A Novel Three-Phase Three-Port UPS Employing a Single High-Frequency Isolation Transformer

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Abstract — A three-phase PWM rectifier and a three-phase PWM inverter are coupled via two four-quadrant full-bridge converter cells and a high-frequency isolation transformer. By employing a third transformer winding and further full-bridge cell battery energy storage is incorporated into the power transfer between rectifier and inverter resulting in a Three-Port UPS concept. The phase shift control of the power flow between the ports is analyzed for square-wave operation of the full-bridge cells. Furthermore, the utilization of the degrees of freedom of the system control, i.e. the extension to duty cycle control for optimizing the system behavior is discussed and control laws ensuring minimum overall system losses are derived. Finally, a control-oriented converter model is proposed and the decoupled control of the power flow of the ports is treated briefly. All theoretical considerations are verified by simulations using PSIM.

I. INTRODUCTION

Highly reliable three-phase on-line uninterrupted power supply systems (UPS) are formed by back-to-back connection, i.e. DC side coupling of a mains-side voltage source rectifier and a load-side voltage source inverter [1]. There, the battery energy storage is directly connected to the DC link or coupled via a DC/DC converter in order to allow a control of the charging or discharging current and a compensation of battery voltage changes. Furthermore, for safety reasons and voltage level adaptation in general a 50/60Hz isolation transformer is employed on the input or output side which, however, constitutes a significant drawback concerning volume/weight, costs and efficiency.

In this paper a novel UPS concept with high frequency isolation of a three-phase PWM rectifier stage (system/port 1), a three-phase PWM inverter stage (system/port 2) and a battery energy storage (system/port 3) is proposed, [2] (cf. Fig.1). There, port 1, port 2 and port 3 are coupled via three four-quadrant full-bridge converter cells operating in phase-shift mode and a single three-winding transformer. Accordingly, the system will be denoted as Three-Port UPS in the following.

For realizing the three-phase rectifier and inverter function of the system 20 (6+4+4+6) power transistors with anti-parallel diodes have to be employed. Therefore, the realization effort is lower than, e.g. for a three-phase high-frequency link AC/AC matrix converter comprising 24 transistors and diodes [4]. As further advantage one has to point out the continuous shape of the input and output currents and the decoupling of the mains and the load concerning power and voltage fluctuations in case sufficient DC link capacitance $C_1$ and $C_2$ is provided for the rectifier and/or the inverter. Alternatively, $C_1$ and $C_2$ could be realized as foil capacitors ensuring only a constant DC link voltage level for switching transients.

The system shown in Fig.1 in general allows a bidirectional energy conversion between two three-phase voltage systems and a DC voltage source and therefore is of potential interest also in connection with future distributed generator of electric energy. There, the mains could, e.g. be formed by a wind turbine or a variable frequency generator powered by a mini micro turbine where the battery storage could compensate power fluctuations of the regenerative energy sources. Also, further other storage elements like flywheels or SMES or a fuel cell or photovoltaic generator could be incorporated in parallel by individual full-bridge converter cells and related transformer windings. An interesting application area of such X-port UPS-type converter system would e.g. be future intelligent nodes of the energy distribution system which also could be extended to a power quality control centre [5] featuring symmetrization of unbalanced loads and active filtering of mains voltage or load current harmonics.

The control of the power flow in the three-port system is in the simplest case by proper phase shift of the individual full-bridge cells operating in square-wave mode. In Section II the system control characteristics are determined based on the equivalent circuit of a three-winding transformer [6] and the dependency of the power transfer between two full-bridge cells on the phase displacement of the control signals. There, in a first step a turns ratio of $N_1:N_2:N_3=1:1:1$ is assumed and the PWM rectifier, the PWM inverter and the voltage-type energy storage are replaced by equal

![Fig1: Proposed three-phase three-port high-frequency link converter (Three-Port UPS, [2]). The high frequency coupling of multiple four-quadrant full-bridge converter cells via a single transformer has been shown in principle in [3] (cf. Fig.8 in [3]), however, no procedure for controlling the system has been given there.](image-url)
The three-port converter can be simplified to a two-port converter if the individual full-bridge converter cell control signals are represented by

\[ P_{12} = \frac{\Delta U_1}{2} \left( 1 - \cos(\phi_2) \right) \]

where \( V_{12} \) is the DC voltage of the converters generating \( u_1 \) and \( u_2 \) and \( T \) denotes the switching cycle. According to (2) the amount and direction of the power flow is determined by the phase shift \( \phi_2 \).

This is also immediately clear from a phasor diagram of the fundamentals of the voltages and currents which is depicted in Fig.3. The fundamental power transferred from port 1 and/or \( u_1 \) to port 2 and/or \( u_2 \) is

\[ P_{12} = \frac{1}{2} \left( U_{12} \right)^2 \frac{\Delta \phi_1}{\sin(\Delta \phi_1)} \cos(\alpha) \]

where \( U_{12} = 4\pi V_1 \), \( U_{23} = 4\pi V_2 \), and \( U_{31} = 4\pi V_3 \) is the absolute value of \( u_1 = y - y_2 \) and \( \phi_1 \) denotes the phase displacement of \( y_1 \) and \( u_2 \) which is oriented perpendicular to \( y_{12} \). For \( \phi_1 > 0 \) (cf. Fig.3(a)), and/or in case \( u_1 \) is leading \( u_2 \), \( \Delta \phi_1 \) is smaller than \( \pi/2 \). Therefore, we have \( P_{12} > 0 \), the power flow is physically oriented from port 1 and/or \( u_1 \) to port 2 and/or \( u_2 \) and reverses for \( \phi_1 < 0 \) since \( \alpha \) then is larger than \( \pi/2 \) (cf. Fig.3(b)).

The Y-equivalent circuit depicted in Fig.2(b) can be transformed into a \( \Delta \)-equivalent circuit [7] shown in Fig.2(c) which allows to determine the resulting power flow of the three-port system by superposition of the power transfer of three two-port systems \( u_1, u_2, u_3 \), and \( u_2, u_3, u_1 \), and \( u_3, u_1, u_2 \). E.g., for \( \phi_1 > 0 \), \( \phi_2 > 0 \) and \( \phi_3 > \phi_2 \) (cf. Fig.4(a)) the power flow is from \( u_1 \) to \( u_2 \), \( u_2 \) to \( u_3 \) and \( u_3 \) to \( u_1 \), \( \phi > \phi_2 \). Therefore, \( u_1 \) is acting as a source and \( u_2 \) is consuming power. Dependent on the relation of \( \phi_2 \) and \( \phi_3 \) of \( V_1, V_2 \) and \( V_3 \), Port 3 and/or \( u_3 \) can be sinking or sourcing power or remain at zero power.

![Fig.2](image-url)  
**Fig.2:** (a) Equivalent circuit of a three-winding transformer; (b) Y-equivalent circuit of the proposed converter used for analyzing the dependency of the power transfer between the ports on the control signal phase displacements; (c) \( \Delta \)-equivalent circuit of the three-port system; (d) equivalent circuit for studying the power flow between two ports when the third port is open.

![Fig.3](image-url)  
**Fig.3:** The phasor diagram of the fundamentals of the voltages and currents when the third port is open.

![Fig.4](image-url)  
**Fig.4:** Equivalent circuit for studying the power flow between three ports. The direction of power flow is only determined by \( \phi_2 \) and \( \phi_3 \), not by \( V_1, V_2 \) and \( V_3 \).
In summary, a power transfer is possible in any direction between any ports simultaneously and the direction is only determined by $\phi_3$ (cf. Fig.4).

C. Control Leaving One Port at Zero Average Power

An important function of an UPS system is to directly supply power from the mains to the load without charging or discharging the battery energy storage. For the three-port UPS this cannot be realized at low battery voltage even if all power transistors of the battery port full-bridge cell are remaining in the off-state as the anti-parallel free-wheeling diodes would be forced into conduction by high output voltages of the mains or load port full-bridge cells. Therefore, $\phi_1$ and $\phi_3$ have to be selected properly in order to achieve $P_i = 0$ and a given value of $P_i = -P_i = P$ (since the sum of the power of three voltage sources has to be zero, $P_1 = P_2 = P_3 = 0$, if losses are neglected).

As the currents in the leakage inductances are directly the currents of the voltage sources of the Y-equivalent circuit, the following considerations are referring to Fig.2(b). Here, again $\phi_1 > 0$, $\phi_2 > 0$ and $\phi_3 > \phi_1$ is assumed. The key waveforms resulting in $P_i = 0$ are shown in Fig.5 for $V_i = V_f = V_j$ and for the general case of different DC voltage levels $V_i$, $V_j$, $V_j$ in Fig.6. According to the law of superposition, $u_0$ contained $u_1$, $u_2$, $u_3$ is

$$u_0 = u_1 L_1 + u_2 L_2 + u_3 L_3$$

(4)

For the power flow of the ports we then have

$$P_1 = \frac{\phi_2 (\pi - \phi_2) V_i^2 L_3 + \phi_1 (\pi - \phi_1) V_i^2 L_2}{2\pi^2 (L_1 L_2 + L_1 L_3 + L_2 L_3)}$$

$$P_2 = \frac{\phi_2 (\phi_2 - \pi) V_i^2 L_3 + (\phi_2 - \phi_1) (\pi - \phi_1) V_i^2 L_2}{2\pi^2 (L_1 L_2 + L_1 L_3 + L_2 L_3)}$$

$$P_3 = \frac{\phi_2 (\phi_2 - \pi) V_i^2 L_3 + (\phi_2 - \phi_1) (\pi - \phi_1) V_i^2 L_2}{2\pi^2 (L_1 L_2 + L_1 L_3 + L_2 L_3)}$$

(5)

From (5) now the relation of $P$ and $\phi_1$ and of $\phi_2$ and $\phi_3$ can be derived under the side condition of $P_1 = 0$ and $P_2 = P = P$. As shown in Fig.7 and Fig.8 for $V_i = 500V$, $V_f = 400V$, $V_j = 360V$, $P_1 = P_2 = P = 5kW$, $P_1 = 0$, $L_1 = L_2 = L_3 = 100\mu H$, and $f = 20kHz$, there are two solutions for $\phi_1$ and $\phi_3$. As a higher phase shift $\phi_1$ and $\phi_3$ result in a higher peak value of the current in the corresponding leakage inductance and/or in higher conduction and switching losses the lower phase shift values $\phi_1$ and $\phi_3$ which have to be selected for the system control.

**Fig.5**: Key waveforms for achieving zero net power flow to port 3 assuming equal levels of the full-bridge converter cell DC voltages, i.e. $V_i = V_f = V_j$.

**Fig.6**: Key waveforms for $P_i = 0$ and/or $P_1 = P_2 = P_3$ for differing levels of the full-bridge converter cell DC voltages $V_i$, $V_j$, $V_j$.

**Fig.7**: Phase shift $\phi_1$ for achieving $P_i = 0$, $P_1 = 5kW$ and $P_1 = 0$. Operating parameters: $V_i = 500V$, $V_f = 400V$, $V_j = 360V$, $L_1 = L_2 = L_3 = 100\mu H$, and $f = 20kHz$.

**Fig.8**: Dependency of phase shift $\phi_1$ on $\phi_3$ for achieving zero net power flow to port 3, i.e. $P_i = 0$ (cf. Fig.7). Operating parameters: as for Fig.7.
III. MINIMIZATION OF THE OVERALL SYSTEM LOSSES

A. Introduction of Duty Cycle Control

In order to gain degrees of freedom for minimizing the overall system losses, duty cycle variation of the full-bridge converter cell output voltages \( u_1, u_2 \) and \( u_3 \) could be introduced in addition to phase shift control as shown in Fig.9. There, the control range of \( \delta_1, \delta_2 \) and \( \delta_3 \) is from 0 to \( \pi/2 \).

B. Zero Circulating Power

The power transferred from \( u_1 \) to \( u_2 \), from \( u_2 \) to \( u_3 \) and from \( u_3 \) to \( u_1 \) will be denoted as \( P_{12}, P_{23} \) and \( P_{31} \) in the following. Aiming for minimizing losses a circulation of power inside the system which would not contribute to the power flow of the ports has to be prevented, i.e.

\[
P_{12} + P_{23} + P_{31} = 0
\]

has to be ensured. Considering \( P_2 = P_{21} + P_{23} \) and \( P_3 = P_{32} + P_{31} \), we have for \( P_{12}, P_{23}, \) and \( P_{31} \) in dependency on \( P_1, P_3 \) with (6)

\[
P_{12} = -\frac{2}{3} P_1 - \frac{1}{3} P_3
\]

\[
P_{23} = \frac{1}{3} P_2 + \frac{1}{3} P_3
\]

\[
P_{31} = \frac{1}{3} P_2 + \frac{2}{3} P_3
\]

For the sake of simplicity the further considerations are restricted to fundamentals of the voltages and currents. Furthermore, \( L_{12} = L_{23} = L_{31}, \phi_2 = 0, \phi_3 = 0 \) and \( \phi_1 = \phi_3 \) is assumed. The corresponding phasor diagram is shown in Fig.10 which \( u_2 \) defines the orientation of the real axis

\[
u_2 = U_2 + 0 \jmath
\]

We then have for \( P_{12} \)

\[
P_{12} = \frac{1}{2} U_1 U_2 \cos(\alpha) = \frac{U_1}{2a_{d_{12}}} U_{12} \cos(\alpha)
\]

Considering the phase displacement \( \phi_2 \) of \( u_1 \) and \( u_2 \) (cf. Fig.9) and/or of \( u_2 \) and \( u_3 \), \( u_2 \) can be expressed as (cf. Fig.10)

\[
u_2 = U_2 \cos(\phi_2) - U_3 \sin(\phi_2) \jmath
\]

Accordingly, \( u_{12} \) formed by \( u_1 \) and \( u_2 \) is

\[
u_{12} = u_1 - u_2 = U_1 - U_2 \cos(\phi_2) + U_3 \sin(\phi_2) \jmath
\]

Introducing the phase displacement \( \alpha \) of \( u_2 \) and \( i_{12}, i_{23} \) which is leading \( i_{12} \) by \( \pi/2 \) can be alternatively formulated as

\[
u_{12} = U_{12} \cos(\alpha) + U_{12} \sin(\frac{\pi}{2} - \alpha) \jmath
\]

\[
u_{12} = U_{12} \sin(\alpha) + U_{12} \cos(\alpha) \jmath
\]

Combining (11) and (12) results in

\[
\hat{U}_{12} \sin(\phi_2) = U_{12} \sin(\alpha) + U_{12} \cos(\alpha) \jmath
\]

\[
\hat{U}_{12} \cos(\alpha) = U_{12} \sin(\phi_1)
\]

Considering (13), the power flow \( P_{12} \) (cf. (9)) now can be represented as

\[
P_{12} = \frac{1}{2} \hat{U}_{12} \hat{U}_{23} \sin(\phi_2) = \frac{1}{2} U_{12} U_{12} \sin(\phi_1)
\]

Taking finally into account that the area \( S_{\triangle OAB} \) of the triangle \( \triangle OAB \) in Fig.10 is

\[
S_{\triangle OAB} = \frac{1}{2} U_1 U_2 \sin(\phi_1)
\]

the power transferred between two ports in general is proportional to the area of the triangle defined by the phasors of the voltages of the ports and is inversely proportional to the equivalent impedance and/or inductance connecting the ports (cf. Fig.2c).

\[
P_{12} = \frac{1}{2} \hat{U}_{12} \hat{U}_{23} \sin(\phi_2) = \frac{1}{2} U_{12} U_{12} \sin(\phi_1)
\]

Accordingly, \( P_{13} \) and \( P_{23} \) can be expressed as

\[
P_{13} = \frac{1}{2} \frac{1}{a_{d_{13}}} S_{\triangle OAC} \quad P_{23} = \frac{1}{2} \frac{1}{a_{d_{23}}} S_{\triangle OBC}
\]

where \( S_{\triangle OAC} \) denotes the area of the triangle \( \triangle OAC \) and \( S_{\triangle OBC} \) denotes the area of the triangle \( \triangle OBC \). Considering \( P_{11} + P_{23} + P_{31} = 0 \), and \( L_{12} = L_{13} = L_{23} \) as assumed above, (17) yields in combination with (16)

\[
P_{12} + P_{31} = R_2 - R_3 - R_{13}
\]

\[
= \frac{1}{2} \frac{1}{a_{d_{13}}} S_{\triangle OAC} - \frac{1}{2} \frac{1}{a_{d_{23}}} S_{\triangle OBC} - \frac{1}{2} \frac{1}{a_{d_{13}}} S_{\triangle OAC}
\]

\[
= 0
\]

Simplifying (18), results in

\[
S_{\triangle OAC} = S_{\triangle OBC} + S_{\triangle OAC}
\]

Accordingly, for preventing circulating power flow, the head of \( u_1 \) shown as point \( C \) in Fig.10 has to be located on the line \( AB \) and/or \( u_2 \) and \( u_3 \) have to be aligned what results in an alignment of \( i_{12}, i_{23}, \) and \( i_{2} = i_{12} + i_{23} \), what considerably simplifies the closed-form system.
The corresponding range of $\phi_3$ is denoted as $\text{Area}_3$ in the following (cf. Fig. 11). In analogy, $\text{Area}_{\overline{A}}$ and $\text{Area}_{\overline{B}}$ are resulting from $\phi_1$ and $\phi_2$. Considering all restrictions, the system operation finally has to be restricted to the $\text{Area}_{\overline{A}}$ defined by the intersection of $\text{Area}_{\overline{A}}$, $\text{Area}_{\overline{B}}$, and $\text{Area}_{3}$.

D. Minimization of Overall System Losses

For realizing the full-bridge converter cells in IGBT technology for a rated power in the range of 5...10kW, the system switching and conduction losses could be derived based on [10]. There, we have for the turn-on and turn-off energy loss of a power transistor and the turn-off energy loss of a power diode

$$w_{\text{Soff}}(u,i) = K_{\text{Soff1}} \cdot u \cdot i + K_{\text{Soff2}} \cdot u^2 \cdot i + K_{\text{Soff3}} \cdot u^2 \cdot i^2 + K_{\text{Soff4}} \cdot u^2 \cdot i^2 + K_{\text{Soff5}} \cdot u^2 \cdot i^2$$

$$w_{\text{Scon}}(u,i) = K_{\text{Scon1}} \cdot u \cdot i + K_{\text{Scon2}} \cdot u^2 \cdot i + K_{\text{Scon3}} \cdot u^2 \cdot i^2 + K_{\text{Scon4}} \cdot u^2 \cdot i^2 + K_{\text{Scon5}} \cdot u^2 \cdot i^2$$

where the indices $S, D$ denote whether the transistor or the diode is considered, and $u$ and $i$ are the voltage and the current being switched. The coefficients $K_i$ are determined by measurements and specified in [10].

For the conduction losses of the valves we have again according to [10]

$$\overline{P}_{S/D} = U_{f/S/D} \cdot \overline{i}_f / D_{\text{rms}}$$

where $f$ denotes the average value and $i_{\text{rms}}$ is the rms value of the transistor or the diode current.

The overall system losses resulting for, e.g. $P_f = 4kW$, $P_f = -3kW$, $P_f = -1kW$, $V_f = 500V$, $V_f = 400V$, $V_f = 360V$, $L_{\text{i}} = L_{\overline{f}} = 100\mu H$, and $f_r = 20kHz$ are shown in Fig. 12. There, operating point $A$ results in minimum overall power semiconductor losses. This is clearly verified by operating point $B$ which is characterized by significantly higher current amplitudes and higher voltage and current phase displacements and/or significantly higher losses (cf. Fig. 13).

Fig. 13 also justifies the approximation of the actual current waveforms with the fundamentals (cf. (b) and (d)). Accordingly, the calculation of $\phi_1$ and $\phi_2$ based on (22) allows an accurate pre-control the system power flow (cf. Fig. 14).

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**Fig. 11:** Operating $\text{Area}_{\overline{A}}$, i.e. admissible range of phase shift $\phi_1$ and phase shift $\phi_2$. Resulting form (22). Assumed operating parameters: $P_f = 4kW$, $P_f = -3kW$, $P_f = -1kW$, $V_f = 500V$, $V_f = 400V$, $V_f = 360V$, $L_{\text{i}} = L_{\overline{f}} = 100\mu H$, and $f_r = 20kHz$.  

**Fig. 12:** Overall system losses in dependency of $\phi_1$ and $\phi_2$. Assumed operating parameters: $P_f = 4kW$, $P_f = -3kW$, $P_f = -1kW$, $V_f = 500V$, $V_f = 400V$, $V_f = 360V$, $L_{\text{i}} = L_{\overline{f}} = 100\mu H$, and $f_r = 20kHz$.  

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C. Range of $\phi_1$ and $\phi_2$

Substituting $\phi = 2\pi n/T$, $U_1 = 4\pi V \cos(\delta_1)$, and $U_2 = 4\pi V \cos(\delta_2)$ in (14), $P_{12}$ can be represented as

$$P_{12} = \frac{1}{2 \alpha_{12}} U_1 \sin(\phi_1)$$

$$= \frac{4\pi}{\alpha_{12}} V_1 \cos(\delta_1) V_2 \cos(\delta_2) \sin(\phi_1)$$

(20)

Analogously, we have for $P_{13}$ and $P_{23}$

$$P_{13} = \frac{4\pi}{\alpha_{13}} V_1 \cos(\delta_1) V_3 \cos(\delta_3) \sin(\phi_1)$$

$$P_{23} = \frac{4\pi}{\alpha_{23}} V_2 \cos(\delta_2) V_3 \cos(\delta_3) \sin(12 - \phi_2)$$

(21)

Referring to (20) and (21) the duty cycles $\delta_1$, $\delta_2$, and $\delta_3$ can be expressed as

$$\delta_1 = \frac{1}{\alpha_{12}} \pi \left[ \frac{2\pi}{\alpha_{12}} P_{12} L_{\text{i}} L_{\text{t}} \sin(\phi_1) \right]$$

$$\delta_2 = \frac{1}{\alpha_{13}} \pi \left[ \frac{2\pi}{\alpha_{13}} P_{13} L_{\text{i}} L_{\text{t}} \sin(\phi_1) \right]$$

$$\delta_3 = \frac{1}{\alpha_{23}} \pi \left[ \frac{2\pi}{\alpha_{23}} P_{23} L_{\text{i}} L_{\text{t}} \sin(\phi_1) \right]$$

(22)

For controlling the system 5 degrees of freedom, i.e. the phase displacements $\phi_1$, $\phi_2$, and the duty cycles $\delta_1$, $\delta_2$, and $\delta_3$ are available. Defining the power flow of two ports, e.g. $P_f$ and $P_{\overline{f}}$ and ensuring zero circulating power two degrees of freedom are remaining, i.e. the converter characteristics can be expressed in dependency of $\phi_1$ and $\phi_2$. Taking into account the restriction of the argument of an inverse cosine function to $[-1, 1]$, (22) results in a limitation of the operating range of the converter and/or in a limitation of the admissible range of $\phi_1$ and $\phi_2$. E.g., there follows considering $\phi_1$ and $\phi_2$,

$$-1 \leq \frac{\pi}{\alpha_{12}} P_{12} L_{\text{i}} L_{\text{t}} \sin(\phi_1) \leq 1$$

(23)
As mentioned in Section 3.C, two degrees of freedom, i.e. $\phi_1$ and $\phi_2$ are available for the system control. Accordingly, the system could be considered as a two input ($\phi_1$ and $\phi_2$) and two output ($P_2$ and $P_3$) control system.

The power $P_2$ of port 2 is the sum of $P_{21}$ (cf. (20)) and $P_{23}$ (cf. (21))

$$P_2 = P_{21} + P_{23} = P_{21} - P_{23}$$

where $K_1 = \frac{4T_s}{\pi l_{12}} V_1 \cos(\delta_1) W_2 \cos(\delta_2)$

$$K_2 = \frac{4T_s}{\pi l_{23}} V_2 \cos(\delta_2) W_1 \cos(\delta_1)$$

Assuming that system operation is close to point $A$ is ensured by proper pre-control (cf. Fig.12, where $\phi_3$ and $\phi_4$ is, and $\delta_{1A}, \delta_{2A}, \delta_{3A}, \delta_{4A}$ could be determined based on (22)) the controller only has to slightly adjust $\phi_1$ and $\phi_2$ in a given operating region. Therefore, for deriving a control-oriented system model, (26) can be linearized at the desired operating point $A$ resulting in

$$P_2 = P_{2A} + \Delta P_2$$

$$= K_{1A} \sin(\phi_{1A}) + K_{2A} \sin(\phi_{2A})$$

$$+ K_{1A} \cos(\phi_{2A}) \Delta \phi_2 - K_{2A} \cos(\phi_{1A} + \phi_{2A}) \Delta \phi_2$$

$$+ K_{2A} \cos(\phi_{1A} + \phi_{2A}) \Delta \phi_2$$

$$= P_{2A} + G_{1A} \Delta \phi_2 + G_{12A} \Delta \phi_3$$

where $G_{1A} = K_{1A} \cos(\phi_{1A}) - K_{2A} \cos(\phi_{1A} + \phi_{2A})$

$$G_{12A} = K_{2A} \cos(\phi_{1A} + \phi_{2A})$$

$$P_{2A} = K_{1A} \sin(\phi_{1A})$$

with

$$K_{1A} = \frac{4T_s}{\pi l_{12}} V_1 \cos(\delta_{1A}) W_2 \cos(\delta_{2A})$$

$$K_{2A} = \frac{4T_s}{\pi l_{23}} V_2 \cos(\delta_{2A}) W_1 \cos(\delta_{1A})$$

In analogy, linearizing $P_3$ at operating point $A$ yields

$$P_3 = P_{3A} + \Delta P_3$$

$$= P_{3A} + \Delta P_3$$

$$= K_{3A} \sin(\phi_{3A}) - K_{4A} \sin(\phi_{3A} - \phi_{4A})$$

$$+ K_{3A} \cos(\phi_{3A} - \phi_{4A}) \Delta \phi_4$$

$$+ K_{4A} \cos(\phi_{3A} - \phi_{4A}) \Delta \phi_3$$

$$= P_{3A} + G_{3A} \Delta \phi_3 + G_{32A} \Delta \phi_4$$

$$G_{3A} = K_{3A} \cos(\phi_{3A}) + K_{3A} \cos(\phi_{3A} - \phi_{4A})$$

$$P_{3A} = K_{3A} \sin(\phi_{3A}) - K_{4A} \sin(\phi_{3A} - \phi_{4A})$$

$$= P_{3A} + G_{3A} \Delta \phi_3 + G_{32A} \Delta \phi_4$$

with

$$K_{3A} = \frac{4T_s}{\pi l_{12}} V_1 \cos(\delta_{1A}) W_2 \cos(\delta_{2A})$$

$$K_{32A} = \frac{4T_s}{\pi l_{23}} V_2 \cos(\delta_{2A}) W_1 \cos(\delta_{1A})$$

The interaction of the control loops (29) can now be eliminated.

In summary, we have for the small signal model of the system

$$\begin{bmatrix} \Delta P_2 \end{bmatrix} = \begin{bmatrix} G_{11} & G_{12} & G_{13} & G_{14} \\ G_{21} & G_{22} & G_{23} & G_{24} \\ G_{31} & G_{32} & G_{33} & G_{34} \\ G_{41} & G_{42} & G_{43} & G_{44} \end{bmatrix} \begin{bmatrix} \Delta \phi_1 \\ \Delta \phi_2 \\ \Delta \phi_3 \\ \Delta \phi_4 \end{bmatrix}$$

The interaction of the control loops (29) can now be eliminated.
using a decoupling network \( H \) (cf. e.g., [11]). The resulting control structure is depicted in Fig.14. Based on

\[
\begin{bmatrix}
\Delta u_2 \\
\Delta u_3
\end{bmatrix} = G_H \begin{bmatrix}
\Delta \phi_u^1 \\
\Delta \phi_u^2
\end{bmatrix}
\]

the decoupling matrix can be found as

\[
H = \begin{bmatrix}
H_{11} & H_{12} \\
H_{21} & H_{22}
\end{bmatrix} = G^{-1} \begin{bmatrix}
G_{11} & G_{12} \\
G_{21} & G_{22}
\end{bmatrix}^{-1} = \frac{1}{G_{11}G_{22} - G_{12}G_{21}} \begin{bmatrix}
G_{22} & -G_{12} \\
-G_{21} & G_{11}
\end{bmatrix}
\]

V. SIMULATION RESULTS

The above theoretical considerations have been verified by digital simulations using PSIM based on the circuit schematic shown in Fig.15 assuming the following operating parameters and component values: \( V_i = 500V, V_j = 400V, V_f = 360V, L_j = L_2 = L_3 = 100\mu H, f_i = 20kHz, C_j = C_2 = C_4 = 2\mu F \).

The simulated system control behavior is shown in Fig.16. At \( t=0 \) the power reference signals are \( P_j = -3kW \) and \( P_f = -1kW \) (a negative power value indicates that power is absorbed by the corresponding port and/or converter cell, in the case at hand power is delivered from port 1 to ports 2 and 3). The power reference signal \( P_j \) is stepped to \(-4kW \) at \( t=0.02s \) and \( P_f \) is stepped to \( 0 \) at \( t=0.07s \). The control shows excellent dynamics and only a weak cross-coupling of the control loops is remaining. A detailed description of the dynamic modeling of the system and the extension of the basic control structure to controlling \( V_f \) and the battery charging current \( f_2 \) will be discussed in a future paper.

VI. CONCLUSIONS

A novel Three-Port UPS formed by linking a three-phase PWM rectifier, a three-phase PWM inverter and a battery energy storage via a single three-winding isolation transformer and corresponding full-bridge converter cells is proposed. According to the theoretical analysis and simulation results the converter system shows the following features:

1) The bidirectional power flow between the ports can be controlled by the phase-shift of the individual full-bridge cells;

2) A utilization of duty cycle control allows to operate the system with minimum power semiconductor losses

3) For employing a decoupling network, the power of the ports can be controlled independently, where operation of the system in the optimum region is ensured by proper pre-control derived from analyzing the system behavior with restriction to the fundamentals of the voltages and currents.

In a next step a DSP-controlled 5kW laboratory model of the system shown in Fig.1 will be realized for verifying the proposed control concept. Furthermore, the optimization of the system behavior will be extended to a proper selection of \( V_j \) and \( V_f \) for a given operating point.

REFERENCES


