Modeling and Voltage Control of Bidirectional Resonant DC/DC Converter for Application in Marine Power Systems

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Abstract: DC/DC converters have gained popularity in a number of industrial applications like electric vehicles or marine power systems, due to their high efficiency and power density levels. Pulse width modulation (PWM) and resonant converters are two main types of the DC/DC converters. Thereby, the resonant converters happen to be the preferred technology in the design of modern marine power systems since these converters are more suitable for the high and middle voltage DC applications. The resonant converters are, however, highly nonlinear systems, which limits the use of linear control methods. In this study, we propose a comprehensive analysis, modeling and control concept of a DC/DC resonant converter in marine power systems. First, a mathematical model of the DC/DC resonant converter in the so-called CLLC topology is derived based on the generalized state-space averaging method. The model is used to design a dual-loop voltage control, which aims to regulate the voltage level at the low-voltage DC bus of the resonant converter. The dual-loop voltage control consists of the primal linear controller, which directly regulates the voltage and the reference generator, which dynamically modifies the voltage reference of the primal controller. The major advantage of the suggested control concept is the improved performance of the simple controller without the need to substitute it as well as the possibility to realize, if required, a multi-rate control concept. Simulation studies under different load conditions show that the suggested modeling and control concept improves voltage control and the closed-loop system response.

Keywords: Marine power systems, Resonant converter, Generalized average model, Reference generator, Model predictive control

1. INTRODUCTION

In order to meet the request for increased efficiency and flexibility, the power systems today are subject to significant infrastructural and operational changes. Some of these changes include increased penetration of renewable energy sources, installation of distributed storage systems, distributed electricity generation or the change of the communication and control architecture. When referring to the power networks in the marine systems, there is a notable interest to replace the existing AC grids with DC networks. As reported in Hansen et al. (2011), the onboard DC grids have series of advantages over AC grids such as fuel savings up to 20 %, more flexible placement of the electric components, improved overall dynamic response and reduced maintenance of the existing engines through more efficient operation. The major change in DC dominated marine power systems is that the frequency line transformers, which are required in AC grids, are replaced by the DC/DC converters. These converters are featured by high efficiency and power density and have a key role as interfacing components between different voltage levels. In particular, in the marine power systems, the resonant converters appear as favored technical solution among the DC/DC converters (Agamy et al. (2017)). Resonant converters come in different topologies and offer a series of practical advantages over conventional non-resonant topologies (like the dual active bridge). Some of these advantages include soft-switching of all switches, capability to operate at a very high switching frequencies, high efficiency and reduced electromagnetic interference.

1.1 Main Contribution

In contrast to the practical advantages the resonant converters offer and the popularity they gained, there are also challenges related to their control. To achieve the voltage control in the power converters, it is a common practice to apply linear control methods, like proportional-integral (PI) control, which assume a linear system representation. These methods are suitable for the fast execution (the sampling time of the converter controllers is in the range of several kHz) and easy to tune by practitioners. The direct application of the linear control methods is not possible in case of the resonant converters, due to the highly nonlinear nature of these power electronic components. For the PWM converters, like the buck-boost DC/DC converter, a simple, linear model is gained using the well-established state-space averaging method in Erickson and Maksimovic (2007). This method relies upon the assumption of small current ripple, which enables to represent the state variables in the model by their averaged values over the switching period, see Zhang et al. (2017). For the resonant converters, which operate at much higher frequencies than...
PWM converters, this assumption is not valid. The modeling of resonant converters is more challenging due to the presence of two different system dynamics, i.e., dynamics of the DC and AC states respectively. The slower changing, DC states are related to the input and output capacitor of the converter, while the fast changing AC states refer to the currents and voltages of the so-called resonant network (see Fig. 1). The increased number of state variables in the mathematical model of the resonant converter simultaneously implies the need for an increased number of measurements to design a control concept. Apart from the modeling challenges, during controller design, we need to consider that these converters do not operate at a single switching frequency — thus, their performance is very sensitive to the selection of this frequency. Therefore, the analysis of the frequency range needs to be conducted first and the control scheme should enable the change of the switching frequency during simulation.

In this work, we consider a bidirectional DC/DC resonant converter for the application in the marine power systems and design a modeling and control concept to regulate the output voltage. This converter serves as an interfacing power electronic component between the high and the low voltage DC bus of the ship power system. We assume a number of DC loads to be connected in parallel to the low voltage DC bus. The DC loads might be turned on and off, changing thereby the total load power during operation. The control problem is to assure that the desired value of the output voltage is pertained and significant voltage drops are avoided in spite of the switching between different load levels. To stabilize the voltage and improve the dynamic response of the converter in closed loop, we design a model-based control concept.

First, we derive a mathematical model of the DC/DC converter in the so called CLLC topology (Jung et al. (2013), Zahid et al. (2015), Zou et al. (2019)), which is suitable for the controller design. Thereby, we apply the generalized state-space averaging method (Sanders et al. (1991)), which relies on the expansion of the periodic functions into Fourier series. To the best authors’ knowledge, using this methodology, a control-oriented state-space model for the CLLC topology has not been suggested yet. Such approach facilitates the representation of the AC system dynamics, which are not negligible at the high switching frequencies of the CLLC converter. In the second step, we propose a control concept for the voltage control at the low voltage bus. The control concept assumes that a standard PI controller is applied to directly control the voltage at the low-voltage bus of the converter by adjusting the frequency. This frequency serves as an input to the frequency modulator which drives the converter internal half-bridges (H-bridges). Instead of replacing the PI controller with the more advanced one, like a model predictive controller (Maciejowski (2002), Bordons and Montero (2015), Kouro et al. (2015)), we pursue a more practical approach and design a double loop concept with an add-on scheme in the outer loop, which supports the primal controller in the inner loop. Such control concept design might be particularly useful in the power converter field, where the sampling frequencies are high and simple primal control approaches are preferred for hardware implementation. The resonant converter in our application can have a switching frequency of up to 20 kHz. As such, it is significant to reduce the complexity of the primal control algorithm and assure its fast execution. Furthermore, in some cases, the controller is already embedded, cannot be modified and we need to come up with a solution to improve the performance without replacing the existing controller. The main task of the add-on scheme, which exploits the idea of reference governors in Kolmanovsky et al. (2014), is to adapt the voltage reference of the PI controller by solving an unconstrained optimal control problem at each sampling time. The reference generator leverages the mathematical model of the CLLC resonant converter to form a closed-loop system model, to predict the time evolution of the closed-loop system and to use this prediction in the definition of the optimal control problem. The generated reference signal corresponds to an artificial reference which should ensure that the actual voltage reference value is reached (under certain performance criteria) for a broad range of load conditions. The adoption of the CLLC state-space model (with Fourier coefficients as AC states) for model-based reference generation has not yet been considered in previous works.

The paper is structured as follows. In Section 2 we introduce the CLLC topology, its properties and derive the characteristic curve for the frequency response amplitude. Section 3 derives the open-loop model of the CLLC converter. The control concept including the derivation of the closed-loop model and the reference generator design is presented in Section 4. In Section 5 simulation results are presented and discussed. Finally, Section 6 concludes and provides an outlook for future work.

2. BI DIRECTIONAL DC/DC RESONANT CONVERTER

2.1 Converter Topology

Fig. 1 shows the circuit topology of the bidirectional resonant converter applied in the middle voltage - low voltage (MV-LV) grid of the considered vessel system. The converter consists of two H-bridges and a resonant network which includes a high-frequency transformer. Thereby the primary bridge comprises the switches \( S1 – S4 \), while the switches of the secondary bridge include the components \( S01 – S04 \). Power switches of the primary bridge convert the power from DC to AC, while the secondary bridge has a rectifying role, converting the power from the AC back to the DC. The transformer \( Tp \) enables galvanic isolation between the primary and the secondary side.

In our application, the transformer’s turns ratio is defined as 1:1. Other components of the resonant network include resonant inductances \( L_{S1} \) and \( L_{S2} \), transformer magnetizing inductance \( L_m \) and the resonant capacitors \( C_{S1} \) and \( C_{S2} \). Since the resonant network is designed such that \( L_{S1} = L_{S2} \) and \( C_{S1} = C_{S2} \), it has a symmetric structure and the resulting converter is denoted as DC/DC converter in symmetric CLLC topology. With \( v_p \) and \( v_s \), we denote the voltage at the input and output of the resonant network, respectively (with \( v_s \) being phase-shifted against \( v_p \)).

2.2 Equivalent Circuit Representation

When the load conditions change, the switching frequency \( f_s \) of the resonant converter might have to be adjusted in
Recall that the designed converter should meet the voltage gain requirements in both directions of the energy transfer. To come up with the frequency region for our application, three frequency regions need to be identified, i.e., \( f < f_{r1} \), \( f_{r2} \leq f \leq f_{r1} \), and \( f > f_{r2} \). As a first step, we define the following impedances:

\[
\begin{align*}
Z_1(j\omega_s) &= j\omega_s L_{S1} + \frac{1}{j\omega_s C_{S1}}, \\
Z_2(j\omega_s) &= j\omega_s L_{m} + \frac{1}{j\omega_s C_{S2}}
\end{align*}
\]

(1) \( Z_3(j\omega_s) = j\omega_s L_{S2} + \frac{1}{j\omega_s C_{S2}} \) and \( Z_4(j\omega_s) = R_{0,e} \).

(2)

With the impedances (1)-(2), the forward frequency response of the resonant network, that is,

\[
H_f(j\omega_s) = \frac{V_p(j\omega_s)}{V_p(j\omega_s)}
\]

(3)

can be derived by transferring the impedances \( Z_3 \) and \( Z_4 \) from the secondary to the primary side of the transformer and by applying Kirchhoff’s 1st and 2nd law on the resulting resonant network circuit. This way, by expanding (3), the resulting frequency response can be summarized as:

\[
H_f(j\omega_s) = \frac{Z_2 Z_1}{Z_1 Z_2 + Z_1 Z_3 + Z_1 Z_4 + Z_2 Z_3 + Z_2 Z_4}
\]

(4)

where, for notational convenience, the dependence on \( j\omega_s \) has been omitted. For the further analysis, it is common practice to express the frequency response (4) in terms of the characteristic parameters of the resonant converter. These are the first resonant frequency \( f_{r1} \), the normalized frequency \( f_n \), that is,

\[
f_{r1} = \frac{1}{2\pi \sqrt{L C}}, \quad f_n = \frac{f}{f_{r1}},
\]

(5)

the second resonant frequency

\[
f_{r2} = \frac{1}{2\pi \sqrt{(L + L_m)/C}},
\]

(6)

the quality factor (providing an indication on the load size)

\[
Q = \frac{Z_r}{R_{0,e}},
\]

(7)

with the converter impedance \( Z_r \) being defined as

\[
Z_r = \frac{L}{C}
\]

(8)

and the inductance ratio

\[
k = \frac{L}{L_m}.
\]

(9)

Due to symmetric structure of the resonant network, it holds \( L = L_{S1} = L_{S2} \) and \( C = C_{S1} = C_{S2} \). By exploiting (5)-(9) to substitute parameters in (4), we come up with a representation of \( H_f(j\omega_s) \) that solely depends on \( f_n \) and \( Q \). That said, we define the numerator \( N(f_n) \) and denominator \( D(f_n,k,Q) \) of \( H_f(j\omega_s) \), i.e.,

\[
N(f_n) = j f_n^3
\]

(10)

\[
D(f_n,k,Q) = Q k + f_n k - 2 Q f_n^2 + Q f_n^4 k + j Q f_n^3 k + j f_n^2 + j 2 Q f_n^2 k
\]

(11)

Eventually the resulting frequency response is given by

\[
H_f(f_n,k,Q) = \frac{N(f_n)}{D(f_n,k,Q)}.
\]

(12)

The DC/DC voltage gain characteristic follows from the amplitude of the frequency response (12):

\[
|H_f(f_n,k,Q)| = \frac{f_n^3}{\sqrt{(D_{Rc}(f_n,k,Q))^2 + (D_{Im}(f_n,k,Q))^2}}
\]

(13)

with \( D_{Rc}(f_n,k,Q) = k (Q + f_n) - Q f_n^2 (2 + k) \),

\[
D_{Im}(f_n,k,Q) = Q f_n^2 (2 k + 1) + f_n^2 (1 + k)
\]

(14)

The gain characteristics in Fig. 3 shows how the gain curve scales for different load values. By analyzing this characteristic we can determine the frequency range in which the converter should be operated for the given load variations such that the voltage at the output can be regulated to the desired value. Three frequency regions can be recognized from the gain curves: Region 1 – the frequency region with frequencies below the second resonant frequency, i.e., \( f < f_{r2} \); Region 2 – between the first and the second resonant frequency, i.e., \( f_{r2} \leq f \leq f_{r1} \); and Region 3 – the one with frequencies above the first resonant frequency, i.e., \( f > f_{r1} \). The converter is not operated in Region 1 because the switches in the primary side suffer from significant losses.

To come up with the frequency region for our application, we recall that the designed converter should meet the voltage gain requirements in both directions of the energy transfer.
a generalized state-space averaging method. The method has previously been used to derive a model of a dual active
bridge (Qin and Kimball (2012), Mueller and Kimball (2018b), Mueller and Kimball (2018a)), but so far no appli-
cations for the CLLC topology have been suggested. The
generalized averaging technique relies on the expansion of
the periodic functions in the Fourier series and on using
more terms of this series than only a DC term. In this
way, a more accurate model is gained, but the order of the
model increases.

In the generalized averaged approach, the state \( x(t) \)
is expanded in a Fourier series:

\[
x(t) = \sum_{k=-\infty}^{k=\infty} \langle x \rangle_k(t) e^{j k \omega_s t}
\]

where \( \omega_s = 2 \pi f_s \) and the complex number \( \langle x \rangle_k \) is the \( k \)-th
coefficient in the Fourier series which equals:

\[
\langle x \rangle_k(t) = \frac{1}{T} \int_{-T}^{t} x(s) e^{-j k \omega_s t} \, ds
\]

\[
= \frac{1}{T} \int_{1-T}^{t} x(t) \cos(k \omega_s t) \, dt
\]

\[
- j \frac{1}{T} \int_{1-T}^{t} x(t) \sin(k \omega_s t) \, dt
\]

In the following, a first harmonic approximation is used
such that apart from the term \( k = 0 \) we include the terms
\( k = \pm 1 \) to represent the state variables, while all other
terms in the Fourier series are neglected. Every state of
the system is thus approximated as follows:

\[
x(t) = (x_0(t) + (x_1)_1(t) e^{j \omega_s t}) + (x_1)_2(t) e^{-j \omega_s t}
\]

\[
= (x_0) + 2 \langle x \rangle_1R \cos(\omega_s t) - 2 \langle x \rangle_1I \sin(\omega_s t)
\]

where \( x_1R \) and \( x_1I \) are the real and imaginary components
of the first Fourier coefficient, respectively.

3.2 Open-Loop CLLC Model

To derive the nonlinear state-space model, let \( s_1(d, \tau) \)
and \( s_2(d, \tau) \) denote the switching functions of the primary
and the secondary bridge for the switching period \( T \). The
voltage at the output of the primary bridge \( v_p \) is then
defined by:

\[
v_p(\tau) = s_1(d, \tau) v_{in}(\tau)
\]

while the voltage at the input of the secondary bridge
equals:

\[
v_s(\tau) = s_2(d, \tau) v_p(\tau)
\]

where \( \tau \) is the time within the switching period and \( d \) is
the phase shift of the converter defined as a fraction of \( \pi \),
i.e., \( d = \Phi/\pi \in [0,1] \) with \( \Phi \) being the phase shift angle.
The form of the switching functions \( s_1 \) and \( s_2 \) depends
on the modulation strategy applied. For the single-phase
shift modulation strategy (Mueller and Kimball (2018a)),
considered in this work, the signals \( s_1 \) and \( s_2 \) are defined as:

\[
s_1 = \begin{cases} 1, & 0 \leq \tau \leq \frac{T}{2} \\ -1, & \frac{T}{2} \leq \tau \leq T \end{cases}
\]

\[
s_2 = \begin{cases} 1, & \frac{dT}{2} \leq \tau \leq \frac{dT}{2} + \frac{T}{2} \\ -1, & 0 \leq \tau \leq \frac{T}{2} \text{ or } \frac{dT}{2} + \frac{T}{2} \leq \tau \leq T \end{cases}
\]

We refer to the equivalent circuit representation given in
Fig. 2 and introduce the following simplification. First,
we assume that the magnetizing inductance of the transformer \( L_m \) can be neglected. Second, the inductance \( L_{eq} \) and capacitance \( C_{eq} \) can be determined as a series connection of corresponding components, i.e.: 
\[
L_{eq} = L_{S1} + L_{S2} = 2L, \quad C_{eq} = \frac{C_{S1}C_{S2}}{C_{S1} + C_{S2}} = \frac{C}{2}
\]  
(21)

This representation simplifies the structure of the resonant network. By considering simplified circuit representations and by leveraging the definitions for \( v_i \) and \( v_c \) in (17)-(18) the following state space equations for the CLLC resonant converter are gained:
\[
\dot{v}_{C1} = -\frac{1}{C_1 R_1} v_{C1} - \frac{1}{C_1} s_1 i_t + \frac{1}{C_1 R_1} v_{\text{in}} \tag{22}
\]
\[
\dot{i}_t = -\frac{1}{L_{eq}} s_1 v_{C1} - \frac{1}{L_{eq}} v_{Ceq} - \frac{1}{L_{eq}} s_2 v_o \tag{23}
\]
\[
\dot{v}_{Ceq} = -\frac{1}{C_{eq}} i_t \tag{24}
\]
\[
\dot{v}_0 = \frac{1}{C_2} s_2 i_t - \frac{1}{C_2 R_0} v_o \tag{25}
\]

In this state-space representation, the input capacitor voltage \( v_{C1} \), the output capacitor voltage \( v_{Ceq} \), the transformer current \( i_t \) and the voltage across the equivalent capacitance \( v_{Ceq} \) are identified as system states. Due to parallel connection, the output capacitor voltage equals the load voltage such that in the following we use \( v_o = v_{Ceq} \).

We can now apply the generalized state-space averaging method on equations (22)-(25). As it follows from (16), every state of the CLLC converter model is a function of the DC average (zero-th Fourier coefficient), of the real and imaginary part of the 1st coefficient in the Fourier series. Therefore, each of these components represents an internal state in the full-order state-space model. The complete state vector can now be summarized as:
\[
x_{OL} = \left[ \begin{array}{c} \dot{v}_{C1} \  \\ \dot{i}_t \  \\ \dot{v}_{Ceq} \  \\ \dot{v}_0 \end{array} \right] \tag{26}
\]

with the subvectors
\[
\dot{v}_{C1} = [ {\langle v_{C1} \rangle}_0, \ {\langle v_{C1} \rangle}_1R, \ {\langle v_{C1} \rangle}_1I], \\
\dot{i}_t = [ {\langle i_t \rangle}_1R, \ {\langle i_t \rangle}_1I], \\
\dot{v}_{Ceq} = [ {\langle v_{Ceq} \rangle}_0, \ {\langle v_{Ceq} \rangle}_1R, \ {\langle v_{Ceq} \rangle}_1I], \\
\dot{v}_0 = [ \ {\langle v_o \rangle}_0, \ {\langle v_o \rangle}_1R, \ {\langle v_o \rangle}_1I].
\]

Simultaneous analysis of the full order model shows a potential for model reduction. The states in the model can be classified into DC states and AC states. The capacitors at the input and the output of the converter are such that the voltages across these elements can be represented by their average components. In a similar way, the internal states related to the resonant network, i.e., current \( i_t \) and voltage \( v_{Ceq} \), have the pure AC nature with the average component equal to zero. For this reason, the state vector is reduced to:
\[
x_{OL} = \left[ \begin{array}{c} \dot{v}_{C1} \  \\ \dot{i}_t \  \\ \dot{v}_{Ceq} \  \\ \dot{v}_0 \end{array} \right] \tag{27}
\]

with the subvectors
\[
\dot{v}_{C1} = [ {\langle v_{C1} \rangle}_0, \ {\langle i_t \rangle}_1R, \ {\langle i_t \rangle}_1I], \\
\dot{i}_t = \ [ {\langle v_o \rangle}_0, \ {\langle v_{Ceq} \rangle}_1R, \ {\langle v_{Ceq} \rangle}_1I].
\]

Applying the fundamental properties of the Fourier coefficients, as in Sanders et al. (1991), on the state vector (27), the generalized state-space averaging model is:
\[
\dot{v}_{C1} = -\frac{1}{C_1 R_1} v_{C1} + \frac{4}{C_1 \pi} (i_t)_{1R} + \frac{1}{C_1 R_1} v_{\text{in}} \tag{28}
\]
\[
\dot{i}_t = \frac{2}{L_{eq}} (v_{Ceq})_{1R} \frac{\sin(\omega_s t)}{\omega_s} \tag{29}
\]
\[
\dot{v}_{Ceq} = -\frac{2}{L_{eq}} v_{Ceq} - \frac{1}{L_{eq}} v_{Ceq} + \frac{2 \cos(\omega_s t)}{\omega_s} \tag{30}
\]
\[
\dot{v}_0 = \frac{1}{R_0 C_2} (i_t)_{1R} - \frac{4 \sin(\omega_s t)}{\pi C_2} (i_t)_{1R} \tag{31}
\]
\[
\dot{i}_t = \frac{4 \cos(\omega_s t)}{\pi C_2} (i_t)_{1R} \tag{32}
\]

The trigonometric terms in (28)-(33) follow from the application of the rules in Sanders et al. (1991) to the switching functions \( s_1(t) \) and \( s_2(t) \). For the Fourier coefficients of these signals it holds, see Qin and Kimball (2012):
\[
\langle s_1 \rangle_0 = \langle s_2 \rangle_0 = 0 \tag{34}
\]
\[
\langle s_1 \rangle_1R = 0, \quad \langle s_1 \rangle_1I = -\frac{2}{\pi} \tag{35}
\]
\[
\langle s_2 \rangle_1R = -\frac{2 \sin(\omega_s t)}{\pi}, \quad \langle s_2 \rangle_1I = -\frac{4 \cos(\omega_s t)}{\pi} \tag{36}
\]

The state-space model (28)-(33) is a nonlinear model and therefore not appropriate for the voltage reference generator design which relies on the linear system representation. Thus, we linearize the model around the steady-state operating point. Let’s assume that \( \tilde{x}_{OL} = \hat{x}_{OL} - X_0 \), \( \tilde{u}_{OL} = u_{OL} - U_0 \) and \( \tilde{y}_{OL} = y_{OL} - Y_0 \) denote small signal perturbations for the state, control input and output variables around the steady-state vectors \( X_0, U_0 \) and \( Y_0 \), respectively. As control input, we use the switching frequency of the converter in radians, i.e., \( \omega_s = \omega_s \). The output voltage is selected as the controlled output, i.e., \( y_{OL} = v_o \). After linearization around the steady-state, we gain the following open-loop model of the CLLC converter:
\[
\dot{\tilde{x}}_{OL} = A_{OL} \tilde{x}_{OL} + B_{OL} \tilde{u}_{OL} \tag{37}
\]
\[
\tilde{y}_{OL} = C_{OL} \tilde{x}_{OL} + D_{OL} \tilde{u}_{OL} \tag{38}
\]

with the system matrix \( A_{OL} \), the input matrix \( B_{OL} \), the output matrix \( C_{OL} \) and the feed-through matrix \( D_{OL} \):
\[
A_{OL} = \begin{bmatrix}
\frac{1}{C_1 R_1} & 0 & 0 & 0 \\
0 & \frac{4}{C_1 \pi} & 0 & 0 \\
\frac{4}{C_1 \pi} & 0 & -\frac{1}{L_{eq}} & 0 \\
0 & 0 & \frac{2 \sin(\omega_s t)}{\omega_s \pi} & 0
\end{bmatrix} \tag{39}
\]
\[
B_{OL} = \begin{bmatrix}
I_{1R} \\
-I_{1R} \\
\frac{I_{1R}}{V_{Ceq,1R}} \\
\frac{V_{Ceq,1R}}{V_{Ceq,1R}}
\end{bmatrix} , \quad C_{OL} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 \end{bmatrix}, \quad D_{OL} = 0
\]
4. CONVERTER CONTROLLER DESIGN

With the open-loop model of the CLLC resonant converter, we propose the dual loop control concept, as shown in Fig. 4, to regulate the voltage at the low voltage bus of the DC/DC converter. The main idea is to improve the response of the closed loop system when the load power is changing and assure that the output voltage remains at 1 kV for all load conditions. The dual loop control concept leverages a simple, linear algorithm as a primal controller. This controller adjusts the switching frequency \( f_s \), depending on the value of the control error, while the phase shift \( \gamma \) is a function of \( f_s \) and the current load. In the outer loop, a voltage reference generator adjusts the artificial voltage reference \( v_{ref}^A \) of the primal controller to track the actual reference \( v_{o,ref} \). To estimate the AC states, related to \( i_t \) and \( v_{Ceq}, \) we propose to utilize a harmonics estimator (not being subject of this paper) which relies on harmonic regression.

Fig. 4. Control architecture with reference generator and low level PI controller.

4.1 Closed-Loop CLLC Model

To design a voltage reference generator, a closed-loop system model must be known. The output voltage of the CLLC converter is controlled using the PI controller in the following digital realization:

\[
\begin{align*}
 u(k) &= K_P e(k) + \gamma(k) \\
 \gamma(k + 1) &= \gamma(k) + K_I T_s e(k)
\end{align*}
\]

where \( e(k) = v_{ref}^A(k) - v_o(k) \) denotes the control error, \( v_{ref}^A \) the artificial reference value of the output voltage \( v_o \). \( T_s \) is the controller sampling time, while \( K_P > 0 \) and \( K_I > 0 \) are the parameters of the PI controller. The integral term (41) has been discretized using forward-Euler discretization.

To form the closed-loop system model, the continuous state-space representation of the CLLC converter is discretized first assuming zero-order hold discretization such that we gain:

\[
\begin{align*}
 \tilde{x}_{OL}(k+1) &= A_{OL,d} \tilde{x}_{OL}(k) + B_{OL,d} \tilde{u}_{OL}(k) \\
 \tilde{y}_{OL}(k) &= C_{OL,d} \tilde{x}_{OL}(k)
\end{align*}
\]

where \( A_{OL,d}, B_{OL,d} \) and \( C_{OL,d} \) are the discretized counterparts of the corresponding matrices in (39).

By inserting the control law (40)-(41) of the PI controller in the discretized system model (42)-(43) and augmenting the state vector with the additional state \( \gamma \), representing the integrator state of the PI controller, we form a discrete-time state-space model of the closed-loop system:

\[
\begin{align*}
 \begin{bmatrix} \tilde{x}_{CL}(k+1) \\ \gamma(k+1) \end{bmatrix} &= \begin{bmatrix} A_{CL,d} & B_{CL,d} \end{bmatrix} \begin{bmatrix} \tilde{x}_{CL}(k) \\ \gamma(k) \end{bmatrix} + \begin{bmatrix} 0 \\ v_{ref}^A(k) \end{bmatrix} \\
 y_{CL}(k) &= C_{CL,d} x_{CL}(k) + u_{CL}(k)
\end{align*}
\]

with

\[
\begin{align*}
 x_{CL} &= \begin{bmatrix} \tilde{x}_{OL} \\ \gamma \end{bmatrix} \\
 y_{CL} &= \tilde{y}_{OL} \\
 u_{CL} &= v_{ref}^A
\end{align*}
\]

where \( \gamma = \gamma - \gamma_0 \) and \( v_{ref}^A = v_{ref}^A - v_{o,ref} \) are the corresponding small-signal perturbations with respect to the operating point \( \gamma_0 \) and \( v_{o,ref} \). This way, in the closed loop model, the deviation \( v_{ref}^A \) from fixed reference \( v_{o,ref} \) (that is, 1 kV) is the new control input.

Remark 1. The operating point \( (X_0, U_0, Y_0) \) of the closed-loop model refers to its steady state for \( v_{o,ref} = 1 \) kV.

4.2 Reference Generation

Adopting the discrete-time state-space model of the closed-loop resonant converter, we can now design a scheme for the voltage reference generation. This scheme is based on the model predictive control concept in which the state-space model is utilized to predict the evolution of the closed-loop system dynamics over the time horizon of \( N_p \) time steps. The predictions are leveraged to define a finite-time optimal control problem which is solved at every time step. The number of time steps is selected, such that the prediction horizon equals at least the settling time \( \tau_{set} \) of the closed-loop system, i.e., \( \tau_{set} = N_p T_s \), where \( T_s \) denotes the sampling time of the reference generator.

The formulation allows to run the reference generator at a different, lower frequency than the PI controller. In this way, a possibility is opened to further reduce the computational effort. To determine the voltage reference for the PI controller, the algorithm minimizes a quadratic cost function subject to the equality constraints that represent the closed-loop system dynamics, i.e.:

\[
\begin{align*}
 \min_{u_{CL.(i)}} \sum_{i=1}^{N_p} Q \left\| (y_{CL.(k+i|k)} - y_{ref}) \right\|_2^2 \\
 &+ \sum_{i=0}^{N_u-1} R \left\| u_{CL.(k+i|k)} \right\|_2^2 \\
 \text{s.t.} \quad x_{CL.(i|k)} = \tilde{x}_{CL.k} \\
 \end{align*}
\]

In (49), the index \( k + i \mid k \) denotes the prediction of the system output at time step \( k + i \) starting from time step \( k \). \( y_{ref} \) is the actual voltage reference that the system should maintain at the low voltage side (in our case 1000 V). The parameter \( Q > 0 \) represents the weighing of the output deviation at time step \( k + i \), while \( R > 0 \) penalizes the magnitude of the control input. Upon solving the quadratic problem (49) at time \( k \), the optimal sequence \( u^*_{CL.(k)} = [u_{CL.(k|k)}^*, \ldots, u_{CL.(k+N_u-1|k)}^*]^T \) of voltage reference changes over the prediction horizon is gained. Only the first element is added to the steady-state value of \( v_{o,ref} = 1000 \) V.
and applied as a reference for the PI controller in the inner loop, i.e., $v_{A\text{ref},k} = v_{\text{ref}} + u_\text{CL,k}|k$. By substituting the constraints into the cost, Problem (49) can be solved fast and the involved matrices even be pre-computed.

In the formulation of the optimization problem (49), the initial condition of the algorithm appears as an equality constraint. At each sampling time, i.e., at each run, the reference generator is initialized at the current state, $x_{CL,k|k}$. Consequently, the system states at time step $k$ have to be either measured or estimated. The voltage at the input capacitance $C_1$ and the output capacitance $C_2$ (the DC states) are directly measured and thus $\langle v_{C_1}\rangle_0$, $\langle v_0\rangle_0$ are known at time $k$. The real and imaginary components of the current $i_t$ and the voltage $v_{C_{eq}}$ (the AC states), though, have to be estimated as indicated at the beginning of the section. As measurements for the estimation, we assume that the current over the primary side inductance $L_{S1}$ and the voltage across the primary side capacitor $C_{S1}$ are available and perform the harmonic regression to determine $\langle i_t\rangle_{1R}$, $\langle i_t\rangle_{1I}$, $\langle v_{C_{eq}}\rangle_{1R}$ and $\langle v_{C_{eq}}\rangle_{1I}$ at time $k$.

5. SIMULATION RESULTS

5.1 Simulation Setup

The parameters of the considered CLLC resonant converter are listed in Table 1.

| Parameter            | Value          |
|----------------------|----------------|
| Input voltage $V_m$  | 5000 V         |
| Inductance $L_{S1}, L_{S2}$ | 40 $\mu$H     |
| Inductance $L_p$     | 150 $\mu$H     |
| Capacitance $C_{S1}, C_{S2}$ | 7.6 $\mu$F  |
| Capacitance $C_1, C_2$ | 100 $\mu$F    |
| Input resistance $R_1$ | 1 m$\Omega$   |

The maximum load equals $P_{max} = 2.5$ MW ($R_0 = 0.4 \Omega$), while the minimum load is $P_{min} = 500$ kW ($R_0 = 2 \Omega$). To demonstrate the performance of the control concept, we consider the load changing sequence in Fig. 5. The loads are denoted in terms of their resistance. $0.4 \Omega$ corresponds to the largest load, $1 \Omega$ to the medium load and $2 \Omega$ stands for the smallest load. In our context the CLLC converter is a frequency modulated device. The frequency for the frequency modulation is set by the PI controller. The duty cycle for the frequency modulation is set to 50%. The phase shift $\phi$ is changing depending on the load and the switching frequency. In our approach, it is determined according to calculus presented in Jung et al. (2013). For the PI controller and the voltage reference generator, the sampling frequencies are chosen to be equal. These sampling frequencies change based on the load conditions. For the largest load, the sampling frequency is 11 kHz, for the medium load 14.2 kHz, and for the maximum load 20.5 kHz. Since the discretized version of the PI controller is applied, with the change of the sampling frequency of the controller, the integral gain is changed as well (see (41)). This way, we realize a gain-scheduling PI controller, i.e., when the load changes, the parameters of the PI controller are adjusted accordingly and a more reasonable comparison with the dual-loop control concept can be made. The parameters of the reference generator are selected as follows: $N_p = 10$, $N_q = 10$, $Q = 5$, $R = 1$.

5.2 Simulation Results

To evaluate the simulation results, we use the following performance criteria (see Table 2): (i) min./max. LVDC voltage and the related overshoot/undershoot, (ii) root mean square error (RMSE) and (iii) the settling time for a 4% tolerance band around 1 kV. We particularly observe the output voltage at time steps when the load level is changing and compare the system response of standard with the dual-loop controller, designed in this contribution. When the load changes from the maximum to minimum or medium load, we obtain performance metrics which are almost equal for both control concepts (white columns in Table 2). Conversely, when changing from the minimum or medium to the maximum load (gray columns in Table 2), the dual loop concept outperforms the standalone PI controller noticeably.

![Fig. 5. Load sequence and the voltage at the LVDC bus.](image_url)

To further analyze these benefits, we have selected two of these time intervals in Fig. 6 (A: min. to max. load; B: min. to medium load). Besides the output voltage, we illustrate the switching frequency, set by the PI controller, along with the phase shift. For time interval A, the dual loop control scheme is able to reduce the voltage overshoot by 12.7% from 1185.7 V to 1035.1 V, the RMSE by 34.3% and the settling time by 51.1%. The initial undershoot, when switching the load, cannot be avoided as it can not be anticipated by the reference generator. Further, we note that the dual loop controller counteracts the predicted over-/undershoot in voltage by adjusting the reference for the PI controller (green solid line) in the opposite direction. Time interval B enforces this observation, showing a 9.3% reduced overshoot, a 27.6% smaller RMSE and a 53.8% reduced settling time. Similar improvement can be recognized for the other corresponding time intervals. In conclusion, we note that the dual loop control concept outperforms the gain-scheduling PI controller and enable to dynamically adapt to excessive load changes.

6. CONCLUSION

We have proposed a modeling and voltage control concept for a DC/DC resonant converter, applied in marine power systems. Leveraging a generalized state-space averaging method, we have presented a control-oriented model of
Table 2. Overview of performance measures during respective load change intervals.

| Load Change [Ω] | 0.4 → 2.0 | 2.0 → 0.4 | 0.4 → 2.0 | 2.0 → 0.4 | 0.4 → 1.0 | 1.0 → 0.4 | 0.4 → 1.0 | 1.0 → 0.4 |
|-----------------|------------|------------|------------|------------|------------|------------|------------|------------|
| Time interval [s]| 0.02–0.04  | 0.04–0.06  | 0.06–0.08  | 0.08–0.1   | 0.10–0.12  | 0.12–0.14  | 0.14–0.16  | 0.16–0.18  |
| **Ref. gen. + PI** |            |            |            |            |            |            |            |            |
| Vmin [V]        | 953        | 812.1      | 952.4      | 805.7      | 947.7      | 855.3      | 960.5      | 846.1      |
| Vmax [V]        | 1057       | 1035.1     | 1053.7     | 1031.3     | 1078.3     | 1023.7     | 1079.8     | 1022.2     |
| RMSE [V]        | 14.6517    | 41.863     | 16.366     | 37.294     | 18.846     | 30.240     | 18.527     | 31.374     |
| Settling Time [s] (4%) | 0.00021    | 0.00065    | 0.0012     | 0.00064    | 0.00030    | 0.00006    | 0.00002    | 0.00070    |

| **PI** |            |            |            |            |            |            |            |            |
| Vmin [V] | 950.9      | 811.9      | 941.5      | 794.7      | 932.9      | 837.8      | 929.5      | 845.5      |
| Vmax [V] | 1048.1     | 1185.7     | 1034.3     | 1175.1     | 1071.3     | 1129.2     | 1071.5     | 1134.0     |
| RMSE [V] | 14.276     | 10.8       | 11.4       | 38.3       | 10.6       | 38.3       | 10.6       | 38.3       |
| Settling Time [s] (4%) | 0.00025    | 0.00133    | 0.00042    | 0.00122    | 0.00040    | 0.00130    | 0.00050    | 0.00100    |

Fig. 6. Controlled output (voltage), voltage reference from ref. generator, control input (switching frequency) and the feedforward term (phase shift).

The resonant converter in the CLLC topology. A linearized version of the model is used to design a dual loop control scheme to regulate the voltage at the low voltage DC side. The dual loop scheme relies on a primal PI controller, which is complemented by a voltage reference generator. A simulation study demonstrates its ability to improve the dynamic system response and to stabilize the voltage when the load is changing. As part of future work, we will examine how the controlled CLLC converter performs (under disturbances elsewhere in the system) when connected to the remaining vessel power system which involves further power converters and local controllers.

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