Abstract: High-voltage direct current (HVDC) is more and more often implemented for long distance electrical energy transmission, especially for off-shore wind farms. In this study, a full DC off-shore wind farm, which requires a high-power and high-voltage DC/DC converter, is considered. In order to reduce the size of the converter, the trend is to increase operating frequency. Silicon carbide (SiC) metal–oxide–semiconductor field-effect transistors (MOSFETs) are becoming industrially available and give scope for the realisation of high-performance DC/DC converters based on modular architectures. This study presents a prospective analysis of the potential of such devices in HVDC power systems. Considering the characteristics of Si insulated-gate bipolar transistor and SiC MOSFET power modules, two DC/DC converter topologies are compared in terms of losses and number of components. In conclusion, a study of the efficiency based on converter energy loss is presented.

1 Introduction

The increase in electrical energy consumption and global warming has forced countries to review their electricity production methods. Renewable energy sources such as wind and photovoltaic farms are already well implanted in Europe but many efforts are still required to meet environmental objectives.

The two main issues involving renewable energy sources are their intermittence and unpredictability. These issues complicate large-scale integration of such renewable energy sources in existing electrical grids. In order to facilitate their integration, two solutions can be found in the literature: ‘Smart Grids’ and ‘Super Grids’ [1].

The Smart Grid is commonly defined as an intelligent electrical grid coupled with a communication network. The combination of these two networks will provide an optimised energy management between production and consumption centres.

The Super Grid can be defined as a strong and powerful electricity network, allowing the transmission over long distances of large amounts of energy, from remote production areas to the consumption centres. For Europe, the combination of these two technologies will allow an optimised electrical energy exchange between countries, with lower losses and better flexibility.

In the case of long-distance overhead lines, submarine cables, high-voltage direct current (HVDC) is often economically more attractive than high-voltage alternating current (HVAC). It is the main reason why HVDC is more and more implemented in power transmission systems.

Nowadays, in off-shore wind farms, an AC collector grid is used to interconnect wind turbines and an AC–DC converter rectifies the current prior to HVDC transmission. Recently, a new architecture, based on a DC collector grid was proposed in the literature [2]. For this architecture, a high-voltage and high-power isolated DC/DC converter is required. After a brief history and state-of-the-art of HVDC power transmission, this paper introduces two suitable DC/DC converter topologies in the first section and discusses them in the second section. In the third section, by considering the characteristics of Si insulated-gate bipolar transistor (IGBT) and silicon carbide (SiC) metal–oxide–semiconductor field-effect transistor (MOSFET) power modules, the two topologies are compared in terms of energy loss. Finally, the conclusion presents the potential of SiC MOSFETs in HVDC power systems.

2 HVDC power transmission systems

2.1 History

The first commercial HVDC system was commissioned in the 1940s in Germany for a ±200 kV–60 MW transmission line based on line commutated converters (LCCs) [3]. Until the 1970s, mercury arc valves were used in HVDC power converters and began to be replaced by thyristors in the late 70s. Today's use of high-voltage and high-current thyristors allows the design of HVDC power converters able to transmit large amounts of power under ultra-high voltages and with low losses. However, since thyristors are not self-commutated devices, this technology suffers from large filtering needs and the inability to controlling the active and reactive power separately.

The arrival on the market of high-voltage and high-current self-commutated devices such as IGBTs allowed the use of voltage-source converters (VSCs) in HVDC power systems. This technology shows several advantages compared with LCC such as the independent control of active and reactive power, reduced filter components and unidirectional DC voltage which allows the use of cheaper cables. Compared with LCCs, VSCs reduce conversion-station footprints. For these reasons VSCs are frequently considered for off-shore power transmission.

2.2 VSCs in HVDC power systems

The choice of VSC topology is linked to the ratings of the power semiconductor devices. For high-voltage applications, a two-level converter using a direct series connection of devices may be considered as the simplest solution but it suffers from high switching losses and higher current and voltage harmonic distortion compared with multilevel topologies [4]. Nevertheless, such a solution was proposed commercially in 1997 for a ±150 kV–400 MW HVDC transmission system [3].

However, classical multilevel topologies such as neutral point clamped and flying cap converters are limited for this application because of the capacitors’ volumes which dramatically increase as a function of voltage [5].

In 2003, the modular multilevel converter (MMC) was presented as a new concept for high-voltage and high-power conversion [6]. This topology consists in several DC–AC elementary converters called ‘submodules’ connected in series on the AC side as it is depicted in Fig. 1. Each submodule is
composed of several switches and a capacitor. Different topologies can be implemented such as half-bridge, asymmetrical H-bridge or H-bridge [7, 8]. However, due to lower losses, the half-bridge topology is usually selected for industrial applications. In addition to its modularity, the MMC presents several advantages such as good voltage sharing for the semiconductors, insensitivity to semiconductor tolerances and parasitic circuit parameters, compared with topologies based on direct series connexion of devices [9]. Moreover, nearest level modulation (NLM) allows good voltage and current quality with no filter requirement [10].

In most cases, silicon (Si) IGBT modules are used in the submodules. In order to operate with a high efficiency, the switching frequency is low [9]. On the other hand, wide band gap semiconductors are becoming available thanks to the technological advances of the past 10 years. Compared to Si, these devices allow higher blocking voltage, higher temperature and higher switching frequency. At the moment, gallium nitride (GaN) and SiC devices show the best advances in terms of maturity and commercial availability. However, in view of their thermal properties and the complexity of manufacturing the good quality bulk substrates required for GaN vertical devices, SiC is more suitable for high-power and high-voltage applications [11].

In a very near future, high-voltage SiC MOSFETs should be industrially available as the market demand increases [12]. In HVDC power systems, the SiC MOSFET would make possible the reduction of switching losses and the increase of switching frequency, resulting in a reduction of passive components. In this paper, the potential of the SiC MOSFET is analysed and discussed in the case of an off-shore wind farm.

2.3 Offshore wind-farm architecture

In Europe, off-shore wind farms have emerged as one of the most promising renewable energy sources [13]. According to the European Wind Energy Association, 400 GW off-shore wind power should be installed by 2030 [14] and consequently, many projects are currently forecasted. The technological progress realised in the last 10 years has allowed the development of powerful wind turbines and 15–20 MW wind turbines should be achievable in the near future [15]. Thus, the trend is to build off-shore wind farms with a high-power generation capacity in the range of GWs.

The use of HVDC power transmission reduces costs, compared with HVAC [16–18]. Indeed, for submarine cables, the HVDC versus HVAC break-even distance is about 50–80 km [19]. Moreover, because the footprints of VSC stations are smaller than with LLC, VSC is generally implemented in offshore wind farms [20]. The electrical architecture of an off-shore wind farm using a VSC-based HVDC power transmission is depicted in Fig. 2a.

Actually, parallel-connected wind turbines generate a low-frequency AC voltage and are connected to AC collector platforms. Then, several transformers are used to step-up the voltage and to provide a galvanic isolation. Finally, the MMC-based AC–DC converter is used to transfer the power to the HVDC link. The AC collector-bus generally uses a medium-voltage alternating current and because of the low frequency (50–60 Hz), the transformers are bulky and require the use of dedicated platforms at considerable cost.

The use of a full DC collector may allow the removal of the AC collector platforms together with a potential capital cost reduction [2]. The diagram of this solution is presented in Fig. 2b. It is proposed that the DC/DC converter has to provide a galvanic isolation with a high-voltage ratio [21], which requires the use of transformers.

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**Fig. 1** Single-phase circuit diagram of the MMC

**Fig. 2** Off-shore wind-farm architectures

(a) AC off-shore wind-farm architecture, (b) Full DC off-shore wind-farm architecture
the AC side by a step-up transformer, as presented in Fig. 3. The proposed architecture would be of interest if one of the referred to the primary side of the transformer two DC/DC converter topologies are compared. The first topology is based on MMCs while the second is more futuristic and is based on the association of elementary isolated DC/DC converters. Consequently, from both side of the transformer, the power factor is increased over the entire power range. In this case, the RMS value of the AC current is computed with (2) while, the phase shifts between the AC voltages and the AC current are given by (3)

\[ P = \frac{3V_1V_2}{2\pi L_f f_{ac}} \sin(\Phi) \]  

\[ I_{ac} = \frac{V_1}{2\pi L_f f_{ac}} \sin\left(\frac{\Phi}{2}\right) \]  

\[ \begin{align*} 
\Phi_1 &= \Phi \\
\Phi_2 &= -\frac{\Phi}{2}
\end{align*} \]

This topology allows the power to be transferred in both directions, according to the sign of phase angle \( \Phi \). Since the AC link is internal to the DC/DC converter, the frequency can be increased in order to reduce the size of the passive components (transformer and capacitor of the submodules). Nevertheless, the use of a unique high-power transformer limits the increase of the frequency. Indeed, the increase of the power density complicates the cooling of the transformer [27–30], but such a study exceeds the scope of this paper.

### 3.2 Association of isolated DC/DC converters

An alternative solution to the front-to-front MMC topology is an association of elementary isolated DC/DC converters in accordance with Fig. 6. With this solution, elementary DC/DC converters are connected using two types of connection: input-series–output-series (ISOS) and input-parallel–output-series (IPOS). The control of the converter enables the transmitted power to be equally shared by the elementary DC/DC converters. Consequently, the voltage and the power ratings of each elementary converter are reduced, allowing the use of medium frequency transformers (MFTs) and SiC MOSFETs. The combination of both technologies reduces the size of passive components without an important increase in switching losses. To realise the elementary converters, several solutions are possible. However, the scope of our paper is limited to topologies using a single-phase medium-frequency transformer, as presented in Fig. 7.

The dual active bridge (DAB) topology consists of two voltage-source inverters linked on the AC side by an inductive device [31]. In order to optimise the converter layout, no external inductance is implemented in the AC link. Consequently, only the leakage inductance of the transformer is used. Several modulation techniques exist but the simplest is phase-shifted modulation [32].

![Fig. 3 Front-to-front-connected MMCs DC/DC converter topology](image)

**Fig. 3** Front-to-front-connected MMCs DC/DC converter topology

![Fig. 4 Single-phase equivalent circuit of the front-to-front MMC topology referred to the primary side of the transformer](image)

**Fig. 4** Single-phase equivalent circuit of the front-to-front MMC topology referred to the primary side of the transformer

### 3 DC/DC converter topologies

The proposed architecture would be of interest if one of the following criteria is fulfilled: lower cost, higher reliability and availability or higher efficiency compared with the AC collector architecture. In this paper, only the efficiency is considered. The key component of the DC collector architecture is the DC/DC converter which is the major contributor to energy loss. In this part, two DC/DC converter topologies are compared. The first topology is based on MMCs while the second is more futuristic and is based on the association of elementary isolated DC/DC converters.

#### 3.1 Front-to-front MMCs

For the DC/DC conversion, on the basis of the MMC topology many variant structures have been proposed in the scientific literature [22–26]. However, when galvanic separation between the input and output terminals is required, front-to-front connected MMCs appear as the most natural implementation to realise a step-up converter. In this structure, two AC–DC MMCs are linked on the AC side by a step-up transformer, as presented in Fig. 3.

The transformer provides the galvanic separation and adapts the voltage and current levels between the primary and secondary sides, so that the MMCs operate with a modulation index close to one. Due to the different voltage and current levels on the primary and secondary sides, the number of submodules is different for the two MMCs. From the system point of view, each phase of the front-to-front MMCs topology can be seen as two AC voltage sources linked by the leakage inductance of the transformer, as depicted in Fig. 4. This circuit is referred to the primary side of the transformer (medium-voltage level).

Several modulation techniques can be implemented to control the MMC but the NLM shows the best efficiency compared with the two-level modulations [25, 26] and it is selected for our study. This modulation allows the reduction of harmonics in the AC link. The vector diagrams presented in Fig. 5 are based on the fundamental waveforms and can be used to compute the parameters of the circuit (currents, phase shifts etc.).

Neglecting the losses in the transformer, the transmitted power is given by (1), where \( V_1 \) and \( V_2 \) are the root-mean-square (RMS) values of the phase voltage, referred to the primary side of the transformer, \( f_{ac} \) the voltage fundamental frequency and \( \Phi \) is the phase shift between \( V_1 \) and \( V_2 \)

\[ P = \frac{3V_1V_2}{2\pi L_f f_{ac}} \sin(\Phi) \]  

The comparison of Figs. 5a and b clearly shows that the operation with \( V_1 = V_2 \) reduces the phase shift \( (\Phi_1 \text{ and } \Phi_2) \) between the AC voltages and the AC current. Consequently, from both side of the transformer, the power factor is increased over the entire power range. In this case, the RMS value of the AC current is computed with (2) while, the phase shifts between the AC voltages and the AC current are given by (3)

\[ I_{ac} = \frac{V_1}{2\pi L_f f_{ac}} \sin\left(\frac{\Phi}{2}\right) \]  

\[ \begin{align*} 
\Phi_1 &= \Phi \\
\Phi_2 &= -\frac{\Phi}{2}
\end{align*} \]
Using this modulation, each inverter provides a two-level, 50% duty-cycle voltage waveform and the power is controlled with the phase shift (Φ) between the two AC voltages.

The addition of the series capacitor in the AC link and the operation with a switching frequency close to the resonance frequency allows the generation of a quasi-sinusoidal current waveform in the transformer which reduces winding and core losses. Beyond that, this operating mode allows a power factor improvement and lower turn-off currents. By considering only the fundamental frequency, the power is expressed by

\[ P = \frac{V_{in}^2}{X_{ac}k \cdot \sin(\Phi)} \]  

(4)

where \( k \) is the voltage ratio referred to the primary side of the transformer, defined by (5) and \( X_{ac} \) is the reactance of the resonant circuit at the fundamental frequency, expressed by (6)

\[ k = \frac{V_{out}}{mV_{in}} \]  

(5)

\[ X_{ac} = \frac{2\pi f_s L_r}{2 - \frac{1}{2\pi f_s C_r}} \]  

(6)

To study the soft-switching operation of the DAB converter, the same vector diagrams as Fig. 5a can be used. For the primary inverter, if current \( I_{ac} \) lags voltage \( V_1 \) (\( \Phi_1 > 0 \)), the current flowing through the switches before turn-off is always positive which is the condition to ensure a zero-voltage switching (ZVS) operation. For the secondary inverter, ZVS operation is obtained if current \( I_{ac} \) leads voltage \( V_2 \) (\( \Phi_2 < 0 \)). Consequently, by studying the sign of Φ, the condition to ensure a ZVS operation for all switches can be found

\[ \Phi \geq \cos^{-1}(k) \text{ if } k \leq 1 \]
\[ \Phi \geq \cos^{-1}\left(\frac{1}{k}\right) \text{ if } k > 1 \]  

(7)

This condition is only valid for an operation above the resonant frequency. According to (7), ZVS is guaranteed over the entire power range for all the switches, if the voltage ratio \( k \) equals 1.

The second considered topology is the series resonant converter (SRC). It consists of one voltage-source inverter which supplies a diodes bridge through a transformer and a resonant circuit, as presented in Fig. 7b. With this topology, the voltage-source inverter provides a 50% duty-cycle two-level voltage. Several AC current waveforms and conduction modes can be obtained according to the switching frequency, the resonant frequency and the output voltage [33]. In this paper, only the operation above the resonant frequency is considered. Consequently, the converter operates in the continuous conduction mode over the entire power range. This conduction mode enables ZVS for all the primary switches. The diode bridge at the secondary side does not allow power reversal and the output voltage is lower than the input voltage according to the following equation [34]:

\[ V_{in} = V_{out} + \frac{\pi X_{ac}^2}{8} (mI_{out})^2 \]  

(8)

However, if power reversal is required, the two inverters may be equipped with SiC MOSFETs whereas only one inverter could be controlled. Moreover, since the power to be transmitted to the wind turbines is lower than the one generated by the wind turbines, the current rating of the SiC MOSFETs of the secondary inverter can be reduced.

Usually, the output voltage \( V_{out} \) is controlled by varying the switching frequency or the phase shift between the inverter legs [33]. In our study case, we assume that the output voltage is imposed by the HVDC line while the wind farm provides the power to be transmitted to shore. Consequently, the DC/DC converter is controlled at a fixed frequency and operates like a classical AC transformer with a low-voltage drop. Thus, the input voltage \( V_{in} \) varies according to the output current as shown in

Fig. 5 Front-to-front MMC topology – vector diagrams at the fundamental frequency and referred from the primary side of the transformer

(a) operation with \( V_1 \neq V_2 \), (b) operation with \( V_1 = V_2 \)

Fig. 6 MVDC/HVDC topology associating isolated elementary DC/DC converters

Using this modulation, each inverter provides a two-level, 50% duty-cycle voltage waveform and the power is controlled with the phase shift (Φ) between the two AC voltages.
Fig. 8. However, in order to balance the voltage and the current between the elementary DC/DC converters, the switching frequency may be adapted in a small range, as was proposed in [35]. In [36], three DC/DC converter topologies were compared in terms of losses. Considering a 10 kV/200 A full SiC power module and a MFT, the study took into account the losses in the switches (conduction and commutation) and in the transformers (core and windings). Mainly thanks to the low losses on the diode rectifier side, the SRC showed the best efficiency, followed by the DAB. For this reason, these two topologies are considered in this study.

4 Comparison of DC/DC converter topologies

In this section, the front-to-front connected MMCs and cascaded DC/DC converter solutions are compared in terms of losses and number of components. A full DC offshore wind-farm application is considered. The specification of the DC/DC converter is presented in Table 1.

<table>
<thead>
<tr>
<th>Table 1</th>
<th>Specification of the DC/DC converter</th>
</tr>
</thead>
<tbody>
<tr>
<td>input voltage, $V_{\text{MVDC}}$</td>
<td>40 kV</td>
</tr>
<tr>
<td>output voltage, $V_{\text{HVDC}}$</td>
<td>320 kV</td>
</tr>
<tr>
<td>nominal power, $P_{\text{dc}}$</td>
<td>600 MW</td>
</tr>
</tbody>
</table>

For the front-to-front MMCs topology, regarding the high-power requirement of the transformer, the classical Si-Fe technology should be used. Consequently, in order to keep a good efficiency, we assume that the fundamental frequency has to be lower than 350 Hz. In order to study the potential of SiC MOSFETs, two configurations, presented in Table 2, are considered.

For the elementary DC/DC converters, the use of a medium frequency AC link allows the size of the transformer to be reduced. However, to allow such a switching frequency, specific designs of the MFTs have to be performed and a lot of technological challenges have to be addressed [37]. Given the literature on this topic, the switching frequency of the DC/DC converters was set to 10 kHz [38–40].

4.1 Assumptions and input data

For the calculations, three types of switches are considered. For the front-to-front MMCs solution, a 3.3 kV/1.5 kA single switch IGBT power module is considered since a low switching frequency is used [41]. As far as SiC MOSFETs are concerned, currently, there are no high-voltage power modules commercially available, only dies can be obtained. Thus, the loss models of full SiC MOSFET power modules were extrapolated from die datasheet and considering 30 dies in parallel for a maximum junction temperature of 125°C. However, a recent publication confirmed our estimations [42]. Table 3 shows the characteristics of the considered power modules.

The conduction losses are estimated using the following equation:

$$ P_{\text{con}} = \frac{1}{2} V_i I_i r_d $$

where $P_{\text{con}}$ is the conduction loss, $V_i$ and $I_i$ are the input voltage and current, and $r_d$ is the conduction resistance.

$$ P_{\text{rec}} = \frac{1}{2} V_i I_i E_{\text{rec}} $$

where $P_{\text{rec}}$ is the reverse recovery loss, $E_{\text{rec}}$ is the reverse recovery voltage, and $r_d$ is the reverse recovery resistance.

$$ P_{\text{off}} = \frac{1}{2} V_i I_i t_{\text{off}} c_{\text{off}} $$

where $P_{\text{off}}$ is the switching loss, $t_{\text{off}}$ is the turn-off time, and $c_{\text{off}}$ is the turn-off loss coefficient.

$$ P_{\text{on}} = \frac{1}{2} V_i I_i t_{\text{on}} c_{\text{on}} $$

where $P_{\text{on}}$ is the switching loss, $t_{\text{on}}$ is the turn-on time, and $c_{\text{on}}$ is the turn-on loss coefficient.

Table 2 | Configurations for the front-to-front MMC solution |
<table>
<thead>
<tr>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Fundamental frequency, Hz</td>
<td>50</td>
</tr>
<tr>
<td>Switching frequency, Hz</td>
<td>250</td>
</tr>
<tr>
<td>Semiconductors</td>
<td>Si IGBTs and SiC MOSFETs</td>
</tr>
</tbody>
</table>

Table 3 | Characteristics of the considered power modules |
<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Static</td>
<td>Dynamic</td>
<td>Static</td>
</tr>
<tr>
<td>3.3 kV Si IGBT power module at 1.8 kV</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$v_0 = 1$ V</td>
<td>$a_{\text{on}} = 4 \times 10^{-7}$; $b_{\text{on}} = 0.0008$; $c_{\text{on}} = 0.5$</td>
<td>$v_0 = 0.5$ V</td>
<td>$a_{\text{rec}} = -2 \times 10^7$</td>
</tr>
<tr>
<td>$r_d = 1.6$ mΩ</td>
<td>$a_{\text{off}} = 1 \times 10^{-7}$; $b_{\text{off}} = 0.0012$; $c_{\text{off}} = 0.1481$</td>
<td>$r_d = 1.5$ mΩ</td>
<td>$b_{\text{rec}} = 0.0011$; $c_{\text{rec}} = 0.4119$</td>
</tr>
</tbody>
</table>

| 3.3 kV SiC MOSFET power module at 1.8 kV | | | | |
| $v_0 = 0$ V | $a_{\text{off}} = 3 \times 10^{-8}$; $b_{\text{off}} = -1 \times 10^{-5}$; $c_{\text{off}} = 0.009$ | $v_0 = 0.7$ V | $r_d = 2.63$ mΩ |
| $r_d = 2.8$ mΩ | $a_{\text{off}} = 3 \times 10^{-8}$; $b_{\text{off}} = -1 \times 10^{-5}$; $c_{\text{off}} = 0.009$ | $r_d = 2.63$ mΩ | $E_{\text{rec}} = 0$ |

| 10 kV SiC MOSFET power module at 5.5 kV | | | | |
| $v_0 = 0$ V | $a_{\text{off}} = -3.6 \times 10^{-7}$; $b_{\text{on}} = 642 \times 10^{-6}$; $c_{\text{on}} = 0.0383$ | $v_0 = 1.3$ V | $r_d = 20$ mΩ |
| $r_d = 25$ mΩ | $a_{\text{off}} = 7.1 \times 10^{-9}$; $b_{\text{off}} = 1.4 \times 10^{-5}$; $c_{\text{off}} = 0.0459$ | $r_d = 20$ mΩ | $E_{\text{rec}} = 0$ |
where \( <i_q> \) is the average current and \( I_q \) is the RMS current flowing through the switch, \( v_0 \) is the threshold voltage (\( v_0 \) equals zero for unipolar devices such as MOSFETs) and \( r_d \) is the dynamic resistance of the switch. The switching losses (turn-on, turn-off and recovery losses) are estimated using a second-order polynomial according to the following equation:

\[
E_{\text{switch/rec}} = E_{\text{on/off}} + E_{\text{on/off-rec}} + E_{\text{on/off-toc}}
\]

(10)

The maximum operating voltage is 1.8 kV for the 3.3 kV devices (Si IGBTs and SiC MOSFETs) and 5.5 kV for the 10 kV SiC MOSFETs. The use of SiC diodes allows us to neglect recovery losses, which is not the case with Si diodes. In our calculations, we consider that the maximum turn-off current is 1.3 kA for the Si IGBT [41]. For a case temperature of 80°C, the maximum dissipated power per module was fixed at 2.5 kW, as allowed by a junction-case thermal resistance of 18°C/kW. As for the SiC MOSFET, we considered a maximal turn-off current of 2.7 kA at 1.8 kV for the 3.3 kV device and 1.2 kA at 5.5 kV for the 10 kV device.

Due to the very specific technology of the transformers [27–30], the estimation of their losses exceeds the scope of this paper and is not considered. A specific study on transformers will be required to complete our investigations.

### 4.2 Sizing of DC/DC converters

Based on the power module characteristics and according to the converter specification in Table 1, the two solutions have been sized according to the following criteria:

i. Operation of the switches within their ratings.

ii. Efficiency of the converter higher than 99% (transformer losses are not taken into account).

iii. Minimal number of components.

For the front-to-front MMCs, the modulation index is set to 0.85 and the leakage reactance of the transformer to 0.2 pu, i.e. 0.7 Ω. The configuration of each MMC is presented in Table 4.

Due the voltage and power ratings, the DC and AC currents are important and require the use of several arms in parallel per phase, as is depicted in Fig. 9, for each MMC. In reality, the paralleling can be obtained by adding several power modules in parallel in the submodule or by using several DC/DC converters with lower power ratings. For the Si IGBT power module, the limited safe operating area forces us to reduce the arm current and to add more branches in parallel as compared to SiC MOSFETs.

For MMCs, the required submodule capacitance and branch inductance values [6] are given by (11) and (12), where \( \Delta V_c \) and \( \Delta I_l \) are, respectively, the ripple in the submodule capacitor voltage and the branch inductor current

\[
P_{\text{cond}} = <i_q> v_0 + I_q^2 r_d
\]

(9)

\[
C = \frac{I_c}{2 \omega_c \Delta V_c}
\]

(11)

\[
L > \frac{V_{dc}}{8 \pi f_{sw} \Delta I_l}
\]

(12)

Usually, the branch inductance value has to be increased in order to limit the current slope during an AC fault. In the case of the DC/DC converter, since the AC voltage is controlled on both sides of the transformer, an over-rating of the inductance is not required.

For the cascaded DC/DC converter solution, the resonant frequency was set to 9.5 kHz. Consequently, 3.3 kV Si-IGBT power modules were not considered because they do not allow operation at such a high switching frequency, with a good efficiency. Table 5 presents the different configurations.

In view of the high voltage and power ratings, several full DC/DC converters (see Fig. 6) have to be connected in input-parallel–output-parallel (IPOP) configuration, as presented in Fig. 10.

In the SRC topology, using 3.3 kV switches, the high amplitude of the AC current in the series impedance causes an important rise of the input voltage. Consequently, the power of each elementary converter has to be reduced in order to limit the input voltage

### Table 4: Sizing of the front-to-front MMC solutions

<table>
<thead>
<tr>
<th>Power Module</th>
<th>Submodules per branch</th>
<th>Primary MMC</th>
<th>Peak current in each arm</th>
<th>Secondary MMC</th>
<th>Total number of submodules</th>
</tr>
</thead>
<tbody>
<tr>
<td>3.3 kV Si IGBT power module</td>
<td>23</td>
<td>16</td>
<td>1.24 kA</td>
<td>2</td>
<td>1.24 kA</td>
</tr>
<tr>
<td>3.3 kV SiC MOSFET power module</td>
<td>23</td>
<td>8</td>
<td>2.5 kA</td>
<td>2</td>
<td>1.24 kA</td>
</tr>
<tr>
<td>10 kV SiC MOSFET power module</td>
<td>8</td>
<td>30</td>
<td>667 A</td>
<td>3</td>
<td>833 A</td>
</tr>
</tbody>
</table>
variations. The sizing and the operation point allow ZVS operation for all the topologies implemented in the elementary DC/DC converters. Consequently, only the turn-off losses are considered. For the elementary DC/DC converter the values of the DC filtering capacitors are computed with (13). Capacitor current $i_c$ is assumed to be the AC part of the input or output current of the converter.

$$C = \frac{1}{\Delta V_c} \int_0^{T_{ac}/2} i_c(t) dt$$ \hspace{1cm} (13)

Table 5 compares the two solutions based on the number of components and the energy stored in the passive components. Voltage and current ripples are 10% of the average value for the front-to-front MMCs solution and 5% of the average value for the cascaded DC/DC converter solution. In this last solution, the frequency of the capacitor currents is twice the switching frequency. In our case, a switching frequency of 10 kHz allows an important reduction of total energy stored in the capacitors. Moreover, the energy stored in the passive components (inductor and capacitor) of the elementary DC/DC converters’ resonant circuit is negligible compared with the energy stored in the DC filtering capacitors. Consequently, it has not been taken into account in the energies given in Table 6.

4.3 Energy loss estimation

In order to compare the two solutions, the semiconductor losses were estimated. For the MMC, the values given in Table 6 were considered and the efficiency was calculated on the basis of the analytical approach proposed in [6]. For the cascaded DC/DC converter, the calculation of the instantaneous current in the series resonant circuit was based on the analytical solution of the differential equation and the sequential analysis of the converter operation. The results given by this method were confirmed by using PLECS simulation software [43]. Knowing the instantaneous current in the AC link, the current in the switches is easily calculated. RMS and average currents through the switches are calculated. Then, relations (9) and (10) with the parameters given in Table 3 are used to compute the losses. Fig. 11 shows the loss distribution at different power levels for both solutions. Compared with Si IGBTs, SiC MOSFETs allow a significant reduction of the

Table 5 Sizing of the cascaded DC/DC converter solutions

<table>
<thead>
<tr>
<th>Topology</th>
<th>Nominal power of each elementary converter, MW</th>
<th>Number of elementary converters in ISOS</th>
<th>Number of elementary blocs in IPOS</th>
<th>Number of full DC/DC converter in IPOP</th>
<th>Total number of elementary converters</th>
</tr>
</thead>
<tbody>
<tr>
<td>3.3 kV SiC MOSFET</td>
<td>DAB</td>
<td>1.1</td>
<td>23</td>
<td>8</td>
<td>3</td>
</tr>
<tr>
<td>power module</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>3.3 kV SiC MOSFET</td>
<td>SRC</td>
<td>0.96</td>
<td>23</td>
<td>9</td>
<td>3</td>
</tr>
<tr>
<td>power module</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>10 kV SiC MOSFET</td>
<td>DAB and SRC</td>
<td>2</td>
<td>5</td>
<td>8</td>
<td>8</td>
</tr>
</tbody>
</table>

Table 6 Comparison of the front-to-front MMCs and cascaded DC/DC converter solutions

<table>
<thead>
<tr>
<th>Topology</th>
<th>Number of submodules/elementary DC/DC converters</th>
<th>Number of IGBT/MOSFET power modules</th>
<th>Number of diode power modules</th>
<th>Total energy in the capacitors</th>
<th>Total energy in the inductors</th>
</tr>
</thead>
<tbody>
<tr>
<td>Front-to-front MMC</td>
<td>3.3 kV SiC MOSFET</td>
<td>3240</td>
<td>6480</td>
<td>70.5 MJ ($f_{ac} = 50$ Hz)</td>
<td>10.5 kJ ($f_{ac} = 50$ Hz)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>10.2 MJ ($f_{ac} = 350$ Hz)</td>
<td>0.63 kJ ($f_{ac} = 350$ Hz)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>84 MJ ($f_{ac} = 60$ Hz)</td>
<td>6.4 kJ ($f_{ac} = 60$ Hz)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>12.1 MJ ($f_{ac} = 350$ Hz)</td>
<td>0.91 kJ ($f_{ac} = 350$ Hz)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>71.1 MJ ($f_{ac} = 50$ Hz)</td>
<td>103.7 kJ ($f_{ac} = 50$ Hz)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>10.2 MJ ($f_{ac} = 350$ Hz)</td>
<td>15 kJ ($f_{ac} = 350$ Hz)</td>
</tr>
<tr>
<td>Cascaded DC/DC converter</td>
<td>3.3 kV SiC MOSFET</td>
<td>2856</td>
<td>5712</td>
<td>131.8 kJ (DAB)</td>
<td>117.1 kJ (DAB)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>621 (SRC)</td>
<td>197.6 kJ (SRC)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>288 (DAB)</td>
<td>112.8 kJ (DAB)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>288 (SRC)</td>
<td>1152 (SRC)</td>
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</tbody>
</table>

Fig. 10 IPOP connection of full DC/DC converters
switching losses but show higher conduction losses when the power level increases.

Fig. 12 presents, for the front-to-front MMC solution, the efficiency dependence on transmitted power. It should be noticed that SiC-MOSFETs-based converters allow a significant increase of the efficiency except at the maximum power level.

Fig. 13 compares the SiC MOSFET based cascaded DC/DC converter and highlights that the SRC is more efficient at the lower power levels.

Finally, Fig. 14 compares the efficiency variation for the both solution equipped with SiC MOSFETs.

With 3.3 kV SiC MOSFETs, the efficiency of the front-to-front MMCs solution is higher than the cascaded DC/DC converter over 60% of the power range. Since the switching losses of the 10 kV SiC MOSFETs are higher, the efficiency of the converter is more dependent on the operating frequency. Consequently, the front-to-front MMCs operating at 50 Hz presents the best efficiency over the entire power range.
4.4 Efficiency considerations

From the previous analysis, it is difficult to conclude on the most efficient topology since the answer depends on the transmitted power level. Nevertheless, since a wind-farm converter does not operate at constant power, it is more convenient to determine the total energy loss over 1 year to find the most efficient solution. For off-shore wind farms, the Weibull distribution is used in order to characterise the wind distribution over a given period [44, 45]. Thus, knowing the efficiency of the converter, the energy loss over a given period can be estimated. The Weibull wind probability distribution is given in the following equation:

\[ P(v) = \frac{K}{A} \left( \frac{v}{A} \right)^{K-1} e^{-\left( \frac{v}{A} \right)^{K}} \]  

where \( v \) is the wind velocity, \( K \) is the shape factor and \( A \) is the scale factor of the distribution. In this paper, a shape factor of 2.2 and a scale factor of 10.57 m/s were used [46]. Fig. 15 presents the corresponding wind distribution.

Knowing the power profile of the wind turbine and the efficiency of the DC/DC converter, the losses versus the wind velocity can be calculated. The power profile of the farm based on 100, 6 MW wind turbines presented in Fig. 15b allows us to calculate the energy loss per year using the following equation:

\[ W_{\text{lost}} = 24 \cdot 365 \cdot \int_{0}^{\infty} P(v) \gamma(v) dv \]  

where \( \gamma(v) \) is the energy loss versus the wind velocity based on the efficiency of the DC/DC converter and \( P(v) \) is the power profile of the wind turbines. This calculation is done assuming 365 days (24 h), that is, an 8760 h operation. Fig. 16 presents the energy loss per year for each solution.

In Fig. 16, it can be noticed that the use of the 3.3 kV SiC MOSFETs reduces energy loss. In the case of the front-to-front MMCs solution at 50 Hz, a significant energy saving is obtained compared with the use of Si IGBTs. For example, considering an energy cost of 150 $/MWh, the money saving reaches 1.2 M$ each year. This gain is even increased with the cascaded DC/DC converter solution. However, the cost-benefit analysis must take into account the price of the SiC MOSFET power modules, which is much higher than that of Si IGBT power modules. However, this price difference should diminish in the coming years due to ongoing developments. Nevertheless, the cascaded DC/DC-converter solution allows the elementary converters to be geographically distributed throughout the farm which, coupled with the higher power density of the converters, would permit a reduction of the footprint of off-shore platforms.

5 Conclusion

In this paper, a comparative analysis of two high-power and high-voltage isolated DC/DC converters was developed for an off-shore wind farm. The first topology is based on an association of two MMCs including a single step-up transformer. The second one, which requires SiC devices, is based on series and parallel associations of elementary DC/DC converters operating at 10 kHz.

The comparison was based on the semiconductor losses and for both topologies, it was shown that SiC MOSFETs allow a significant energy saving. However, regarding the maturity of SiC components, only the front-to-front MMCs solution using IGBTs, which requires a single transformer operating at low frequency, seems accessible.

On the other hand, by reducing the operating voltage and the power rating of the transformer, the frequency can be increased. This is the aim of the cascaded DC/DC converter solution which
allows the option of reducing the off-shore platform area. Nevertheless, regarding the MFT, a feasibility study should be carried out. The key issue concerns the high-voltage isolation requirement due to the series connection of the elementary DC/DC converters [47, 48], and thermal aspects due to the higher power density [27–30, 37]. Regarding these technological challenges, it is still too early to determine if the use of SiC MOSFETs and MFTs would have an economical interest in off-shore wind farms. However, the results presented in this paper encourage further work on these new technologies.

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7 References

Fig. 16 Converter energy loss per year for each solution

