Abstract—Power-electronics-based Medium-Voltage Direct Current (MVDC) shipboard power systems consist of several interconnected feedback-controlled switching converters. Such systems suffer from stability degradation owing to the negative incremental resistance of the Constant-Power Loads (CPLs). To tackle the stability problem, system-level stabilizing control methods were proposed acting on either the load-side or the source-side converters. In this paper, the three-phase dual-active bridge (DAB3) dc-dc converter is considered to realize the MVDC bus. A master-slave control approach is introduced which not only ensures system-level stability but also optimizes the load-sharing among the DAB3 converters. A nonlinear control technique for the master controller is compared with the PI controller with linearized feed-forward compensation. The comparison is performed through simulations in Matlab/PLECS.

Index Terms—DC-DC power converters, Voltage Control, Control systems, Negative Resistance Circuits, Dual-Active Bridge, Nonlinear control

I. INTRODUCTION

Addressing the key technical issue of bus-voltage regulation will enable the development of MVDC shipboard power systems [1]–[3]. The advantages of MVDC distribution for shipboard power systems include higher fuel efficiency, weight and space saving and flexibility in the system design [4]. In such systems, the individual or groups of loads are normally supplied through power-electronic converters which are directly connected to the MVDC distribution bus. The load converters, under tight feedback regulation, behave at their input terminals as Constant-Power Loads (CPLs) within their control-loop bandwidth. In the small-signal sense, these CPLs exhibit a negative incremental input resistance characteristics which is the origin of the undesired destabilizing effect [5], [6]. Conditions under which the instability takes place and methods for the stability analysis of MVDC systems have been reviewed and discussed in [7]. In the past, for the MVDC Integrated Ship Power Systems (ISPSs), efforts have been focused on the design of load converters, i.e. those interfacing the MVDC distribution bus and the loads. In order to prevent the MVDC ISPS from voltage instability, the negative input resistance behavior of the load converters can be relaxed through dedicated control algorithms [8], stabilizing virtual impedances [9] or through forced disconnection before the MVDC system operates out of the stable range [10]. Nevertheless, the load-side stabilizing control approach in MVDC ISPSs presents difficulties in the standardization and usage of commercial off-the-shelf converters since every load converter would have to include a special stabilizing control adapted to the present system before inserting a load. Moreover, the load-side stabilizing controllers always deal with the trade-off of stability-margin improvement vs. the feedback-bandwidth reduction [9].

Recent developments of all-electric naval vessels are based on MVDC ISPS approach. A conceptual high-performance MVDC ship system was presented as an IEEE Standard and adapted for this work [3]. In this concept, the shipboard loads are connected to the MVDC bus directly via load center or dc-dc converters. These loads might exhibit CPL characteristic which, in specific conditions, can lead to voltage instability [5]. These conditions occur due to sudden change of the operating point caused by connection or disconnection of large loads or loss of generators. In addition, the number of load-side converters is usually larger than the number of generation-side converters [3], making the source-side stabilizing control a superior solution [11].

The generating units are connected to the MVDC bus through a multi-stage power-conversion system consisting of rectifier(s) and dc-dc converter(s). Replicating the feature of terrestrial power systems on ISPSs such that the generation side is responsible for the system stability, irrespective of the load, will enable the plug&play capability for every load and the use of commercial off-the-shelf components. Furthermore, the generation-side stabilizing control allows downsizing of the input and output filters while guaranteeing the operational objectives in the presence of wide variations of load or generation capacity [4]. To this end, DAB3 is considered in this work to form the MVDC bus to which multiple loads can be connected. The galvanic isolation, the soft-switching capability over a wide operating range resulting in highly efficient operation and simple implementation make DAB3 a suitable converter for the MVDC ISPS. The DAB3s are controlled such that the bus voltage remains stable regardless of the type and connection/disconnection of loads. The paper compares a nonlinear voltage controller with the PI-based controller [12], [13] including linearized feed-forward compensation for the parallel-connected DAB3s. Moreover, an effective approach for the load-sharing based on a master-slave control of the source-side converters is presented.

This paper is organized as follows: Section II introduces the overall MVDC ISPS and the topology of DAB3 dc-dc
converter. Section III briefly presents the small-signal model of the DAB3 followed by the linearization of the transfer function at the operating point. Section IV describes the master-slave control architecture followed by the two voltage-control strategies and the inner controller with instantaneous current control. A case study is performed in Section V to investigate the stability of the controllers and to illustrate the performance and operating characteristics of the DAB3-based ISPS. Finally, the conclusions are presented in Section VI.

II. MVDC SYSTEM DESCRIPTION

Fig. 1 shows a simplified schematic of the considered MVDC ISPS taken from [3], [14], [15]. It consists of a number of multi-megawatt MV ac generators whose outputs are rectified to the nominal voltage \( U_s = 8.9 \text{kV} \). A galvanically isolated dc-dc conversion stage forms the MVDC bus with the regulated voltage of \( U_s = 6 \text{kV} \). Different loads with a total active power of 60 MW are connected to this MVDC bus. In order to realize this conversion stage, a number of parallel-connected DAB3s of Fig. 2 are considered in this work, as shown in Fig. 3. According to [14], [16], a 3% ripple of the nominal voltage of the MVDC bus is acceptable in steady state. During a transient condition, the voltage sag should not exceed 0.25 p.u [3], which corresponds to a minimum transient voltage of 4.5 kV for the case considered in this work.

The DAB3 [17], shown in Fig. 2, consists of two fully controllable three-phase bridges connected back-to-back through a medium/high-frequency transformer. One of the bridges acts as the inverter whereas the other one operates in rectification mode. Both of these bridges are switches in block-mode i.e. all of the active devices operate at 50% duty cycle. The phases within one bridge are mutually phase-shifted by 120°. The power flow between the two bridges can be controlled by phase-shifting the similar phases on both sides by an angle \( \varphi \). The power flows from the leading to the lagging bridge.

The DAB3 offers a couple of advantages which make it suitable for the aforementioned application:

- It is galvanically isolated through the transformer which electrically decouples the primary and secondary sides. Thus, a fault on either side remains on that side and does not affect the components on the other side.
- The voltage levels on the two sides can vary significantly by suitably choosing the turns ratio \( n \) of the transformer.

\[
P = \frac{U_p^2}{2\pi f_s \cdot L_o} \cdot d \cdot \varphi \cdot \left(\frac{2}{3} - \frac{\varphi}{2\pi}\right) \quad \text{for} \quad 0 \leq \varphi \leq \frac{\pi}{3}
\]

where \( L_o \) is the primary-referred leakage inductance of the transformer and the dynamic voltage ratio \( d = \frac{n U_u}{u_{cpl}} \). This is the fundamental equation of DAB3 that will be used in the next sections to model the dynamics of the multi-converter system and to develop the control structure. The operation of the converter for \( \pi/3 \leq \varphi \leq 2\pi/3 \) is not considered in this work because of presence of large reactive currents in the DAB3 transformer, potentially leading to poor system efficiency.

III. SMALL-SIGNAL DAB3 MODEL

The state-space-averaging (SSA) model of DAB3 connected to a resistive load is presented in [12], [13]. The model shows a nonlinear element in the equation which is based on the phase-shift angle \( \varphi \) [12], [13], [19]. Starting from the model in [13] and adding the resistive load and the ideal model of the CPL [5] [14] (cf. Fig. 3), the following nonlinear model equation is obtained:

\[
\frac{d u_s}{d t} = \left(\frac{2}{3} - \frac{\varphi}{2\pi}\right) \cdot d \cdot U_p - \frac{u_s}{R_L C_f} - \frac{P_{cpl}}{u_s C_f}
\]

where \( u_s \) is the instantaneous voltage across the output capacitance \( C_f \), \( R_L \) is the equivalent resistance of the linear load, \( P_{cpl} \) is the rated power of the CPL, and \( 2\pi f_s L_o \) is the inductive reactance at the switching frequency.

In order to comply with the control approach described in [12], [13], the feed-forward transformation block (cf. Fig. 7) is given by the following relation:

\[
\varphi^* = \frac{2\pi}{3} - \frac{2\pi}{3} \sqrt{1 - \frac{9 \cdot f_s \cdot L_o \cdot I_{sDC}^2}{n \cdot U_p}}
\]
The equation (3) correlates the output reference current $I_{sDC}^*$, the converter design and operational parameters with the actual control output $\varphi^*$ of the controller. The application of the linear analysis requires the system to be linearized at a certain operating point. The values for the reference current and the reference phase shift can be evaluated by solving the system around the operating point. Applying the linearization technique to (3) results in the following small-signal relation:

$$\varphi^* \simeq \frac{2\pi}{3} - \frac{2\pi}{3} \frac{1 - k \cdot I_{sDC}^*(0)}{\sqrt{1 - k \cdot I_{sDC}^*(0)}} + \frac{\pi \cdot k \cdot I_{sDC}^*(0)}{3\sqrt{1 - k \cdot I_{sDC}^*(0)}}$$

$$\simeq \frac{\pi \cdot k \cdot I_{sDC}^*(0)}{3\sqrt{1 - k \cdot I_{sDC}^*(0)}} = H \cdot I_{sDC}^*$$ (4)

where $I_{sDC}^*(0)$ is the reference current calculated at the operating point and $k = \frac{9f_s \cdot L_p}{n \cdot U_p}$.

The linearization technique is also applied to the nonlinear element of the model equation (2). Neglecting the constant term of the linearized equation results in the following small-signal relation:

$$\varphi \left( \frac{2}{3} - \frac{\varphi}{2\pi} \right) \simeq \varphi \left( \frac{2}{3} - \frac{\varphi_0}{\pi} \right)$$ (5)

where $\varphi_0$ is the phase shift calculated at the operating point. Replacing the linearized element obtained from (4) in (5) and back-substituting in (2) results in the following small-signal model of the system:

$$\frac{du_s}{dt} = H \cdot I_{sDC}^* \cdot \left( \frac{\frac{2}{3} - \frac{\varphi_0}{\pi}}{2\pi f_s \cdot L_p \cdot C_f} \right) \cdot d \cdot U_p - \frac{u_s}{R_0 C_f} + \frac{u_s}{C_f R_0}$$ (6)

where $R_0$ is the negative resistance of the CPL. The procedure for obtaining the negative resistance is defined in [20].

From (6), the control-to-output transfer function $G(s)$ is calculated as:

$$G(s) = \frac{u_s(s)}{I_{sDC}(s)} = H \cdot \frac{\frac{\frac{2}{3} - \frac{\varphi_0}{\pi}}{2\pi f_s \cdot L_p \cdot C_f} \cdot d \cdot U_p}{s + \frac{1}{\tau_f} \left( \frac{1}{C_f} - \frac{1}{R_0} \right)}$$ (7)

Equation (7) highlights that the small-signal stability condition of DAB3 only depends on the percentage of CPL load. The output capacitance of the converter does not have any additional impact on the stability and it can be designed depending on the desired characteristic of the output voltage, e.g. voltage ripple specification. The leakage inductance $L_0$ has a small effect on the stability, since it does not appear on the denominator of the transfer function. Nevertheless, the leakage inductance has to be carefully designed, given that it controls the dynamic of the transformer current and consequently the output current [12]. The small-signal model of the DAB3 in the time domain and Laplace domain, expressed in (6) and (7) respectively, will be used in the section IV-A for the definition of the stability limit of the linear controller.

IV. MASTER-SLAVE CONTROL OF THE MVDC ISPS

As stated in the previous section, the dc-dc conversion stage consists of parallel-connected DAB3 converters to form the common MVDC bus. The load side is characterized by an equivalent constant-voltage load (CVL), implemented with a resistance, and an equivalent CPL, which is realized by applying the definition of the ideal CPL to a current source [4], [14], [20]. The main goal of the voltage-control strategies applied to the DAB3 system is to maintain MVDC-bus voltage stable in the event of fast load changes or reconfiguration of the system itself, as presented in [4], [14]. The chosen control structure for this purpose is the Master-Slave Control (MSC), as shown in Fig. 3. This structure is particularly suitable for shipboard application where the distances among the master and the slave controllers are smaller than the typical terrestrial grid applications. When the distances among converters become larger such as in terrestrial application, the MSC however, does not represent an optimal solution. This is primarily due to the high dependence on the knowledge of the system configuration and topology which might be vary in a microgrid application. Moreover at larger distances, the communication of references from the master controller to the distributed converters can also be problematic because of the delays in the communication channels. Some potential techniques for compensation of the communication delay can be found in [21]–[23].

Fig. 3: Proposed master-slave control
The master controller implements the voltage control, presented in section IV-A, defining the current-control references for the individual DAB3 slave controllers. Based on on the total reference current and on the rated power of the converters, the master controller also enables the connection of the optimal number of DAB3s to the MVDC bus, as shown in Fig. 4, ensuring optimal load-power sharing. The ability of the master to manage the number of DAB3s connected to the main bus improves the reliability of the entire system. Furthermore, each DAB3 has its own slave controller, shown in Fig. 7, which implements the inner current-control loop and the instantaneous current control (ICC) [12], [24] which are briefly described in section IV-B. All these features make the MSC an interesting solution for the MVDC ISPS.

Another potential approach is the well-known droop-based control, where decoupling of the power converters and hence, the power sharing are achieved by means of a voltage drop over a fictitious resistance [11]. Compared to the MSC, the droop-based control introduces a voltage error in the steady state that has to be compensated in a centralized or decentralized manner [4], [11]. Moreover, the droop strategy does not ensure optimal current sharing among the converters since the sharing is based only on the droop coefficients. Therefore, for this specific application, the MSC is chosen as compared to the droop control.

In the proposed MSC architecture, the slave controllers do not communicate with each other. In the case of failure of the communication channel between the master and the slave controller, a converter would remain connected to the MVDC bus resulting in an unexpected and potentially undesired operation. A possible solution that can be adopted is based on the fact that the slave controllers, even though they are not directly exchanging data, are physically connected by means of the network switches. If the tree topology is adopted, the intrinsic redundant characteristic allows the slave controller to reach its neighbours even though one of the communication channel is broken [25]. On detecting a communication fault, the slave controller disconnects itself from the MVDC bus by trying to reach the circuit breaker through an available communication path. After a certain time-out, the master controller enables one of the backup converters, since the system design with MSC helps to ensure a backup converter even at full load power.

A. Voltage Control

In this section, two control techniques are discussed to control the voltage of the MVDC bus. This control is implemented in the master controller that generates the enable signals and the current references for all the DAB3 converters.

1) PI Control with Linearized Feed-Forward: Based on the linearized model of the system, described in Section III, the stability of the DAB3 controlled by a PI controller can be investigated. The equations of the closed loop system are obtained by adding the definition of the PI controller to the equation (6) resulting in the following:

\[
\frac{du_s}{dt} = H \cdot \zeta \cdot \left( U_p^* - u_s \right) + K_i \cdot \gamma - \frac{u_s}{R_1 C_1} + \frac{u_s}{R_0 C_1} \tag{8}
\]

where

\[
\frac{d\gamma}{dt} = (U_p^* - u_s) \quad \text{and} \quad \zeta = \frac{\left( \frac{2}{3} - \frac{2\omega}{\pi} \right)}{2\pi f_p L_0 C_1} \cdot d \cdot U_p
\]

and \(K_p, K_i\) are the proportional and the integral coefficients of the PI controller, whereas \(U_p^*\) is the reference value for the MVDC-bus voltage. The characteristic equation of (8) is given by the following:

\[
s^2 + \left( H \cdot \zeta \cdot K_p + \frac{1}{R_1 C_1} - \frac{1}{R_0 C_1} \right) s + H \cdot \zeta \cdot K_i = 0 \tag{9}
\]

According to the Routh–Hurwitz criterion for linear systems, the stability limits can be analyzed by setting the coefficients of \(s\) in (9) as greater than zero leading to the following relations:

\[
\begin{cases}
K_p > \left( \frac{1}{R_1} - \frac{1}{R_0} \right) \cdot \frac{2\pi f_p L_0}{H \cdot \left( \frac{2}{3} - \frac{2\omega}{\pi} \right) \cdot d \cdot U_p} \\
H \cdot \zeta \cdot K_i > 0
\end{cases}
\tag{10}
\]

If these conditions are satisfied, the system is said to be closed-loop stable. From (10) it is evident that the output filter capacitance of DAB3 converter does not affect the stability margin of the system. Taking into account the stability limits of the closed-loop system, the design of the linear voltage controller can be based only on the desired performance,
without considering the instabilities introduced by the unstable (CPL) load. The equation (10) also demonstrates that the stability conditions of (7) it is enhanced for load consisting mostly of CPL load.

As indicated in the Fig. 5, a feed-forward (FF block) compensation is introduced to improve the dynamics of the voltage controller. At each time step, the MVDC-bus voltage \( u_s \) and the total load current \( i^{\text{load}}_\text{DC} \) are measured to calculate the total equivalent load resistance \( R_{\text{eq}} = u_s/i^{\text{load}}_\text{DC} \) which is used for calculation of the feed-forward component \( I_{\text{FF}} \) of the current reference through linearization as follows:

\[
I_{\text{FF}} = \frac{U^*_s}{R_{\text{eq}}} + \frac{U^*_s - u_s}{R_{\text{eq}}}
\] (11)

This is added to the output of the linear PI controller to get the current reference \( I^{\text{ref}}_\text{ref} \), as indicated in Fig. 5. This technique ensures fast tracking of the reference and the PI controller primarily compensates for the modelling errors and removes the steady-state error.

2) Immersion and Invariance Control: Since the model of the DAB3 is nonlinear, the application of an advanced control can make use of the nonlinearities of the system. Moreover, the introduction of the nonlinear controller permits the understanding of the applicability of linear controller for the nonlinear system. In this work, the Immersion and Invariance (I&I) controller is considered for the voltage control of the DAB3. The I&I control is adopted because it guarantees the global asymptotic stability (GAS) by choosing a global asymptotically stable target system. However, the controller is characterized by a high dependence on the system parameters which reduces its robustness [26]. A potential alternative is the Sliding Mode control (SMC) which is theoretically more robust because of less dependence on the system parameters. However, the standard formulation of SMC demonstrates some difficulties in the practical applications primarily because of the chattering phenomena that can degrade the quality of the control [27]. In order to solve this problem, many different adjustments to the first formulation of the SMC have been derived which comprehend the use of a high-order SMC [26], methods for controlling the duty cycle [28] [29] and prediction methods [30] [31]. All these methods are model dependent which reduces the theoretically high robustness of the SMC. Moreover, the authors in [26] have demonstrated that the high-order SMC is a special case of the I&I principle. Therefore, the I&I control has been adopted in this work owing to its intrinsic GAS characteristic which can be ensured by following the calculation steps in [32].

The I&I control technique is based on the definition of an invariant and attractive manifold, as described in [32], and in its discrete form in [33]. It begins with the definition of a lower-order and global asymptotically stable system \( \tilde{\xi} = \alpha(\xi) \), which is the target system of the I&I control strategy. By means of an immersion \( \pi: \xi \rightarrow x \), the application of the I&I guarantees that every trajectory \( x(t) \) of the system is the equivalent image in \( \tilde{\xi}(t) \), through the mapping function. Therefore, given the asymptotic stability of the target system, the immersed system is also asymptotically stable. Additionally, the manifold must be held invariant and attractive by defining a proper control law, that has to keep also the trajectory bounded. An interesting result of this control method, as many other manifold-based control strategies [34], [35], is that the definition of the control law is not strictly related with the evaluation of the Lyapunov function for the system. The formulation of the control output is based on the steps defined in [32], [33], which guarantee that if all the hypotheses are satisfied, the non-linear system is I&I-stabilizable with the target dynamic \( \dot{\xi} = \alpha(\xi) \).

The system to be stabilized is described by the nonlinear equation (2) and by the definition of an auxiliary state variable \( M \), which is the integral of the voltage error. This variable has been introduced in order to add an integral action on the control law. The resulting system is then described by:

\[
\begin{align*}
\frac{dM}{dt} &= K \cdot (u_s - U^*_s) \\
\frac{du_s}{dt} &= \frac{\varphi(\frac{\xi}{R_{\text{eq}}} - \frac{\xi}{R_{\text{eq}}})}{2\pi\int f_s \cdot \sigma_s} \cdot d \cdot U_p - \frac{u_s}{R_{\text{eq}}} \frac{\rho_{\text{eq}}}{u_s} + M
\end{align*}
\] (12)

where \( K \) is the coefficient of the integration. The system is rewritten with the new variables \( x_1 = M - M^* \) and \( x_2 = u_s - U^*_s \). Since the integral of the voltage error should be zero in steady state, \( M^* = 0 \). Therefore, the new system is described by the following equations:

\[
\begin{align*}
\frac{dx_1}{dt} &= K \cdot x_2 \\
\frac{dx_2}{dt} &= \frac{\varphi(\frac{\xi}{R_{\text{eq}}})}{2\pi\int f_s \cdot \sigma_s} \cdot d \cdot U_p - \frac{x_2}{R_{\text{eq}}} \frac{\rho_{\text{eq}}}{x_2 R_{\text{eq}}} + x_1
\end{align*}
\] (13)

The first element to be introduced for the calculation of the control law is the target system and its dynamic. The I&I control theory does not define any restriction on the definition of the target system, but for a non-linear system, a linear target system is generally applied [32]. The linear target system that has been used is given by:

\[
\dot{\xi} = -\xi
\] (14)

which has a global asymptotically stable equilibrium point in \( \xi^* = 0 \). By fixing \( \Pi_1(\xi) = \xi \), where \( s(\xi) \) is the mapping function, the equilibrium point is \( \Pi_1(\xi^*) = 0 = M^* \). Therefore, the first hypothesis of [32], [33] holds. Considering the mapping function, the system is rewritten in this form, where \( c(\Pi(\xi)) \) is the control input of the system in the new formulation:

\[
\begin{align*}
\Pi_1 &= -\xi = K \cdot \Pi_2 \\
\Pi_2 &= \frac{\partial \Pi_2}{\partial \xi} \cdot \dot{\xi} = c(\Pi(\xi)) \cdot \left( \frac{2}{\rho_{\text{eq}}} - \frac{c(\Pi(\xi))}{\pi} \right) \cdot d \cdot U_p \frac{2}{\pi} f_s \cdot \sigma_s \cdot L_{\text{eq}} \cdot C_t \\
&\quad - \frac{\Pi_2}{R_{\text{eq}}} \frac{\rho_{\text{eq}}}{\Pi_2 R_{\text{eq}}} + \Pi_1
\end{align*}
\] (15)

By applying the theory defined in [32], the implicit definition of the manifold can be described as:

\[
\rho(x) = x_2 - \Pi_2 \left( \Pi_1^{-1}(x_1) \right) = 0
\] (16)

It has been explained in [32] that the manifold is not explicitly defined, but described by setting the function (16) equal to zero. This means that the goal of the control law is to drive the off-the-manifold trajectories of the system to zero, which are described by \( z = \rho(x) \).
From the first equation of the system (15), the description of \( \Pi_2 = -\xi/K \) is obtained. Therefore the implicit definition of the manifold becomes \( \rho(x) = x_2 + x_1/K = 0 \) and the off-the-manifold dynamic is defined by the following equation, where \( \Psi(x,z) \) defines the control output that makes the system \( I & J \)-stabilizable [32]:

\[
\dot{z} = \rho(x) = \Psi(x,z) \cdot \left( \frac{2}{3} - \frac{\Psi(x,z)}{2n} \right) \cdot d \cdot U_P - \frac{x_2}{R_{dc}C_1} + x_1 + x_2
\]

In order to have bounded off-the-manifold trajectories [32], [33], this equation is set to be \( \dot{z} = -z \). The control law is then calculated by substituting \( \rho(x) = z = x_2 + x_1/K \) and solving the resulting equation for \( \Psi(x,\rho(x)) \). Since the output is actually the reference angle \( \varphi^* \), the result is substituted in (3) and brought back to the original state variables \( u_s \) and \( M \), obtaining the control output \( I_{dc}^* \).

To complete the control method, the last hypothesis presented in [32], [33] must be verified, which says that the trajectories of the closed-loop system must be bounded and they must satisfy \( \lim_{t \to \infty} z(t) = 0 \). In order to make the calculation of the closed-loop system easier and more intuitive, the term \( \eta \) has been defined as:

\[
\eta = \Psi(x,\rho(x)) \cdot \left( \frac{2}{3} - \frac{\Psi(x,\rho(x))}{2n} \right) \cdot d \cdot U_P
\]

By using (17), \( \eta \) can be described by the following equation:

\[
\eta = \frac{x_2}{R_{dc}C_1} + \frac{P_{cpl}}{x_2C_1} - x_1 - x_2 - z
\]

The equation (19) is substituted into (13) and the dynamics of the closed loop system can be expressed as:

\[
\begin{cases}
\dot{z} = -z \\
\dot{x}_1 = Kx_2 \\
\dot{x}_2 = -x_2 - z
\end{cases}
\]

The trajectories of the system are clearly bounded and \( \lim_{t \to \infty} z(t) = 0 \) is satisfied. Therefore \( x^* = [0, U^*_s] \) is a global asymptotically stable equilibrium of the closed-loop system. The scheme of the nonlinear voltage controller is represented in Fig. 6.

**TABLE I: DAB3 Parameters**

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( P_s )</td>
<td>8 MW</td>
</tr>
<tr>
<td>( f_s )</td>
<td>1 kHz</td>
</tr>
<tr>
<td>( U_P )</td>
<td>8.9 kV</td>
</tr>
<tr>
<td>( U_s )</td>
<td>6 kV</td>
</tr>
<tr>
<td>( R_{dc}, R_a )</td>
<td>5.4 mΩ, 2.4 mΩ</td>
</tr>
<tr>
<td>( L_a )</td>
<td>380 µH</td>
</tr>
<tr>
<td>( C_t )</td>
<td>600 µF</td>
</tr>
</tbody>
</table>

**TABLE II: Controllers Parameters**

<table>
<thead>
<tr>
<th>Proportional Action</th>
<th>Integral Action</th>
</tr>
</thead>
<tbody>
<tr>
<td>( K_p = 1 )</td>
<td>( K_i = 1 )</td>
</tr>
<tr>
<td>( 1/k )</td>
<td>( K = 1 )</td>
</tr>
</tbody>
</table>

**TABLE III: Controls’ performance with ICC**

<table>
<thead>
<tr>
<th>Connections</th>
<th>20 MW CVL</th>
<th>20 MW CPL</th>
</tr>
</thead>
<tbody>
<tr>
<td>Controls</td>
<td>PI</td>
<td>I&amp;I</td>
</tr>
<tr>
<td>Overshoot (in %)</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>Undershoot (in %)</td>
<td>7.4</td>
<td>6.75</td>
</tr>
<tr>
<td>Settling time (in ms)</td>
<td>5.1</td>
<td>2.3</td>
</tr>
</tbody>
</table>

**B. Inner Current Control and Instantaneous Current Control**

The inner current-control loop (CCL) calculates the required phase-shift angle for the \( \theta \)th DAB3 converter in order to obtain the reference current \( I_{dc}^* \) at the output of that converter. Like the voltage-control loop, the CCL has a feed-forward component and a PI compensator, as indicated in Fig. 7. The feed-forward block uses (3) to calculate \( \varphi^* \). The PI controller compensates for the modelling and steady-state errors.

As indicated in [12], [24], an abrupt change in the phase-shift angle could lead to oscillations of the transformer phase currents which also manifest in the form of overshoot and oscillations in the dc-bus voltage. In order to balance the phase currents during the transient interval and hence achieve better MVDC-bus voltage regulation, the method II of [12] for the instantaneous current control is employed in this work. This method adjusts the switching instants of different devices of DAB3 in order to counter the offset voltages applied across the transformer phase windings during the transient interval. The details of this method can be looked up in [12], [24].

**V. SIMULATION RESULTS**

The MVDC system, introduced in Section II, is implemented and simulated in Matlab/PLECS framework. Each DAB3 is rated for a nominal power of \( P_s = 8 MW \) considering the largest Injection-Enhanced Gate Transistors (IEGTs) [36] available in the market. IEGTs are IGBTs capable of handling larger currents with lower collector-emitter on-state voltage. The parallel-connected DAB3s are connected on the primary side to a voltage source of 8.9 kV whereas the secondary side, connected to the main MVDC bus, is maintained at 6 kV by the voltage controller. The total load power of 60 MW is divided among the eight DAB3 converters. The
parameters of the DAB3 converters are defined in Table I where \(R_p, R_s\) are the primary and secondary transformer winding resistances and \(L_a\) is the primary-referred total leakage inductance of the DAB3 transformer.

The control algorithms are developed as scripts by means of the C-script building block in PLECS. Therefore, the control laws are discretized with a time step \(T_s = 1/\omega_s = 1\) ms, applying the typical discretization of linear controller for the modified PI controller. The discretization of the I&I controller leverages the Euler's discretization of nonlinear continuous-time systems. The controllers' parameters are given in Table II, where the integrator coefficient \(K\) of the I&I control is chosen equal to the coefficient \(K_1\) of the linear PI controller. This allows a fair comparison among the two control laws, since the two controllers exhibit the same integral action. The implementation of the master-slave control follows the scheme presented in Section IV and has been carried out with C-script block in PLECS.

Fig. 8 and Fig. 9 show some system variables with the PI and I&I controller, respectively, under similar load variations. The test starts with a base load of 15 MW of resistive load. At time \(t = 6\) ms, a CVL of 20 MW, modelled with a resistive load, is connected and at \(t = 36\) ms, a CPL of 20 MW is connected to the MVDC bus amounting to a total load power of 55 MW. Afterwards, at \(t = 66\) ms, a ramped disconnection of loads brings the total load power to 20 MW in 40 ms. The top plot shows the MVDC-bus voltage \(U_{act}^s\) compared to the reference voltage \(U_{act}^s\). The middle plot shows the total instantaneous load power. The bottom plot shows, as an example, the output dc current \(i_{DC1}\) for the 1st, 5th and 8th DAB3 converter.

As it can be noticed from the figures, the PI control with feed-forward and the I&I control are both able to achieve a fast restoration of the voltage in case of large load reconfigurations, maintaining the constraints presented in section II. Their performances are both satisfying, even though the I&I control exhibits a slightly faster response during transient time. The comparison of the two control strategies in term of overshoot, undershoot and settling time is represented in Table III, where the settling time is calculated as the time to reach the \(\pm 3\%\) of voltage deviation, as described in [14], [16].

Another aspect that can be highlighted with the simulation results is the effectiveness of the master-slave approach. For demonstrating that, only the plot of output currents suffices since it shows the connection of different number of converters based on the load level. At the beginning of the simulation, the total amount of load is more than the nominal power of a single DAB3 but less than two. Therefore, the master control, based on the reference output current, connects only
the minimum number of converters needed, in this case only two. Afterwards, at the connection of the first CVL, the algorithm connects five converters, and finally at 55 MW, seven out of eight DAB3 converters are connected to the main bus. The effect of the connection and disconnection of the DAB3s can be observed in the bottommost plots of Fig. 8 and Fig. 9.

As can be seen, the eighth DAB3 turns on momentarily at the connection of CPL at 36 ms in order to meet the high current demand to restore the MVDC-bus voltage. Since, in a real system, turning on and off of such a converter for a couple of milliseconds is not feasible, a hysteresis can be implemented to avoid such turn-on/off the converters.

It is also worth noticing and mentioning here that because of the instantaneous current control of each DAB3, there are no noticeable overshoots of the dc-side currents. In order to prove this, the same load variations are performed after turning the ICC off in the inner CCL. The results for the two controllers are presented in the Fig. 10 and Fig. 11 respectively. Comparing with the previous figures with ICC (cf. Fig. 8 and Fig. 9), it is clear that without ICC, not only are the semiconductor devices stressed more because of the high overshoots in the phase currents but also the dc-bus capacitors as well. Moreover, the dynamics of the system are negatively affected in terms of the settling time, as shown in Table IV. In order to evaluate the effect of system parameters on the control performance, the value of $L_\alpha$ is varied with a normal distribution around the nominal value defined in Table I, with a standard deviation of 10%. The resulting $L_\alpha$ for the eight DAB3 converters is given in Table V. The controllers however calculate the control output based the nominal value of $L_\alpha$. The MVDC bus voltage is given in Fig. 8 and Fig. 9 in the top plot as $U_{\text{s, dist}}^{\text{act}}$. The plot shows that the variation of $L_\alpha$ introduces a small difference on the bus voltage, primarily because the feed-forward transformation block, described in section III, uses a value of $L_\alpha$ which is different from the real system. However, it has a complicated control law whose implementation might be challenging in a digital controller.

VI. CONCLUSION

This paper presented the application of DAB3 converters in a parallel-connected configuration supplying an MVDC bus in an ISPS. The control of the system is carried out through a master-slave configuration and the linear and a nonlinear controller for the voltage control have been compared. Moreover, the previous enhancements, like ICC of the DAB3 [12], [24], are implemented in order to make this investigation inline with the recent developments in the DAB3 performance.

It has been demonstrated that the linearized system controlled by the PI control is closed-loop stable in presence of a CPL. The linear and the nonlinear controllers show comparable performance. The nonlinear I&I controller has the advantage of ensuring, by definition, the global asymptotic stability of the system. However, it has a complicated control law whose implementation might be challenging in a digital controller.
In order to validate the robustness of the controller, a simulation with normally distributed values of the transformers' leakage inductance is performed. The results demonstrated that it had a limited impact on the controllers' performance. Some suggestions are given on the practical implementation of the proposed controllers on digital devices.

The paper has also demonstrated the applicability of the parallel-connected DAB3s for shipboard application as an alternative to the buck converter of [4], [14] in an MVDC ISPS. Maintaining the central voltage-control approach, the master-slave configuration allows the collection, processing and analysis of all the relevant data in a single controller. It also ensures stable system operation under load reconfigurations resulting in converter(s) connection/disconnection. Furthermore, the master-slave approach presented in the paper allows the DAB3 converter to work in the optimal conditions, reducing the overall losses and maintaining always the optimal number of DAB3s connected to the MVDC bus.

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REFERENCES


TABLE IV: Controls’ performance without ICC

<table>
<thead>
<tr>
<th>Connections</th>
<th>20 MW CVL</th>
<th>20 MW CPL</th>
</tr>
</thead>
<tbody>
<tr>
<td>Controls</td>
<td>PI</td>
<td>PI</td>
</tr>
<tr>
<td>Overshoot (in %)</td>
<td>0.56 0.88</td>
<td>1.25 1.5</td>
</tr>
<tr>
<td>Undershoot (in %)</td>
<td>6.38 6.31</td>
<td>6.63 6.58</td>
</tr>
<tr>
<td>Settling time (in ms)</td>
<td>4.6 4.7</td>
<td>15.9 14.9</td>
</tr>
</tbody>
</table>

TABLE V: Values of $L_o$ in µH

<table>
<thead>
<tr>
<th>DAB</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
<th>8</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_o$</td>
<td>371</td>
<td>378</td>
<td>395</td>
<td>385</td>
<td>370</td>
<td>398</td>
<td>370</td>
<td>389</td>
</tr>
</tbody>
</table>

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