An Innovative Bidirectional Isolated Multi-Port
Converter with Multi-Phase AC Ports and DC Ports

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An Innovative Bidirectional Isolated Multi-Port Converter with Multi-Phase AC Ports and DC Ports

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Abstract—This paper presents an innovative bidirectional isolated multi-port converter with multi-phase AC ports and DC ports, which is a key element of Solid-State Transformers (SST) utilized for example in a wind energy generation system. The multi-port converter allows the direct coupling of the three-phase AC system of the power generator with the three-phase AC utility grid and an additional DC storage unit applying a single high-frequency transformer structure. The topology is stackable and hence, single converter modules can be connected in series at the input/output ports for medium or high voltage or in parallel for low voltage applications. The converter is operated utilizing a time-varying phase-shift control to draw or inject sinusoidal currents with a corresponding amplitude and phase at the AC ports. The topology, its operating principle including the theoretical analysis and simulation results of a prototype system are provided.

Keywords—Multi-Port Converter, AC-DC Converter, Multi-Phase AC Port, Bidirectional, Isolated

I. INTRODUCTION

The use of renewable energy sources is constantly increasing in order to replace limited energy sources like coal, oil or uranium to reduce greenhouse gas emissions and account for the exhaustible primary energy sources. Wind energy represents an important part of today’s renewable energy generation which demands suitable high power electronic equipment interfacing the generator, energy storage systems and the utility grid. For wind energy generation systems, different wind turbine concepts exist [1], [2]: For instance fixed speed wind turbines employing asynchronous squirrel cage induction generators or partial variable speed wind turbines with variable aerodynamic rotor resistance where both concepts show a direct connection of the generator and the grid via a transformer. Concepts with variable speed wind turbines show either a partial-scale (also known as the doubly-fed induction generator concept) or a full-scale AC-AC power converter with a subsequent low-frequency transformer to connect to the utility grid.

The partial- or full-scale AC-AC power converter used in the latter two concepts usually consists of two stages, a three-phase AC-DC rectifier and a subsequent three-phase DC-AC inverter [3]. The common DC-link additionally allows the connection of energy storage systems [4] to compensate for energy shortages during low wind conditions. Besides unidirectional power converter solutions employing a cost-efficient diode rectifier on the generator side, several bidirectional single-cell and multi-cell power converters have been proposed [5]. Single-cell power converters include two-level and multi-level converters in back-to-back configuration like diode-clamped or flying-capacitor multi-level converter [6].

Multi-cell power converters comprise the modular multilevel converter [7], [8] in back-to-back configuration [9] and the cascaded H-bridge converter [8], where additional dual-active-(full-)bridge or dual-half-bridge modules form an additional isolation stage applying a high-frequency transformer [5], [10].

In this paper, an innovative bidirectional isolated multi-port converter with multi-phase AC ports and DC ports for use as a full-scale power converter in a wind energy generation system...
as shown in Fig. 1 is proposed. The multi-port converter allows the direct coupling of a three-phase AC system of a power generator with the three-phase AC utility grid and an additional DC storage unit applying a single high-frequency transformer structure. The single-stage power conversion leads to a low number of switching devices and gate drives. Due to the integrated transformers providing isolation and voltage adaptation, the low-frequency transformer on the grid side is fully eliminated which in turn saves volume and weight. Furthermore, compared to three-phase converter solutions, the proposed multi-port converter is stackable and hence, single converter modules can be connected in series or parallel at the input/output ports for medium or high voltage applications and properly scaled according to voltage and power needs at the ports. The operating principle allows to control the apparent power at the AC ports independently from each other, the power at the DC port is then given inherently. Additionally, the proposed converter concept offers the possibility to couple the power generator directly to a Medium or High Voltage DC (MVDC/HVDC) distribution grid.

In the following, first the converter topology and especially the introduction of multi-phase AC ports is presented in section II. Then, section III shows how the operating principle comprising the modulation and control strategy in terms of a mathematical analysis. Finally, simulation results of a prototype system are provided in section IV.

II. CONVERTER TOPOLOGY

Fig. 1 shows the proposed multi-port converter in a wind energy generation system. The converter topology consists of AC port switching networks $S_a, S_b, S_c$ which connect to a first three-phase AC system with phases $a, b, c$ and the star point $N_{abc}$ and AC port switching networks $S_A, S_B, S_C$ which connect to a second three-phase AC system with phases $A, B, C$ and the star point $N_{ABC}$. The frequencies and the phases of the two three-phase AC systems do not have to be equal. Furthermore, a DC port switching network $S_{dc}$ is attached to a DC-link. All of the switching networks are coupled with each other through six two-winding transformers $T_a, T_b, T_c, T_{AB}, T_{BC}, T_{CA}$ where the three secondary windings of the transformers belonging to one three-phase AC system form a series interconnection.

The switching networks in Fig. 1 consist of half-bridges with a clamping switch (also known as T-type circuit) with bidirectional switching devices when coupled to a low-frequency (LF) AC port and of full-bridges with unidirectional switching devices when coupled to a DC port. Each switching network applies a high-frequency (HF) square-wave voltage with or without clamping interval to the corresponding winding or the series connection of windings with an amplitude equal to half of the voltage which occurs on the LF AC side or the full DC voltage on the DC side and a phase angle in relation to a chosen reference. The switching frequency is chosen to be well above the frequencies of the two three-phase AC systems and the capacitors assumed to be large enough, so that the amplitudes of the generated square-wave voltages can be considered as constant during one switching cycle.

For simplicity, in the following investigations, only one part of the multi-port converter which is depicted in Fig. 2 and represents the connection of the generator side with the DC-link is analyzed. The analysis of the grid side of the converter can be done in a similar way.

A. Multi-Phase AC Ports

Bidirectional isolated multi-port converters in literature usually exhibit three DC ports which are coupled through a three-winding transformer and are controlled by the phase-shifts between the square-wave voltages applied to the windings [11]–[13]. To achieve high efficiency in terms of low switching and conduction losses, the DC port voltages are kept at mainly constant voltages or additional duty cycle control is introduced to compensate for voltage variations [11], [12]. In [12], Zero-Voltage-Switching (ZVS) conditions are based on keeping the half-cycle voltage-second products (half-cycle voltage-time integrals) applied to the windings equal. In the case of an AC-DC two-port converter employing a primary half-bridge, whose voltage varies in time, with the use of a secondary fullbridge, the voltage-second product applied to the secondary winding can be adjusted to the one of the primary side.

If a multi-phase AC system with phase voltages, whose absolute values add up to a nearly constant sum over time, is considered, a multi-phase AC port is formed which can be coupled directly to a DC port through transformer structures shown in Fig. 3. An equivalent circuit of the converter is given in Fig. 4a. Hence, for controlling the converter, a nearly constant sum of the half-cycle voltage-second products applied to the primary windings is available over the whole AC system period. By using this nearly constant control variable, which is further described in section III, the converter efficiency can be kept high.

The term nearly constant in this context does not mean negligible ripple. In case of a symmetrical three-phase AC system, the sum of the absolute phase voltages with amplitude $V_{abc}$ corresponds to a six-pulse wave form with a constant average value $6V_{abc}/\pi$ over the AC system period exhibiting a $2(\pi/3 - 1)\% = 9.4\%$ peak-to-peak voltage ripple.
B. Transformer Structure for Port Coupling

The use of the nearly constant sum of the absolute values of the phase voltages requires a transformer structure where the HF voltages applied to the windings are added. Such a structure comprises for instance three two-winding transformers as depicted in Fig. 3a. There, the voltages \( v_{p,a}, v_{p,b}, v_{p,c} \) are summed up through the series interconnection of the secondary windings. Magnetically, the sum of the voltages corresponds with the sum of the winding fluxes. Therefore, the voltage sum can be translated into a winding flux sum which leads to a single four-winding transformer shown in Fig. 3b where the core winding fluxes are summed up through the secondary winding.

In Fig. 3a, the voltages \( v_{p,a}, v_{p,b}, v_{p,c} \) applied to the primary windings impress the primary winding fluxes \( \Phi_{p,a}, \Phi_{p,b}, \Phi_{p,c} \) according to Faraday’s Law. The voltage \( v_{s,dc} \) applied to the series connection of the secondary windings predefines the flux sum \( \Phi_{s,a} + \Phi_{s,b} + \Phi_{s,c} \). The mismatch of the primary and secondary winding fluxes per core results in the leakage fluxes \( \Phi_{s,a}, \Phi_{s,b}, \Phi_{s,c} \) which show the same value for negligible magnetizing fluxes if all secondary windings have the same number of turns. Hence, all of the windings exhibit the same leakage inductance current referred to a specific winding. The three transformers show the primary referred leakage inductances \( L_{\sigma,a}, L_{\sigma,b}, L_{\sigma,c} \).

In case of the single four-winding transformer depicted in Fig. 3b, the voltages \( v_{p,a}, v_{p,b}, v_{p,c} \) applied to the primary windings impress the winding fluxes \( \Phi_{p,a}, \Phi_{p,b}, \Phi_{p,c} \) whereas the voltage \( v_{s,dc} \) applied to the secondary winding predefines the winding flux \( \Phi_{s} \). Again, the mismatch of the impressed winding fluxes causes the leakage fluxes \( \Phi_{s,a}, \Phi_{s,b}, \Phi_{s,c} \) which can be summed up to a total leakage flux. Also for this transformer structure, neglecting the magnetizing fluxes, all of the windings show the same leakage inductance current referred to a specific winding. The four-winding transformer exhibits a total leakage inductance \( L_{\sigma} \) referred to the primary side.

It is concluded, that the two transformer structures given in Fig. 3 exhibit a comparable electrical behaviour seen from the windings for negligible magnetizing fluxes. An equivalent electrical circuit is given by Fig. 4a where the total leakage inductance \( L_{\sigma} = L_{\sigma,a} + L_{\sigma,b} + L_{\sigma,c} \) referred to the primary side is drawn.

III. OPERATING PRINCIPLE

Like the three-port DC-DC converters discussed in [11]–[13], the proposed multi-port converter is operated by phase-shift control where the HF square-wave voltages applied to the transformer windings are phase-shifted against each other to control the power flow at the ports. The total leakage inductance \( L_{\sigma} \) of the transformer structure acts as decoupling and energy transfer element between the square-wave voltages as can be seen from Fig. 4a.

A. High-Frequency Square-Wave Voltage Summation

The transformer structures depicted in Fig. 3 sum up the HF square-wave voltages \( v_{p,a}, v_{p,b}, v_{p,c} \) which are generated by the switching networks \( S_{a}, S_{b}, S_{c} \) and phase-shifted towards the square-wave voltage \( v_{s,dc} \) by the angles \( \phi_{a}, \phi_{b}, \phi_{c} \) as can be seen in Fig. 5. There, also the resulting voltage sum \( v_{p,s} \) and the transformer leakage inductance current \( i_{L_{\sigma}} \) referred to the primary side over two switching periods \( T_{s} = 1/f_{s} \) are drawn.

The adaptation of the phase angles \( \phi_{a}, \phi_{b}, \phi_{c} \) within the period of the AC system enables keeping the half-cycle voltage-second product of the voltage sum \( v_{p,s} \) (shaded area in Fig. 5) at a nearly constant value. Therefore, a nearly constant ratio between the half-cycle voltage-second product of the voltage sum \( v_{p,s} \) and the voltage \( v_{s,dc} \) generated by switching network \( S_{dc} \) is achieved. In other words, this means that the amplitude ratio of the HF fundamentals of \( v_{p,s} \) and \( v_{s,dc} \) which are shown in Fig. 5 is time-independent.

Summarized, the above mentioned nearly constant control variable of the sum of the half-cycle voltage-second products...
enables to compose a voltage sum $v_{p,s} = v_{p,a} + v_{p,b} + v_{p,c}$ with a nearly constant half-cycle voltage-second product and a constant amplitude of the HF fundamental, respectively.

**B. High-Frequency Fundamental Model**

The multi-port converter depicted in Fig. 2 can be modelled by an equivalent circuit according to Fig. 4a by means of HF square-wave voltage sources for each switching network and a primary referred total leakage inductance $L_{\sigma}$ of the applied transformer structure. The leakage inductance current $i_{L,\sigma}$ flows through all of the equivalent sources and hence defines, together with the port voltage, the instantaneous powers $p_a, p_b, p_c$ which has to be delivered to or drawn from the corresponding port.

In the following, for the sake of simplicity, the analytical considerations are based on a HF fundamental model of the converter where higher order harmonics of the HF square-wave voltages are neglected. The HF voltages considering the clamping intervals $2\delta_{abc}, 2\delta_{dc}$ (see Fig. 5) and the resulting leakage inductance current are then described as complex HF phasors which are given by

$$v_{p,a} = 2v_a \cos (\delta_{abc}) e^{j\phi_a},$$

$$v_{p,b} = 2v_b \cos (\delta_{abc}) e^{j\phi_b},$$

$$v_{p,c} = 2v_c \cos (\delta_{abc}) e^{j\phi_c},$$

$$v'_{s,dc} = \frac{4n_{abc}V_{dc} \cos (\delta_{dc})}{\pi} e^{j\phi_{dc}},$$

$$i_{L,\sigma} = \frac{v_{p,a} + v_{p,b} + v_{p,c} - v'_{s,dc}}{j\omega_{abc} L_{\sigma}}.$$  \hspace{1cm} (5)

The HF phasors rotate with the angular frequency $\omega_a = 2\pi f_a$. Additionally, $v_{p,a}, v_{p,b}, v_{p,c}$ show a time-dependent amplitude caused by the AC voltage waveforms and a time-dependent phase caused by the time-varying phase-shift control. The square-wave voltage $v_{s,dc}$ is chosen to be the reference for the phase-shift control as shown in Fig. 5 and its phasor is placed on the real axis of the complex coordinate system, hence $\phi_{dc} = 0$ (see Fig. 4b).

**C. Modulation and Control Strategy**

For the proposed multi-port converter, a time-varying phase-shift control is applied, where the phase angles $\delta_a, \delta_b, \delta_c$, according to (1)-(3) represent the control functions whose waveforms have to assure sinusoidal phase currents $i_a, i_b, i_c$ with a given amplitude $I_{abc}$ and a given phase-shift $\phi_{abc}$ towards the phase voltages

$$v_a = \hat{V}_{abc} \cos (\omega_{abc}t),$$

$$v_b = \hat{V}_{abc} \cos \left(\omega_{abc}t - \frac{2\pi}{3}\right),$$

$$v_c = \hat{V}_{abc} \cos \left(\omega_{abc}t - \frac{4\pi}{3}\right).$$  \hspace{1cm} (8)

To get the required phase currents, the control functions are chosen in such a way that the corresponding instantaneous powers $p_a, p_b, p_c$ are drawn from or delivered to the ports which in turn are modelled by the equivalent HF voltage sources. The average active powers flowing out or into these sources over one switching cycle $T_s$ can be determined with the HF model according to

$$p_a = \text{Re} \left\{ \frac{1}{2} v_{p,a} L_{L,\sigma} \right\} = v_a \frac{i_{L,\sigma} \cos (\delta_{abc})}{\pi} \cos (\phi_a - \phi_1),$$

$$p_b = \text{Re} \left\{ \frac{1}{2} v_{p,b} L_{L,\sigma} \right\} = v_b \frac{i_{L,\sigma} \cos (\delta_{abc})}{\pi} \cos (\phi_b - \phi_1),$$

$$p_c = \text{Re} \left\{ \frac{1}{2} v_{p,c} L_{L,\sigma} \right\} = v_c \frac{i_{L,\sigma} \cos (\delta_{abc})}{\pi} \cos (\phi_c - \phi_1).$$

Taking the reactive power consumed by the AC capacitors $C_{ac}$ into account, the reference value of the control current can be represented by a phasor

$$I_{ctrl} = I_{abc} - I_{C}.$$  \hspace{1cm} (12)
The load is represented by a controlled current source \(I_{ctrl}\). The phasor of the phase reference current is

\[
\hat{I}_{abc} = \hat{i}_{abc} e^{j\phi_{abc}}.
\]

\(\hat{i}_{abc}\) denotes the reference amplitude whereas \(\phi_{abc}\) is the reference phase of the AC port currents \(i_a, i_b, i_c\).

In the time domain, the reference values of the control currents per phase are then given by

\[
i_a^s = \hat{I}_{ctrl} \cos(\omega_{abc} t + \phi_{ctrl}), \tag{15}
i_b^s = \hat{I}_{ctrl} \cos\left(\omega_{abc} t - \frac{2\pi}{3} + \phi_{ctrl}\right), \tag{16}
i_c^s = \hat{I}_{ctrl} \cos\left(\omega_{abc} t - \frac{4\pi}{3} + \phi_{ctrl}\right), \tag{17}
\]

with amplitude \(\hat{I}_{ctrl} = |I_{ctrl}|\) and phase \(\phi_{ctrl} = \angle I_{ctrl}\).

The control functions \(\phi_a, \phi_b, \phi_c\) can be determined using the nonlinear system of equations

\[
i_a^s = i_a = \hat{i}_{L\sigma} \cos\left(\frac{\delta_{abc}}{\pi}\right) \cos(\phi_a - \phi_i), \tag{18}
i_b^s = i_b = \hat{i}_{L\sigma} \cos\left(\frac{\delta_{abc}}{\pi}\right) \cos(\phi_b - \phi_i), \tag{19}
i_c^s = i_c = \hat{i}_{L\sigma} \cos\left(\frac{\delta_{abc}}{\pi}\right) \cos(\phi_c - \phi_i), \tag{20}
\]

while considering \(\hat{i}_{L\sigma} = |i_{L\sigma}|\) and \(\phi_i = \angle i_{L\sigma}\).

By comparing (15)-(17) with (18)-(20) and knowing that \(\delta_{abc}, \delta_{dc} \in [0, \pi/2]\), the most obvious control functions which lead to sinusoidal AC port currents show the mathematical form

\[
\phi_a = \omega_{abc} t + \phi_s, \tag{21}
\phi_b = \omega_{abc} t - \frac{2\pi}{3} + \phi_s, \tag{22}
\phi_c = \omega_{abc} t - \frac{4\pi}{3} + \phi_s. \tag{23}
\]

The amplitude and the phase of the reference values of the control currents according to (15)-(17) are then given by

\[
\hat{I}_{ctrl} = \frac{\hat{i}_{L\sigma} \cos(\delta_{abc})}{\pi}, \tag{24}
\]

\(\phi_{ctrl} = \phi_s - \phi_i, \tag{25}\)

where \(\delta_{abc}, \phi_s\) represent the control variables.

Inserting the control functions \(\phi_a, \phi_b, \phi_c\) according to (21)-(23) into (1)-(3) and adding the voltage phasors, the composed voltage sum phasor

\[
\hat{v}_{p,s} = \hat{v}_{p,a} + \hat{v}_{p,b} + \hat{v}_{p,c} = \hat{v}_{p,s} e^{j\phi_s} = \frac{3\hat{V}_{abc} \cos(\delta_{abc})}{\pi} e^{j\phi_s} \tag{26}
\]

results, which shows a time-independent amplitude \(\hat{v}_{p,s}\) and a time-independent phase angle \(\phi_s\). Applying the rotation operators \(e^{j\phi_a}, e^{j\phi_b}, e^{j\phi_c}\) on the sinusoidal AC voltages \(v_a, v_b, v_c\), a fixed space vector is formed, which is then rotated over a switching cycle \(T_s\). Due to the time-independent phasors \(\hat{v}_{p,s}, \hat{L}_{s,dc}\), also the transformer leakage inductance current \(\hat{I}_{L\sigma}\) shows a constant amplitude and phase over time. Fig. 4b depicts a phasor diagram of the HF phasors \(\hat{v}_{p,s}, \hat{L}_{s,dc}, \hat{L}_{L\sigma}, \hat{L}_{L\sigma}\) in case of capacitive AC port currents.

D. Accuracy of High-Frequency Fundamental Model

In general, a periodic square-wave signal \(s(t)\) with clamping interval \(2\delta_s\), amplitude \(h\) and angular frequency \(\omega_s\) like the voltages drawn in Fig. 5 can be represented by its fourier series

\[
s(t) = \frac{4h}{\pi} \sum_{k=1}^{\infty} \cos\left((2k-1)\delta_s\right) \sin\left((2k-1)\omega_st\right). \tag{27}
\]

When adding the HF square-wave voltages \(v_{p,a}, v_{p,b}, v_{p,c}\) while considering their phase-shifts from (21)-(23) and their amplitudes according to the half of (6)-(8), it can be seen from the fourier series (27), that harmonics with an odd multiple of three of the HF fundamental frequency \(\omega_s\) cancel out and are not present in the voltage sum \(v_{p,s}\). Nevertheless, if \(v_{s,de}\) exhibits these orders of HF harmonics, corresponding current harmonics in the transformer leakage current will be driven. This causes power shares on the AC ports which lead to LF current harmonics exhibiting odd multiples of three of the fundamental AC system frequency \(\omega_{abc}\). To suppress these LF AC current harmonics, \(\delta_{dc} = \pi/6\) can be chosen which in turn cancels out the corresponding HF harmonics in the square-wave voltage \(v_{s,de}\). Besides the fundamental, HF current harmonics of order 5, 7, 11, 13, 17, ... remain in the transformer leakage current and contribute to a small AC current distortion. The control functions above are derived by setting

\[
p_{abc} = p_{abc,(1)} \tag{28}
\]

where \(p_{abc}\) denotes the reference value of the instantaneous power of the considered phase and \(p_{abc,(1)}\) the HF fundamental power. In a next step, also higher order HF power shares are considered by solving

\[
p_{abc} = \sum_{k=1}^{\infty} p_{abc,(k)} \tag{29}
\]

in order to get improved control functions \(\phi_a', \phi_b', \phi_c'\) and a lower input current THD.
IV. Prototype System

For validating the theoretical analysis, a prototype system of the multi-port converter drawn in Fig. 2 is simulated in GeckoCIRCUITS [14] where the AC system voltage and frequency are fixed at 230 V<sub>rms</sub> and 50 Hz. The DC port is considered to be at a constant voltage of 400 V<sub>dc</sub>. The simulation model corresponds to Fig. 2 and applies three two-winding transformers, which show negligible high magnetizing inductances and primary referred leakage inductances of 3 µH each. The converter is operated at a constant frequency of 20 kHz. Table I summarizes the simulation parameters.

A. Solutions of Control Variables

In order to obtain the desired AC port phase currents with corresponding amplitude I<sub>abc</sub> and phase φ<sub>abc</sub>, the control variables δ<sub>abc</sub>, φ<sub>a</sub> have to be determined. This can be done by solving the nonlinear system of equations (18)-(20) while taking the control functions (21)-(23) into account. Solutions of the control variables for the prototype system with parameters given in Table I are depicted in Fig. 7 for phase angles φ<sub>abc</sub> = {−π/2, −π/4, 0, π/4, π/2} versus phase current amplitude I<sub>abc</sub>.

![Fig. 7. Solutions of the control variables δ<sub>abc</sub>, φ<sub>a</sub> for the prototype system with parameters from Table I and δ<sub>dc</sub> = π/6 for phase angles φ<sub>abc</sub> = {−π/2, −π/4, 0, π/4, π/2} versus phase current amplitude I<sub>abc</sub>.](image)

![Fig. 8. Phasor diagrams for the HF fundamental model for phase control angles φ<sub>ctrl</sub> = −π/2 (a), φ<sub>ctrl</sub> = −π/4 (b), φ<sub>ctrl</sub> = 0 (c), φ<sub>ctrl</sub> = π/8 (d), φ<sub>ctrl</sub> = π/4 (e) and φ<sub>ctrl</sub> = π/2 (f).](image)

<table>
<thead>
<tr>
<th>TABLE I</th>
</tr>
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<tbody>
<tr>
<td>SIMULATION PARAMETERS OF THE PROTOTYPE SYSTEM.</td>
</tr>
<tr>
<td>AC system voltage</td>
</tr>
<tr>
<td>AC system frequency</td>
</tr>
<tr>
<td>DC port voltage</td>
</tr>
<tr>
<td>Switching frequency</td>
</tr>
<tr>
<td>Transformer turns ratios</td>
</tr>
<tr>
<td>Transformer leakage inductances</td>
</tr>
<tr>
<td>Transformer magnetizing inductances</td>
</tr>
<tr>
<td>Total transformer leakage inductance</td>
</tr>
<tr>
<td>Inductors</td>
</tr>
<tr>
<td>Capacitors</td>
</tr>
<tr>
<td>Inductor</td>
</tr>
<tr>
<td>Capacitor</td>
</tr>
</tbody>
</table>

B. Simulation Results

The multi-port converter is simulated in steady-state with δ<sub>dc</sub> = π/6 in AC-to-DC operation with reference values I<sub>abc</sub> = 20 A and φ<sub>abc</sub> = π/8 for the AC phase currents. The control variables are then given by δ<sub>abc</sub> = 0.7543 and φ<sub>a</sub> = 0.4207. In the simulation model according to Fig. 2, the star point of the converter N<sub>abc</sub> and the star point of the grid are connected to earth through 1 Ω earth resistances. The DC side is also grounded via a 1 Ω earth resistance. The simulation
results without feedback control applying the derived control functions are shown in Fig. 9 and Fig. 10.

V. CONCLUSION

An innovative bidirectional isolated multi-port converter with multi-phase AC ports and DC ports is presented, which is ideally suited for use in a wind energy generation system. The concept of multi-phase AC ports is introduced, which allows the direct coupling of a multi-phase AC system, whose absolute phase voltages add up to a nearly constant sum over time, with one or more DC ports over a single high-frequency transformer structure. The proposed converter is operated by a time-varying phase-shift control to draw or inject sinusoidal currents with a corresponding amplitude and phase at the AC ports. The analytical description and simulation results of a prototype system for validating the converter topology are provided.

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