switches $T_1$, $T_2$ operate in different duty ratio depending on weighting factors, the voltage $V_{AB}$ varies between zero and a combination of the different voltage levels. e.g. $V_{dc,m}$, $2V_{dc,m}$, $3V_{dc,m}$, $4V_{dc,m}$ in a switching cycle. Fig 5(a) shows this operation. It can be seen that there is five distinct voltage levels of the converter.

B. Distributed Power Sharing (Current Dynamics)

Mode – 1 zero to discharging mode: Fig 5(b) shows the experimental result for the distributed control scheme at the moment of connecting to the grid. It can be seen that the converter distinctly utilises the hybrid modules where the highest and the lowest module currents are almost in 1:8 ratio. The module currents are dependent on the instantaneous weighting factors. Table III shows the comparison of calculated and measured currents.

Mode – 2 zero to charging mode (current dynamics): Fig 5(c) shows the experimental result when the converter switches to charging mode. The second module is charged at a significantly higher current than the remaining modules. Module currents are widely different (1:10 ratio exists between the highest and lowest module currents). The steady state values of module currents are presented in Table III.

Mode – 3 charging to discharging mode (current dynamics): Fig 6(a) shows the experimental result when the converter switches from charging mode to discharging mode. It is important to note that the current sharing between the modules is different in charging and in discharging due to the differences in weighting functions between the two modes. The details are given in Table III.

Mode – 4 discharging to charging mode (current dynamics): Fig 6(b) shows the experimental result when the converter switches from discharging mode to charging mode. It is to be noted that the all the module currents in charging and discharging mode are different and the current sharing changes after switching from discharging to charging mode.

Fig. 5 Experimental result of converter operation: a) multilevel operation, b) zero to discharging, c) zero to charging scale time 20ms/div, module currents 5A/div

Fig. 6 Experimental results of current sharing while switching the mode: a) charging to discharging, b) discharging to charging, c) module bypassing, scale time 20ms/div, and module currents 5A/div

Fig. 7 Duty ratio distribution of $T_1$, $T_2$: a) charging to discharging, b) discharging to charging, c) module bypassing: Scale: duty ratio 0.5/div, time 20ms/div
C. Module Bypassing (Current Dynamics)

Fig 6(c) shows the experimental result of module bypassing during discharging. In this experiment, module – 4 (24V, 16Ah) has been bypassed by making \( T_4 \) OFF and \( T_{d4} \) ON. The remaining modules take a higher share of the currents to keep the same power. A momentary drop in the currents occurs due to the sudden dip in \( V_{dc} \) which reduces \( d_{ov} \) and subsequently \( d_i/s \) but then recovers quickly. This shows the fault tolerance of the converter when one of the module bypasses.

### Table II

<table>
<thead>
<tr>
<th>Type/Name</th>
<th>Rating/specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>OptiMOS for H-bridge dc-dc modules</td>
<td>100V 40A – (FDFP085N10A)</td>
</tr>
<tr>
<td>Boost inductors of dc-dc modules ((L))</td>
<td>1.5mH, 15A, ( R_L = 20m\Omega )</td>
</tr>
<tr>
<td>dc-link inductor ((L_d))</td>
<td>1.5mH, 15A, ( R_L = 40m\Omega )</td>
</tr>
<tr>
<td>module dc-link capacitors</td>
<td>2200µF, 100V</td>
</tr>
<tr>
<td>Switching frequency of the dc-dc module and inverter</td>
<td>10kHz</td>
</tr>
<tr>
<td>module dc-link voltage ((V_{dc,a}))</td>
<td>80V</td>
</tr>
<tr>
<td>Operating central dc-bus voltage ((V_{dc}))</td>
<td>150V</td>
</tr>
<tr>
<td>Line side filters ((L, C))</td>
<td>3mH, 3mH and 10 ( \mu )F</td>
</tr>
<tr>
<td>Nominal Grid voltage ((V))</td>
<td>120V (peak), 50Hz</td>
</tr>
<tr>
<td>Test power level ((P))</td>
<td>450W</td>
</tr>
<tr>
<td>Battery module – 1 (lead acid)</td>
<td>12V, 10Ah – ( P_{max} = 14V ), ( V_{max} = 9.5V ), Initial ( Z = 0.015\Omega )</td>
</tr>
<tr>
<td>Battery module – 2 (lithium titanate)</td>
<td>24V, 60Ah – ( P_{max} = 27V ), ( V_{max} = 18V ), Initial ( Z = 0.006\Omega )</td>
</tr>
<tr>
<td>Battery module – 3 (NiMH)</td>
<td>7.2V, 5.5Ah, ( V_{max} = 8.5V ), Initial ( Z = 0.01\Omega )</td>
</tr>
<tr>
<td>Battery module – 4 (lead acid)</td>
<td>24V (2x12V), 16Ah lead acid – ( P_{max} = 18V ), ( V_{max} = 28V ), Initial ( Z = 0.022\Omega )</td>
</tr>
</tbody>
</table>

### Table III

<table>
<thead>
<tr>
<th>Cases</th>
<th>Measured initial SOC/OCV before changing a mode</th>
<th>Calculated current references (Eq. A.13)</th>
<th>Measured steady-state module currents</th>
</tr>
</thead>
<tbody>
<tr>
<td>Mode – 1</td>
<td>( OC_{101} = 11.21V )</td>
<td>( h_{bat,1} = 1.5A ), ( h_{bat,2} = 11.5A ), ( h_{bat,3} = 3A ), ( h_{bat,4} = 3.9A )</td>
<td>( h_{bat,1} = 1.3A ), ( h_{bat,2} = 10A ), ( h_{bat,3} = 2.5A ), ( h_{bat,4} = 3.5A )</td>
</tr>
<tr>
<td>Mode – 2</td>
<td>( OC_{102} = 12.8V )</td>
<td>( h_{bat,1} = -0.6A ), ( h_{bat,2} = -6.1A ), ( h_{bat,3} = -0.4A ), ( h_{bat,4} = 2.0A )</td>
<td>( h_{bat,1} = -0.5A ), ( h_{bat,2} = -5A ), ( h_{bat,3} = -0.55A ), ( h_{bat,4} = 1.8A )</td>
</tr>
<tr>
<td>Mode – 3</td>
<td>( OC_{103} = 22.6V )</td>
<td>( h_{bat,1} = 9.5A ), ( h_{bat,2} = 1A ), ( h_{bat,3} = 9.6A ), ( h_{bat,4} = 3.2A )</td>
<td>( h_{bat,1} = 9.6A ), ( h_{bat,2} = 15A ), ( h_{bat,3} = 3A )</td>
</tr>
<tr>
<td>Mode – 4</td>
<td>( OC_{104} = 23.5V )</td>
<td>( h_{bat,1} = -0.9A ), ( h_{bat,2} = -0.4A ), ( h_{bat,3} = -0.5A ), ( h_{bat,4} = -2.1A )</td>
<td>( h_{bat,1} = -0.7A ), ( h_{bat,2} = -4.8A ), ( h_{bat,3} = -0.4A ), ( h_{bat,4} = -1.9A )</td>
</tr>
</tbody>
</table>

D. Duty Ratio Distribution

The converter in Fig 1 is controlled in boost-buck mode where the pair of module switches \( S_i, S_j \) and \( T_{di}, T_{dj} \) operates in PWM fashion to utilize the hybrid modules. Fig 7 shows how the duty ratios are used to change the power sharing during the transition of modes. Fig 7(a) shows the distribution of module duty ratio while switching from charging to discharging and Fig 7(b) shows the similar waveform while moving from discharging to charging mode. On the other hand, Fig 7(c) shows the module duty ratio distribution when module – 4 bypasses. It can be seen that the module duty ratios \( (d_i) \) redistribute among themselves to keep the overall duty ratio \( (d_o) \) constant all the time because the total dc-bus voltage \( (V_{dc}) \) is controlled to a desired value through a central voltage loop as shown in Fig 2.

E. Module voltage dynamics

Fig 8 shows the dynamics of module dc-link voltages \( (V_{dc,i}) \) at the start up. Note all the voltages are equal \((80V)\) in steady-state as expected from the design presented in section V after undergoing slightly different transient response.

F. Validation of overall control

Fig 9 shows the experimental results for grid side and dc-link side control system operation at a moment in time when the inverter is switched from the charging to discharging mode. It can be seen that the phase angle of the line side current changes with respect to the line voltage when the dc-link current \( (I_{dc}) \) moves from negative to positive. It is to be noted that the total DC-link voltage \( (V_{dc}) \) stays constant during the transition due to the central dc-link voltage control loop.

G. Converter Efficiency

The overall converter efficiency from the prototype was measured in two steps: a) efficiency of the modular dc-dc converter consisting of four modules, b) efficiency of the overall converter along with the grid-tie inverter. The
cascaded boost-multilevel buck converter from the experimental prototype has a measured efficiency of around 96% at a 10 kHz switching frequency at the test power level. The cascaded converter has a high efficiency because the LV Trench MOSFETs (OptiMOS) used in the H-Bridge module have a very low on-state resistance \( R_{\text{dson}} = 8\, \text{m}\Omega \) as shown in Table II. The overall efficiency of the converter was found to be around 92% when the inverter is included in the calculation, however there is scope to improve this using more efficient devices in the inverter.

H. Charging/Discharging Trajectory of Hybrid Batteries

Normal Condition: In order to validate the full charging/discharging trajectory, all the battery modules were started at different initial SOC and/or voltage levels. The converter was run for a long time using the distributed strategy. The estimated state-of-charge (SOC) is plotted during charging and discharging as shown in Fig 10 and Fig 11 respectively under normal conditions using the method described in Appendix. A zero SOC corresponds to the minimum capacity condition and a unity SOC corresponds to the fully charged condition in this case. It can be seen that the module with a lower initial SOC has a larger slope compared to the module with a higher initial SOC during charging and vice-versa during discharging. The module SOC’s reach zero or unity at around the same time using the proposed strategy.

Parameter variation: Capacity fade is an important practical phenomenon in second life application. In order to validate the proposed strategy under this condition, two identical batteries of the same voltage (12V) and capacity (10Ah) were put in parallel through a dc-breaker. As a result of this, the combination behaves as a 20Ah capacity. Mid-way through the discharging experiment, one battery was disconnected through the breaker to reduce the capacity. As a result, the effective charge or capacity was halved while impedance was doubled, and this was picked up by the capacity estimator as explained in Appendix. The current sharing was affected due to this change because of the change in the weighting factors according to (A.13). This is shown in Fig 12 where the current share taken by battery – 1 is reduced while that taken by the remaining modules is increased and the discharging trajectory continues as expected. In this way, the proposed strategy remains valid even under varying parameter conditions.

VII. CONCLUSION

A modular boost-multilevel buck based converter topology and a module based distributed control architecture are proposed, analysed and experimentally validated to integrate any set of second life batteries to a grid connected energy storage system and to optimally utilise them. The proposed converter is found to be efficient and is capable of utilising widely different battery modules characteristics using the boost-buck control mode. The control structure is based on weighting factors which are dependent on module battery characteristics such as, initial SOC, voltage, impedance and capacity in order to take account their different performances. The results show the suitability of the topology and control structure when widely different batteries are present. The application is focused on single-phase but the proposed converter structure and control architecture can be extended into three-phase and other modular energy systems.
APPENDIX

There is a hybrid mix of batteries. The strategy adopted in this paper is to ensure that the charging/discharging trajectory of the hybrid battery modules during a charging or discharging cycle will all arrive at their maximum or minimum values at the same time. The following definitions have been assumed:

- A battery capacity has been taken as the maximum charge left (Q_max in C or Ah) that a battery can deliver to a load.
- A battery is modelled as an open circuit voltage (OCV) with series impedance (Z) \( \leftrightarrow \) \( V_{batt,i} = OCV_i \pm i_{batt,i} Z_c \).
- Each battery module has different initial SOC (or OCV_i) at start, different nominal voltages and also different capacity (Q_max).
- Remaining charge or capacity left in a battery module (Q) is expressed as the product of Q_max and SOC.

A. Weighting Factor Derivation

The fundamental charge equation for a battery module is given by (A.1) where the \( Q_0 \) is the initial charged stored. Now assume, the module battery charge \( Q(t) \) is some function \( f(.) \) of its open-circuit voltage OCV(t):

\[
Q(t) = Q_0 + \int_0^t i_{batt} \, dt = f(OCV(t)) \tag{A.1}
\]

Therefore,

\[
f(OCV(t)) = f(OCV_0(t)) + \int_0^t i_{batt} \, dt \tag{A.2}
\]

Where, OCV_0 is the measured module open-circuit voltage at time \( t = 0 \) (V). This OCV_0 indicates the initial SOC. For the charging/discharging strategy chosen at some time \( T = T_{min/max} \) where the charging/discharging process must cease.

Therefore, OCV (t)\(_{\text{t=0}}\) \( \equiv \) OCV_0 max:\n
\[
OCV_{\text{min/max}}: \text{ Module maximum or minimum open-circuit voltage at time } T_{\text{min/max}}. \text{ Under these conditions (A.2) becomes:}
\]

\[
f(OCV_{\text{min/max}}) = f(OCV_0) + \int_{T_{\text{min/max}}}^t i_{batt} \, dt \tag{A.3}
\]

Re-arranging in terms of the module charging current gives:

\[
\int_{T_{\text{min/max}}}^t i_{batt} \, dt = f(OCV_{\text{min/max}}) - f(OCV_0) \tag{A.4}
\]

Equation (A.5) can be solved for an approximately constant charging/discharging current condition as follows:

\[
i_{\text{batt}} \approx \frac{f(OCV_{\text{min/max}}) - f(OCV_0)}{T_{\text{min/max}}} \tag{A.6}
\]

Where, \( i_{\text{batt}} \) is the signed magnitude of the charging/discharging current. This assumption is accurate if the minimum battery voltage of a module is higher than the change in battery voltage range between fully charged to discharged otherwise inaccuracies may be introduced.

In equation (A.6), OCV_{\text{min/max}} should be known for a particular module based on pre-characterization and OCV_0 can be measured taking the first sample instant of open circuit voltage (OCV) calculated from the battery model (= V_batt \( \pm I_{\text{batt}} Z \)) as a charge/discharging cycle begins.

\[
T_{\text{min/max}} \text{ will be common across all the modules as they are to be charged/discharged at the same time. However, in a typical grid support application, } T_{\text{min/max}} \text{ will be unknown to the hybrid modules. To eliminate } T_{\text{min/max}} \text{ from (A.6), the converter power balance equation can be used as shown below. }
\]

\[
P = \sum_{k=1}^n V_{\text{batt,k}} I_{\text{batt,k}} \tag{A.7}
\]

Where \( n \) is the number of active modules and each module \( k \) has a different voltage, \( V_{\text{batt,k}} \) and current \( I_{\text{batt,k}} \).

Now, substituting (A.6) into (A.7) for each module \( k \), gives:

\[
P = \sum_{k=1}^n V_{\text{batt,k}} \frac{f(OCV_{\text{min/max},k}) - f(OCV_0,k)}{I_{\text{batt,k}}} \tag{A.8}
\]

Re-arranging (A.9) gives the desired module battery current:

\[
i_{\text{batt,1}} = P \left( \frac{f(OCV_{\text{min/max},1}) - f(OCV_0,1)}{\sum_{k=1}^n V_{\text{batt,k}} f(OCV_{\text{min/max},k}) - f(OCV_0,k)} \right) \tag{A.10}
\]

\[
i_{\text{batt,k}} = P \left( \frac{f(OCV_{\text{min/max},k}) - f(OCV_0,k)}{\sum_{k=1}^n V_{\text{batt,k}} f(OCV_{\text{min/max},k}) - f(OCV_0,k)} \right) = P \omega_i \tag{A.11}
\]

If a straight-line relationship is assumed between OCV (t) and Q (t) such as reported in [6], then \( f(OCV) \) can be written in the following form:

\[
OCV_k(t) = OCV_{\text{min+k}} + \frac{(OCV_{\text{max+k}} - OCV_{\text{min+k}})}{Q_{\text{max,k}}} Q_k(t) \tag{A.12}
\]

Substituting, (A.12) in (A.11) gives the desired \( \omega_i \):

\[
\omega_i = \left( \frac{\sum_{i=1}^n V_{\text{batt,i}} OCV_{\text{max,i}} - OCV_{\text{min,i}}}{\sum_{k=1}^n V_{\text{batt,k}} (OCV_{\text{max,k}} - OCV_{\text{min,k}})} \right) \frac{Q_{\text{max,i}}}{Q_{\text{max,k}}} \tag{A.13}
\]

B. Parameter Estimation

The proposed sharing strategy requires two important parameters: initial OCV or SOC (OCV_i), b) capacity (Q_max). The OCV_i in the weighting factor has to be updated at the end of a charging or discharging cycle as explained in Fig 2. The battery capacity (or Q_max) has to be tracked during the operation because long term battery degradations will consequently affect this capacity. Apart from the capacity, the internal impedance (Z) is also prone to vary with the SOC, age and degradations. The variation of that acts as an indicative of power fade because the state-of-health (SOH) of a battery is a combination of power fade and capacity fade as reported in

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Therefore, both the parameters are estimated to help with the power sharing and also to track the SOH online.

Part – 1: Impedance (Z) Estimation:
There are many methods to estimate the impedance of battery: a) a pulse power based method [23] b) a ripple based online impedance method similar to [24], c) EIS based techniques [25]. The pulse power based method [23] is straightforward but it requires the battery to be at rest for at least 2-5 minutes between the tests. This method has been used to find impedance of a battery as part of the pre-characterization tests prior to connecting a battery module to the converter system as shown in Table II. However, it is difficult to provide a pulse current reference externally to the battery modules when the inverter connected energy storage system is providing the necessary grid support because the power command is decided by the inverter which is dependent on the grid side demand. EIS based methods are expensive and are suitable in off-line applications. This paper uses a ripple based method described in Fig A.1 during the power converter operation. The principle concept is to use the high frequency inductor ripple current of the associated module dc-dc converter and corresponding high frequency ripple of the battery terminal voltage to calculate the internal impedance. Therefore, it is deemed more appropriate in this application because this ripple is always present during the converter operation. The switching frequency is in the order of multiples of kHz’s, so it can be assumed that the SOC/OCV does not change significantly during such small switching interval (in the order of ‘µs’).

Two different equations can be written in each switching period: at \( t = 0 \) and \( t = dT_s \) as shown below (where \( d \) represents the converter duty ratio and \( \pm \) refers to charging/discharging condition). The ripple part of the voltage and current are extracted through a low pass filter as shown in Fig A.1. The validation of the method is presented in Fig A.2 where an external impedance of 0.033Ω has been put in series with a 24V battery. It can be seen that the method is able to track the variation of impedance both increase and decrease dynamically taken out. It can be seen that the method is able to validate the method is presented in Fig A.2 where an external impedance of 0.033Ω has been put in series with a 24V battery.

Moreover, the past researches show that the battery internal impedance (Z) is related to the battery capacity (\( Q_{max} \)) either through a linear relationship [25], [27] or through a nonlinear relationship [28]. This means if one parameter changes there will be a subsequent change in the other. Therefore, both power and capacity fade are implicitly included in the formula. The combined SOH could be calculated directly using this method because both the quantities are estimated.

The proposed strategy could be valid for various temperature because any variation of temperature causes the capacity \( Q_{max} \) and impedance (Z) to change because both are functions of temperature [28]. In this work, both are estimated through online monitoring. This will consequently affect the current sharing through the weighting function in (A.13). However, in this work, the batteries were kept at ambient temperature. Further work may be necessary to explicitly validate this.

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