Abstract—This paper presents a new solid-state step-up resonant Marx modulator (S3RM2) with a continuous output current for offshore wind energy applications. The developed topology is based on the Marx generator concept, where magnetic switches are replaced by solid-state switching devices. The proposed converter is characterized by resonant switching transitions to achieve minimal switching losses and maximum system efficiency. Therefore, a higher switching frequency is conceivable to attain a higher power density. A double module consists of the 4-active switches operating at the output voltage up to seven times as high as the input voltage. An appropriate output capacitor size is considered to eliminate output voltage ripples and work as charge storage. The series-modular and cascade configurations of the S3RM2 have the advantage of being readily applicable to multilevel power switching converters with an arbitrary number of levels. The developed topology has been implemented on a 5-kW prototype converter to verify its feasibility.

I. INTRODUCTION

With increasing integration of renewable energy generation into power grids, the transmission using the DC level is becoming more and more useful than the AC transmission in terms of efficiency because no reactive power is generated/consumed by DC transmission cables [1-3]. In this regard, the emerging applications, such as offshore wind farms and photovoltaic generations, need a high step-up DC/DC conversion system to interface with high power transmission networks [4-5].

On the other hand, the heavy electrical components, such as bulky and complex generators, transformers, and converters, create serious issues as they must be taken to the places that offshore wind turbines are located. Thus, the offshore wind turbines should cope with the following challenges to make high-power-density power conversion systems a feasible alternative. Bulky and huge electrical components result in high investment costs due to a more difficult erection and the equipment transportation from shore to the installation sites [5]. In other words, if the high-power-density electrical components are utilized in offshore wind farms, offshore floating platforms can be fully assembled and installed with a reduced burden for large vessels. In addition, there is a greater need for the high reliability owing to the inherent lack of turbine access at sea which makes operation and maintenance more difficult. Smaller stages of the conversion and the device count mean lower costs, fewer failure points, and lower power losses. Hence, a power conversion system herein should feature high power density, high efficiency, high reliability, and low costs.

Step-up or boost converters are theoretically able to achieve infinitely high voltage conversion ratios; however, the maximum gain is practically limited by circuit imperfections, such as parasitic elements and switch commutation times [7-8]. To overcome this limitation, a hybrid boost or buck/boost converter was proposed to achieve a high conversion ratio for offshore wind farms applications [9]. Nonetheless, because the duty ratio of the main switch is quite large to achieve high-voltage gain, the switching frequency is relatively low to reduce losses and leads to enough turn-off time for switches. Hence, increasing the size of passive elements, such as boost inductors and filter capacitors, is inevitable due to the low switching frequency. On the other hand, full-bridge isolated conversion systems have been widely used for higher voltage gains. These systems rely on a high frequency transformer to gain a high power density [10-11]. Unfortunately, high frequency transformers with large turn ratios are difficult to design at high voltage and mega power levels [9].

Recently, the common types of switched-capacitor converters based on a Marx generator topology are considered as an attractive solution to meet the requirements such as high power density and control simplicity. The Marx generator is capable of achieving high voltage boost ratios and is suitable for high pulsed power technologies [12-13]. The Marx generator boosts a DC voltage by connecting in series of capacitors that are charged in parallel to an input voltage. However, they produce megawatt peak powers only for a few 100 ns duration and result in a discontinuous output current waveform. Moreover, a number of switches are required to achieve high step-up ratio, which causes significant switching losses and modulation complexity. In [14], a resonant switched-capacitor (RSC) converter was investigated, where
an extra inductor was added to form a sinusoidal fashion with the capacitors to perform a soft switching. In [15], a multilevel modular RSC topology was proposed with significant benefits, including its modular structure, low voltage stress of the switches, and reduced switching loss. However, all the aforementioned topologies require a large number of capacitors and inductors that dramatically increase the physical size and cost of converters.

This paper describes a new solid-state step-up resonant Marx modulator (S3RM2) for offshore wind energy systems. A 7-level S3RM2, including principle of the operation and analysis, is presented in Section II. In Section III, a 15-level S3RM2 is verified via the simulation and compared to an RSC converter to demonstrate its advantages for high voltage and high power offshore wind applications. Finally, the experimental results on a 5-kW prototype converter are presented in Section IV.

II. S3RM2 CONFIGURATION

A. General Topology

Fig. 1 shows a general configuration with an AC grid and an AC/DC converter in the front-end of the proposed S3RM2. A full bridge converter supplies the S3RM2 with a constant input DC voltage. A large capacitor is assumed to be used for the energy storage and the voltage filter at the output of the AC/DC converter. The S3RM2 uses a new arrangement of the solid-state switches, diodes, capacitors, and inductors.

B. Principle of S3RM2 Operation

A 7-level S3RM2 with two modules is shown in Fig. 2(a). It is composed of four resonant capacitors (Cru1, Cru2, Crd1, and Crd2), two output capacitors (Cuo and Cdo), four resonant inductors (Lru1, Lru2, Lrd1, and Lrd2), two output resonant inductors (Luo and Ldo), six diodes (Db1, Db2, Dbo, Dc1, Dc2, and Dco), and four switches (SB1, SB2, SC1, and SC2). Here, the following assumptions are made to simplify the analysis:

1) All the switches, diodes, capacitors, and inductors are ideal;
2) All the capacitances are equal and the inductors have same values;
3) The switching frequency is less than the resonant frequency to achieve a zero-current switching (ZCS) [14-15];
4) Vs is an ideal DC voltage source and the load is modeled by a pure resistor (Rload).

1) Mode I \([t_0, t_1]\) [Fig. 2(b)]

In the beginning of this mode (\(t=t_0\)), SB1 and SB2 are ON while SC1 and SC2 are OFF (see Fig. 3(a) and (b)). The charging currents flow through (Db1, SB1) and (Db2, SB2), as depicted in Fig. 2(b). Therefore, Cru1 and Cru2 are charged whereas Crd1 and Crd2 were previously charged one time and two times the input voltage level in Mode III, respectively. The resonant inductor currents then rise and fall in a sinusoidal fashion, as shown in Fig. 3(e) and (f). This mode ends when the resonant inductor currents reach zero and all the energy stored in three resonant loops has been transferred to Cru1, Cru2, and Cdo at \(t=t_1\). At this point, Cru1, Cru2, and Cdo are charged up to one time, two, and three times the input voltage level, respectively. Then, SB1 and SB2 can be OFF under the zero-current condition, as shown in Fig. 3(h).

Here, an opposite energy transmission is not allowed because Db1, Db2, and Dbo make three unidirectional paths in the resonance circuit and block the reverse current flows.

2) Mode II \([t_1, t_2]\) [Fig. 2(c)]

In this mode, all the switches and diodes are turned OFF. The resonances stop at three loops as depicted in Fig. 2(c). Therefore, the inductor currents are equal to zero. The resonant capacitor voltages of Cru1, Cru2, Crd1, and Crd2 are unchanged. The output capacitor voltages of Cuo and Cdo were charged up to the three times the input voltage
level in Mode I and Mode III, respectively) are discharged to the load as shown in Figs. 2(c) and 3(g).

Fig. 2. Equivalent Circuits of the 7-level S'RM². (a) Circuit configuration. (b) Mode I [t₀, t₁]. (c) Mode II [t₁, t₂]. (d) Mode III [t₂, t₃].

Fig. 3. Key waveforms of the 7-level S'RM² at steady-state. (a) and (b) Switching patterns. (c) Input current. (d) Output inductor currents. (e) and (f) Resonant inductor currents. (g) Output capacitor currents. (h) and (j) Switch currents.

3) Mode III [t₂, t₃] [Fig. 2(d)]

At the instant t=t₂, SC₁ and SC₂ are turned ON, while S B₁ and S B₂ are OFF. It can be seen from Fig. 3(j), the currents through SC₁ and SC₂ are increased by a soft-switching operation with the half-cycle resonant shape. In this mode, C ru₁ and C ru₂ are discharged to C uo while C rd₁ and C rd₂ are charged through a resonant phenomenon as shown in Fig. 2(d). In this mode, D C₁, D C₂, and D C₀ make three unidirectional paths in the resonant circuit in order to avoid opposite energy transmission. Then, the current through L uo is decreased to zero after the half-resonant period (refer to Fig. 3(d)). At the time of t₃, SC₁ and SC₂ become OFF under the zero-current condition as illustrated in Fig. 3(j).

4) Mode IV [t₃, t₄] [Fig. 2(c)]

The operation of this mode is similar to that of Mode II. Therefore, all the switches and diodes are turned OFF. The resonances stop at three loops as shown in Fig. 2(c). The inductor currents are equal to zero (see Fig. 3(d), (e), and (f)).
Applying the charge balance principle to $C_{uo}$ leads to

$$
-\int_{t_0}^{t_2} P_0 \frac{dt}{V_o} + \int_{t_2}^{t_3} P_0 \frac{dt}{V_o} + \int_{t_3}^{t_4} P_0 \frac{dt}{V_o} = \int_{t_0}^{t_4} P \frac{dt}{V_o} .
$$

(1)

where $P_0$ and $V_o$ are the average output power and the average output voltage, respectively. If $t_0=0$, $t_1=t_2=T_s/2$, and $t_3=t_4=T_s$, then (1) can be rewritten as

$$
i_{t-u0} (t) = -\frac{\pi P_0}{V_o} \sin(\omega_o t)
$$

(2)

where $T_s$ and $\omega_o (\omega_o=2\pi/T_s)$ are the switching period and the resonant frequency (1/√($C_{ru}L_{ru}$)).

By applying the principle of charge balance to $C_{ru2}$, it follows that

$$
\int_{t_0}^{T_s/2} i_{C_{ru2}} (t)dt = \int_{T_s/2}^{T_s} i_{t-u0} (t)dt .
$$

(3)

Therefore, the resonant capacitor and inductor currents, and the output capacitor current of $C_{uo}$ are obtained as

$$
i_{C_{ru2}} (t) = i_{t-u2} (t) = \frac{\pi P_o}{V_o} \sin(\omega_o t) ,
$$

(4)

$$
i_{C_{ru1}} (t) = i_{t-u1} (t) = \frac{2\pi P_o}{V_o} \sin(\omega_o t) ,
$$

(5)

$$
i_{C_{uo}} (t) = \begin{cases} -\frac{P_o}{V_o} & 0 \leq t \leq T_s/2 \\
-\frac{\pi P_o}{V_o} \sin(\omega_o t) - \frac{P_o}{V_o} T_s/2 & \frac{T_s}{2} \leq t \leq T_s . \end{cases}
$$

(6)

For the k-module $S^kRM^2$ as shown in Fig. 1, the resonance inductor currents, the input current, the switch currents, and the average output voltage can be expressed as

$$
i_{r-nh} (t) = i_{t-nh} (t) = \frac{2^{k-h} \pi P_o}{V_o} \sin(\omega_o t) \quad (h=1,2,...,k) ,
$$

(7)

$$
i_{in} (t) = \frac{P_o}{V_o} + \left(2^k - 1\right) \frac{\pi P_o}{V_o} \sin(\omega_o t) ,
$$

(8)

$$
i_{S_{ch}} (t) = \begin{cases} 0 & 0 \leq t \leq T_s/2 \\
-\frac{2^{k-h+1} \pi P_o}{V_o} \sin(\omega_o t) & T_s/2 \leq t \leq T_s \quad (h=1,2,...,k) , \end{cases}
$$

(9)

$$
i_{S_{bh}} (t) = \begin{cases} \frac{2^{k-h+1} \pi P_o}{V_o} \sin(\omega_o t) & 0 \leq t \leq T_s/2 \quad (h=1,2,...,k) \\
0 & T_s/2 \leq t \leq T_s , \end{cases}
$$

(10)

$$
V_o = (2^{k+1} - 1) V_c .
$$

(11)

It can be observed from (11) that the k-module $S^kRM^2$ offers a potential for the large voltage gain by the least number of the passive components and switches owing to the exponential effect.

III. SIMULATION RESULTS AND EVALUATION OF $S^kRM^2$

Offshore wind farms in the MW range are required to interface the high voltage power systems. In order to cope with this situation, the cascade $S^kRM^2$ configuration can be introduced to achieve a high voltage gain and a high rated power. A 10-MW wind turbine with an output of 6.6 kV LL,rms is considered as an input source (see Fig. 4). This voltage will be boosted to 150 kV for the HVDC transmission through a 15-level $S^kRM^2$. The 15-level $S^kRM^2$ was developed by using MATLAB/Simulink with the PLECS Blockset [16]. The resonant frequency is set to 3 kHz. All the capacitors and inductors are selected as 100 µF and 28 µH, respectively.
Fig. 5 shows the simulation waveforms for the switching patterns, the resonant inductor currents, the output inductor currents, and the output capacitor currents from the top to bottom. It can be observed from Fig. 5(b) and (c) that all the switches can be ON and OFF under the zero-current condition. The simulation results closely match the analysis and operation of the S^3RM^2 in Section II.

Here, the series-modular and cascade S^3RM^2 configurations are evaluated against the RSC converter in [15]. The series-modular power conversion system has advantage concerning reliability at locations where maintenance and exchange of parts is a main issue, such as offshore wind energy systems. For example, if a single module fails, the converter can still function at a reduced power level.

The series-modular S^3RM^2 must boost the 10 kV DC rectifier output to the 140 kV DC. Therefore, two modular cells are designed for equal voltage gains of 7 and identical input powers of 5 MW. Fig. 8 shows the series-modular configuration that each cell includes a 7-level S^3RM^2 as shown in Fig. 2(a). Both cells operate from equal input voltages and draw the same input currents. The switch voltage stresses are 10 kV and 20 kV for S_B1 and S_B2, respectively. The diode voltage stresses are 10 kV for D_B1 and 30 kV for D_B2 and D_Bo.

The cascade S^3RM^2 is designed to boost the 10 kV DC rectifier output to the 150 kV DC. Therefore, the 3-module S^3RM^2 must provide a voltage gain of 15 as indicated in Fig. 4. The switch voltage stresses are 10 kV, 20 kV, and 40 kV for S_B1, S_B2, and S_B3, respectively. The switch currents can be obtained from (9) and (10).

The RSC converter must provide the voltage gains of 14 and 15 to compare the device count to the series-modular and cascade S^3RM^2 configurations. Therefore, two RSC converters with 13 and 14 modular cells are required. Each module consists of one resonant inductor and one resonant capacitor, one output filter capacitor, and two diodes. The values of the output filter capacitor are much larger than those of the resonant capacitor to fix the resonant frequency during a switching period. The voltage stress of the main switches reaches the input voltage level (10 kV) and the switch current stresses can be obtained in [15].

For all the converters, each switch or diode is made up of several series and parallel-connected devices to withstand the rated current and voltage. These converters are compared based on the following features: 1) total number of devices; 2) passive component weights; and 3) losses.

**Table I**

<table>
<thead>
<tr>
<th>Approach</th>
<th>Diode (6.5 kV, 500 A)</th>
<th>IGBT (6.5 kV, 250 A)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Series-modular</td>
<td>100</td>
<td>132</td>
</tr>
<tr>
<td>S^3RM^2</td>
<td>78</td>
<td>120</td>
</tr>
<tr>
<td>RSC converter</td>
<td>78</td>
<td>120</td>
</tr>
</tbody>
</table>

Table I and II show that the device count comparisons between the S^3RM^2 configurations and the RSC converter. It can be seen from Table I and II that the cascade S^3RM^2 compared to the series-modular S^3RM^2 requires more device count due to the higher voltage and current stresses of the switches.
TABLE II
Device count comparisons for the 150 kV DC output voltage

<table>
<thead>
<tr>
<th>Approach</th>
<th>Diode (6.5 kV, 500 A)</th>
<th>IGBT (6.5 kV, 250 A)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cascade S’RM²</td>
<td>110</td>
<td>200</td>
</tr>
<tr>
<td>RSC converter</td>
<td>84</td>
<td>120</td>
</tr>
</tbody>
</table>

Fig. 8. Normalized loss comparisons between S’RM² configurations and RSC converter. (a) Series-modular S’RM² and RSC converter. (b) Cascade S’RM² and RSC converter.

Fig. 9. Normalized weight and volume comparisons of the passive components between S’RM² configurations and RSC converter. (a) Series-modular S’RM² and RSC converter. (b) Cascade S’RM² and RSC converter.

Normalized comparisons among the converters are depicted in Fig. 8(a) and (b). The suitable high power capacitors and inductors are selected from AVX, GS-ESI, and REO [17-19]. In these comparisons, it is assumed that all the resonant capacitors have same values.

In Fig. 8(a) and (b), the passive components losses are reduced for the S’RM² configurations. The normalized total loss values indicate that the S’RM² configurations yield the system loss nearly similar to that of the RSC converter, especially in the case of series-modular S’RM². It can be observed from Fig. 9(a) and (b) that the passive component volumes and weights are significantly reduced for the S’RM² configurations compared to the RSC converter. This is because the fact that the passive component counts in the RSC converter are much larger than those of the S’RM² configurations. For example, the RSC converter uses 14 inductors and 28 capacitors, whereas the cascade S’RM² requires 8 inductors and 8 capacitors with lower current ratings in its circuit.

Furthermore, the output filter capacitor values in the RSC converter are much larger than the resonant capacitor values, while all the capacitors have the same values in the S’RM² configurations. If the same low voltage ripple is considered for all the resonant capacitors, the capacitor weights in the RSC converter will be four times heavier than those of the S’RM² owing to the higher current levels for the resonant capacitors. The physical size of power conversion systems is an important issue because they must be taken to the places that offshore wind turbines are located. Bulky and huge
electrical components represent high investment costs due to a more difficult erection, large cranes, lifting vessels, and the equipment transportation from the shore to the installation sites. Therefore, high-power-density power conversion systems can offer the opportunity to install lightweight components such as tower and platform systems for offshore wind energy systems. Conceptual comparisons of the S$^3$RM$^2$ configurations to the RSC converter show that the S$^3$RM$^2$ is superior to the RSC converter in terms of the passive component count, current rating, volume and weight.

IV. EXPERIMENTAL RESULTS

A 7-level S$^3$RM$^2$ was implemented on a 5-kW laboratory prototype converter to verify the theoretical developments presented above. The proposed converter was designed to boost a 100 V input to the maximum output voltage of 700 V. The switching frequency is about 2.75 kHz. The specifications of the capacitors and inductors are represented in Table III.

<table>
<thead>
<tr>
<th>Components</th>
<th>Symbols</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resonant inductor</td>
<td>$L_{ru}$, $L_{rd}$</td>
<td>7</td>
<td>µH</td>
</tr>
<tr>
<td>Resonant inductor</td>
<td>$L_{r2}$, $L_{rd2}$</td>
<td>14</td>
<td>µH</td>
</tr>
<tr>
<td>Output inductor</td>
<td>$L_{ruo}$, $L_{rdo}$</td>
<td>14</td>
<td>µH</td>
</tr>
<tr>
<td>Resonant capacitor</td>
<td>$C_{ru1}$, $C_{rd1}$</td>
<td>400</td>
<td>µF</td>
</tr>
<tr>
<td>Resonant capacitor</td>
<td>$C_{r2}$, $C_{rd2}$</td>
<td>200</td>
<td>µF</td>
</tr>
<tr>
<td>Output capacitor</td>
<td>$C_{uo}$, $C_{do}$</td>
<td>200</td>
<td>µF</td>
</tr>
</tbody>
</table>

Fig. 11. Operating waveforms of the proposed converter. (a) Gate-emitter voltage of switch $S_{ru}$, (b) and (c) Output capacitor currents. (d) and (e) Output inductor currents. (f) Output voltage. (g) Output capacitor voltage. It can be observed from Fig. 11(b), (c), (d), and (e) that the current waveforms closely match the analysis of (2) and (6) in Section II and the simulation results of Fig. 5(c) and (d). The peak value of the output inductor current is 22.5 A, which is consistent with the analysis of (2). The output voltage $V_o$ and the output capacitor voltage $V_{cuo}$ are shown in Fig. 11(f) and (g). They have peak to peak voltage ripples of 4 V and 10 V, respectively. The empirical voltages are in a close agreement with the simulation results of Fig. 6(b) and (c). The test results clearly show that the proposed circuit can enhance the power density and efficiency owing to the minimum switching losses.

V. CONCLUSION

A new solid-state step-up resonant Marx modulator (S$^3$RM$^2$) with a continuous output current is proposed for offshore wind energy applications. The soft-switching action is provided by the resonant condition of the circuit, thus, switching losses are minimal in both ON and OFF instants, and the power density of the system can be enhanced by increasing switching frequency. The output capacitors present a continuous output current operation during a switching period. These characteristics make the proposed converter a viable option for high voltage and high power offshore wind applications that require high power density and high efficiency. The simulation and experimental results have been performed to verify the feasibility of the proposed converter.
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