Dynamic Modeling of a VSC-HVDC Converter

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Abstract—The aim of this paper is to present a nonlinear dynamic model for Voltage Source Converter-based HVDC (VSC-HVDC) links that can be used for dynamic studies. It includes the main physical elements and is controlled by PI controllers with antiwindup. A linear control model is derived for efficient tuning of the controllers of the nonlinear dynamic model. The nonlinear dynamic model is then tuned according to the performance of an ABB HVDC Light model.

Index Terms—Converters, HVDC transmission, Power system dynamics, Power system modeling, Pulse width modulation converters, Reactive power control, Voltage control, VSC-HVDC,

I. INTRODUCTION

In the previous decade, the infed of Renewable Energy Sources (RES) such as wind and photovoltaics has increased [1]. A further increase is expected; this increase will put additional stress on the already heavily loaded power systems. Power electronic controlled devices such as Flexible AC Transmission Systems (FACTS) and High Voltage DC (HVDC) links can play an important role in controlling and stabilizing the power system. With a higher penetration of RES it is assumed that power oscillations will increase as conventional power plants are decommissioned, leading to a reduction of inertia. Power electronic devices may be able to mitigate this issue, as they have the ability to react on electromechanical transients. One option is to install HVDC links for power system control, for instance, for transient stability control or for power oscillation damping as in [2]. The main purpose of HVDC links still is power transfer over long distances or across seas.

Voltage Source Converter-based HVDC (VSC-HVDC) technology has several advantages over Current Source Converter-based HVDC (CSC-HVDC) technology including independent control of active and reactive power. Further VSC-HVDC can be connected to weak networks, and the direction of power flow can be reversed much faster [3], [4]. This makes VSC-HVDC transmission an attractive solution for connecting more RES to the network, as the controllability is increased and the system operating costs decreased [5].

For dynamic studies it is important to have a suitable VSC-HVDC model. Detailed models where all switching actions are modeled have a high computational effort and are not well-suited for large power systems studies. When reducing the complexity of the model we should be careful not to neglect the physical behavior of the VSC-HVDC link. Dynamic models of VSC-HVDC converters are developed in [6], [7] and [8]. These models differ in complexity and accuracy. However, how the tuning for the controllers of the converters was done has not been addressed.

This paper presents a modified dynamic VSC-HVDC model of [6] and [7] which can be used for dynamic power systems studies. A VSC-HVDC two-level converter is modeled in the dq0 frame representation which is compliant with existing dynamic AC models such as generators and FACTS devices. The VSC-HVDC model includes the dynamics of the physical elements such as the AC phase reactor, the DC converter capacitance and the DC cable dynamics as well as the dynamics of the underlying converter control. Not only is the physical system modeled, also the tuning of the cascaded PI controllers is presented, using a linearized control model. This works very well for the active power. However for the reactive power, the linearized control model has its limit. We should investigate in future work how with nonlinear controller tuning the performance of the controller model can be improved. The model is then compared with the performance of an ABB VSC-HVDC light model [9].

This paper is structured as follows: Section II introduces the physical modeling of the VSC-HVDC link, Section III presents the control structure of the rectifier and inverter and Section IV the tuning of the rectifier and inverter controller with a linearized control model. Section V shows the validation results of the proposed VSC-HVDC model with the ABB HVDC light model and Section VI concludes the paper.

II. VSC-HVDC PHYSICAL MODEL

The variable representation used for this model is a complex phasor representation in the dq0 frame and are denoted by underlined variables, e.g. $\underline{v}_i = v_i^q + jv_i^d$. The dq0 Transformation, or Park’s Transformation, is mainly used in the derivation of models of electrical machines [10]. The dq0 frame rotates with the system frequency and is also known as a rotating reference frame. It is is well suited for electromechanical transient studies and gives a well representation of the converter averaged model [10], [11].

The VSC-HVDC system consists of a physical part as well as controllers for the Voltage Source Converter (VSC) converters. The physical model components are depicted in Fig. 1. The VSC-HVDC link is separated into three main subsystems: Two AC systems, two VSC converters and one DC system. A step down transformer connects the VSC-HVDC link to the AC grid. It converts the AC grid voltage to the needed voltage for the VSC converter. A phase reactor connects the rectifier (VSC 1) with the transformer. The rectifier connects the AC to the DC voltage. A bipolar DC cable transports the power to the inverter (VSC 2). The inverter converts the DC voltage back to AC. The phase reactor filters the high switching harmonics and a step up transformer adjusts the AC voltage from the inverter to the AC grid voltage.

Most VSC based pulsedwidth modulation (PWM) converters have a cascaded control structure, consisting of an inner and an outer control loop [12]. The inner control loop controls the currents and the outer control loop the powers. The outer control loop calculates the current references for the inner...
control loop. The current controllers will thereafter calculate the modulation index needed for the VSC converters to keep the desired power setpoints. Fig. 2 shows the control scheme of the entire VSC-HVDC link. The VSC-HVDC link can be controlled by four external control signals. The rectifier controls the active power flowing over the VSC-HVDC link and the reactive power at the AC terminal. The inverter controls the reactive power at its AC terminal and regulates the DC voltage. The DC voltage controller only consists of a single control loop.

A. AC Model

The AC part of the VSC-HVDC link consists of an ideal step down transformer represented with the inductance \( L_t \) which connects the VSC-HVDC link to the AC grid. A phase reactor, represented by \( X_t \), and \( R_t \) connects the VSC converter with the transformer as depicted in Fig. 1. The q-axis of the rotating dq0 reference frame for each converter is aligned with the AC system voltage \( U_{ac,l} \), hence \( u^d_{l,t} = 0 \). This leads to

\[ u^{q}_{l,t} = j u^{q}_{s,t} \]  

where \( l = 1 \) represents the rectifier and \( l = 2 \) the inverter. We chose that the AC current \( i_d \) is to be directed into the converter. The dynamics of the current are calculated as follows:

\[ \frac{X_t}{\omega} \frac{d}{dt} X_{i,t}^{q} = -R_t i^{q}_{l,t} - u^{d}_{c,t} + u^{q}_{l,t} \]  

\[ \frac{X_t}{\omega} \frac{d}{dt} X_{i,t}^{d} + X_{i,t}^{d} = -R_t i^{d}_{l,t} - u^{q}_{c,t} + u^{d}_{l,t} \]  

where \( u^{d}_{l,t} \) and \( u^{q}_{l,t} \) represent the voltage at the converters and \( u^{d}_{c,t} \) and \( u^{q}_{c,t} \), the voltage after the step down transformer. We neglected the current dynamics of the transformer because for all other transformers in the AC grid the dynamics are not modelled. The voltage equations are

\[ u^{d}_{l,t} = X_{i,t}^{d} \]  

\[ u^{q}_{l,t} = X_{i,t}^{q} \]  

We solved (4) and (5) for \( u^{d}_{l,t} \) and \( u^{q}_{l,t} \) respectively and substituted into (2) and (3) will give us the dynamic AC equations

\[ \frac{X_t}{\omega} \frac{d}{dt} X_{i,t}^{q} = (X_t + X_t) i^{q}_{l,t} - R_t i^{d}_{l,t} - u^{d}_{c,t} \]  

\[ \frac{X_t}{\omega} \frac{d}{dt} X_{i,t}^{d} = - (X_t + X_t) i^{d}_{l,t} - R_t i^{q}_{l,t} - u^{q}_{c,t} + u^{d}_{l,t} \]  

B. DC Model

The DC circuit of the VSC-HVDC link consists of the converter capacitance \( C_{converter} \) and a bipolar DC cable. The cable is modeled as a \( \pi \)-equivalent with the elements \( R_{dc}, L_{dc} \) and \( C_{cable} \). The equivalent DC capacitance \( C_{dc} \), as depicted in Fig. 1, we calculate as

\[ C_{dc} = C_{converter} + \frac{1}{2} C_{cable} \]  

The dynamic circuit equations of the DC side are

\[ C_{dc} \frac{d}{dt} I_{dc,1} = I_{dc,1} - I_{cable} \]  

\[ C_{dc} \frac{d}{dt} I_{dc,2} = I_{dc,2} + I_{cable} \]  

\[ L_{dc} \frac{d}{dt} U_{dc,1} = U_{dc,1} - I_{dc,2} - R_{dc} I_{cable} \]  

C. Converter Model

The VSC converter used for this model is assumed to be a lossless two level PWM converter. It is modelled as an averaged switched converter model. The higher harmonics of the switchings of the Insulated Gate Bipolar Transistors (IGBTs) are neglected. In this paper, the 0 component is neglected and thus only symmetric AC systems can be simulated [11]. This is consistent as the AC grid is as well represented in the dq reference frame. The relationship between the AC and DC side is given as

\[ u_{s,t} = K_0 M_{l} \cdot U_{dc,1} \]  

where \( M_{l} = M_{l}^d + j M_{l}^q \) represents the modulation index of the converter and is the control signal to control the setpoints of the converter. \( K_0 \) is the modulation constant and in case of rectangular modulation is \( K_0 = \frac{\sqrt{3}}{2} \) [13]. We assumed that the converters are lossless. Thus the power balance of the AC and the DC side has to be fulfilled.

\[ u^{d}_{c,t} + u^{q}_{c,t} = P_{ac,t} = P_{dc,t} = 2 \cdot I_{dc,1} U_{dc,1} \]
Each converter has two control inputs. The rectifier is controlled by $M_{d1}^d$ and $M_{q1}^d$ and the inverter by $M_{d2}^q$ and $I_{dc,2}$. Unlike in [7], there is no q-current controller at the inverter side, as shown in Fig. 2. This would overdetermine the system. Therefore the q-part of current $I_{dc,2}$ is calculated using the power equality (13).

III. VSC-HVDC CONTROLLER STRUCTURE

The goal of the converter control is to control the active power $P_1$ at the AC grid bus at the rectifier side and the reactive powers $Q_1$ at each AC grid bus. Fig. 2 shows the detailed control structure of the VSC-HVDC link. The power expressions in d-q frame,

$$P_1 = u_{d1}^q \cdot i_{q1}^d$$,
$$Q_1 = u_{d1}^q \cdot i_{q1}^d$$,

show that $P_1$ and $Q_1$ can be controlled independently by setting the associate currents $i_{d1}^q$ and $i_{q1}^d$. This will provide the references for the current controllers, the inner control loop.

A. Current Controller

The inner and faster control loop controls the currents. It receives the setpoints form the slower outer control loop as seen in Fig. 2. It is seen in (6) and (7) that the currents are cross-coupled because of the dq transformation. Feed-forward decoupling terms have to be introduced to decouple the currents.

$$\Delta u_d^d = i_{q1}^d (X_r + X_l)$$,
$$\Delta u_d^q = i_{q1}^d (X_r + X_l)$$

This leads to two independent current controllers. Fig. 3 shows the scheme of the d-current and the q-current controllers. For each controller, a PI controller, as depicted in Fig. 4, was used. It includes saturation and anti windup. The parameters for the the d-current controller are $K_p = K_{p,dc}^d$ for the gain and $K_i = K_{i,dc}^d$ for the integral term. The q-current controller the parameters $K_p = K_{p,dc}^q$ give the gain and $K_i = K_{i,dc}^q$ the integral term. $K_A$ is the gain of the integrator anti windup circuit. The PI controller calculates the limited internal voltage $u_{lim,1}$. With the feed forward system voltage $u_{lim}^d$, the feed forward decoupling term $\Delta u_d^d$ and the feed forwarded DC voltage $U_{dc,1}$, the modulation indices $M_{d1}^d$ and $M_{q1}^d$ are calculated and sent to the converter model. The modulation indices are constrained by their physical limits

$$M_{d1}^d \leq M_{d1}^q \leq M_{d1}^\text{max}$$,
$$M_{q1}^d \leq M_{q1}^q \leq M_{q1}^\text{max}$$

B. Active & Reactive Power Controller

The outer control loop of the VSC converter controls the active and reactive power. The PI controller used for the active power controller has the same structure as the PI controller depicted in Fig. 4. The parameter for the gain is $K_p = K_{p,dc}^d$ and for the integral term is $K_i = K_{i,dc}^d$. The PI controller calculates the reference current $i_{ref,1}^d$ needed for the desired active power. The current reference is limited to,

$$i_{min}^d \leq i_{ref,1}^d \leq i_{max}^d$$

C. DC Voltage Controller

The fourth control loop is controlling the DC voltage $U_{dc,2}$ at the inverter side. The DC voltage controller consists of a single PI controller of the same structure as shown in Fig. 4. It is a saturated PI controller with an integrator anti windup circuit. The parameter for the gain is $K_p = K_{p,dc}$ and for the integral term $K_i = K_{i,dc}$. The PI controller calculates the DC reference current $I_{dc,2}$ to keep the DC voltage at its desired value. The DC reference is also limited

$$I_{dc,\text{min}} \leq I_{dc,2} \leq I_{dc,\text{max}}$$

This ensures that the converter will operate within its limits.

IV. LINEARIZED CONTROLLER TUNING

This section will describe how the controllers were tuned using a linear control model. However with the linearization we loose the nonlinearity of the saturation of the controllers. As we will see in Section V this works very well for the active power control loop but not for the active power control loop. For each control loop, we derived a linearized transfer function and we calculated the controller parameters using MATLAB.
A. Linearized Plant

The AC model and the converter model from Section II-A and II-C are linearized at the current operating point \( x_0 \). This results in following combined linearized AC plant

\[
\begin{bmatrix}
\frac{X_{ld}}{\omega} \\
\frac{X_{ld}}{\omega}
\end{bmatrix} = \begin{bmatrix}
-R_t & X_t + X_i \\
-(X_t + X_i) & -R_t
\end{bmatrix} \begin{bmatrix}
\hat{i}_d^p \\
\hat{i}_d^q
\end{bmatrix} + \begin{bmatrix}
-2K_0U_{dc,1,0} & 0 \\
0 & -2K_0U_{dc,0,1}
\end{bmatrix} \begin{bmatrix}
\dot{M}_d^q \\
\dot{M}_d^d
\end{bmatrix}
\]

\[
+ \begin{bmatrix}
0 & -2K_0M_{ld,0,1} \\
1 & -2K_0M_{ld,0,1}
\end{bmatrix} \begin{bmatrix}
\hat{u}_{sl,1} \\
\hat{u}_{s,1}
\end{bmatrix}
\] (23)

where \( \hat{i}_d^p, \hat{i}_d^q, \dot{M}_d^q, \dot{M}_d^d, \hat{u}_{s,1} \text{ and } \hat{u}_{sl,1} \) denote the linearized values in the form of \( \hat{x} = x - x_0 \). This nomenclature will be used from now on, where \( x \) is a placeholder for \( x_0 \) the steady state and \( \hat{x} \) the deviation from the steady state. This leads to the transfer function for the d-current

\[
G_{id,l}(s) = \frac{\hat{i}_d^p(s)}{M_d^d(s)} = -2K_0U_{dc,1}(2K_0M_{ld,1,0}^d \dot{U}_{dc,1}(s) + (X_t + X_i)\hat{i}_d^q(s) + \frac{1}{\omega} s + R_t)
\] (24)

and for the q-current

\[
G_{iq,l}(s) = \frac{\hat{i}_d^q(s)}{M_d^q(s)} = -2K_0U_{dc,1}(2K_0M_{ld,1,0}^q \dot{U}_{dc,1}(s) - (X_t + X_i)\hat{i}_d^q(s) + \hat{u}_{s,1}^d + \frac{1}{\omega} s + R_t)
\] (25)

The DC model of Section II-B is linearized at its operating point which leads to

\[
\begin{bmatrix}
C_d \dot{U}_{dc,1} \\
C_d \dot{U}_{dc,2} \\
L_{dc} \dot{I}_{cable}
\end{bmatrix} = \begin{bmatrix}
\frac{P_{ld,0}}{U_{dc,0}} & 0 & -1 \\
0 & 1 & 0 \\
1 & -1 & -R_{dc}
\end{bmatrix} \begin{bmatrix}
\dot{U}_{dc,1} \\
\dot{U}_{dc,2} \\
\dot{I}_{cable}
\end{bmatrix}
\]

\[
+ \begin{bmatrix}
0 & 1 \\
1 & 0 \\
0 & 0
\end{bmatrix} \begin{bmatrix}
\dot{U}_{dc,0} \\
\dot{U}_{dc,0} \\
\dot{U}_{dc,0}
\end{bmatrix}
\] (26)

The transfer function for the DC voltage is derived to

\[
G_{dc}(s) = \frac{\dot{U}_{dc,2}(s)}{\dot{I}_{dc,2}(s)} = \frac{1}{sC_{dc} - 1}
\] (27)

B. Linearized Current Controller Tuning

In the following section the current controller from Section III-A will be linearized. However the saturation and the antiwindup circuit of the PI controllers and the limits on the modulation indices are neglected in the linearization. The linear transfer function for the d-current controller is

\[
K_{id,l}(s) = \frac{\hat{i}_{ref,1}^d(s)}{M_d^d(s)} = \frac{1}{2K_0U_{dc,1}}(2K_0M_{ld,1}^d \dot{U}_{dc,1}(s) - (X_t + X_i)\hat{i}_d^q(s) + \frac{K_{p,ld,1}s + K_{i,ld,1}}{s})
\] (28)

and for the q-current controller is

\[
K_{iq,l}(s) = \frac{\hat{i}_{ref,1}^q(s)}{M_d^q(s)} = \frac{1}{2K_0U_{dc,1}}(2K_0M_{ld,1}^q \dot{U}_{dc,1}(s) - (X_t + X_i)\hat{i}_d^q(s) - (X_t + X_i)\hat{i}_d^q(s) + \frac{K_{p,q,1}s + K_{i,q,1}}{s})
\] (29)

Therefore we derived the open loop transfer function of the inner control loop, as depicted in Fig. 2. For the d-axis we receive

\[
l_{id,l}(s) = K_{id,l}(s) \cdot G_{id,l}(s) = \frac{K_{p,ld,1}(s + \frac{K_{i,ld,1}}{K_{p,ld,1}})}{s + \frac{K_{p,ld,1}}{K_{i,ld,1}}} \cdot \frac{s \omega}{s + \frac{\omega}{R_t}}
\] (30)

By choosing the negative zero of the controller

\[
-\frac{K_{i,ld,1}}{K_{p,ld,1}} = -\frac{R_t \omega}{X_t}
\]

the negative pole of the reactor is cancelled. This leads to the closed loop transfer function

\[
H_{id,l}(s) = \frac{\hat{i}_{ref,1}^d(s)}{i_{ref,1}^d(s)} = \frac{1}{s + \frac{K_{p,ld,1}}{K_{i,ld,1}}} + \frac{1}{\frac{R_t}{\tau_{id}} + 1}
\] (31)

The response from \( i_{ref,1}^d(s) \) to \( i_{ref,1}^d(s) \) is based on a first-order transfer function whose rise time is given by \( T_{id}, \tau_{id} \) is a design parameter and can be chosen freely. It should be sufficiently small for a fast current controller. Therefore the parameters for the PI controller are

\[
K_{p,ld,1} = \frac{X_t}{\omega}, \quad K_{i,ld,1} = \frac{R_t}{\tau_{id}}
\] (32)

For the q-current controller tuning, the same design procedure is applied.

C. Linearized Active and Reactive Power Controller

The transfer function of the active and reactive power controller is derived in this section. The active power controller from Section III-B with the plant is linearized to

\[
k_{p,l}(s) = \frac{K_{p,pl,1}s + K_{i,pl,1}}{s}
\] (33)

The saturation and the antiwindup circuit are neglected in the linearization. The linear plant model for the response of \( i_{ref,1}^q \) to \( P_1 \) is

\[
G_{p,1} = \frac{\hat{P}_1(s)}{i_{ref,1}^q(s)} = \frac{1}{s \tau_{iq} + 1} + \hat{u}_{s,1,0}^q \hat{u}_{s,1,0}^q(s)
\] (34)
For tuning purposes, we assume that the disturbance \( \hat{u}_{s,1}^q(s) \) is small and can be neglected. This leads to following open loop transfer function

\[
l_{p,1}(s) = k_{p,1}(s) \cdot G_{p,1}(s) = \frac{u_{s,1,0}(K_{p,1} + K_{l,1})}{s(s\tau_q + 1)} .
\]

which brings us to the closed loop transfer function

\[
H_{p,1} = \frac{\hat{P}_1(s)}{\hat{P}_{ref}(s)} = \frac{1}{s^2 + \frac{u_{s,1,0}K_{p,1}}{\tau_q} + 1} \cdot \frac{s}{\frac{s\tau_q + 1}{\frac{s\tau_q}{\tau_q}}} .
\]

The active power response follows a second-order transfer function. It can be written in standard form

\[
H_{p,1} = \left(\frac{\alpha}{\omega_n} + 1\right) \cdot \left(\frac{s}{\omega_n} + 2\zeta\left(\frac{s}{\omega_n}\right) + 1\right) ,
\]

where \( \zeta \) is the damping ratio, \( \omega_n \) the undamped natural frequency and \( \alpha \) is a scaling factor of the zero [14]. A coefficient comparison gives

\[
\alpha = \frac{2K_{l,1}\tau_q}{K_{p,1}} , \quad \omega_n = \frac{u_{s,1,0}K_{p,1}}{\tau_q} ,
\]

\[
2\zeta\omega_n = \frac{u_{s,1,0}K_{p,1}}{\tau_q} .
\]

Standard quantities like the peak time \( t_p \) or the overshoot \( M_p \) express the transfer function and can be defined. Ref. [14] gives a formula for both quantities in relation to \( \zeta \) and \( \omega_n \). They can be solved for \( \zeta \) and \( \omega_n \), leading to:

\[
\zeta = \frac{t_p}{\sqrt{\ln\left(M_p^2 + \pi^2\right)}}, \quad \omega_n = \frac{\pi}{\sqrt{1 - \zeta^2}}
\]

and substituted into (37) to solve for the PI-parameters:

\[
K_{p,1} = \frac{2\omega_n\tau_q - 1}{u_{s,1,0}} , \quad K_{l,1} = \frac{\omega_n^2\tau_q}{u_{s,1,0}} .
\]

If \( \alpha \) is large, the zero will be far removed from the poles and thus the zero will have little effect on the response. The effect of the zero is to increase the overshoot \( M_p \). Ref. [14] shows that the zero has little effect if \( \alpha \geq 3 \). If the desired design parameters fulfill this constraint then \( K_{p,1} \) and \( K_{l,1} \) can be chosen according to (41). Otherwise a numeric pole zero placement has to be performed to achieve the design requirements.

For the reactive power controller, the same design procedure can be used with the assumption that the disturbance \( \hat{u}_{s,1}^q(s) \) is zero.

\[\text{D. Linearized DC Voltage Controller}\]

The fourth control loop is comprised of the DC voltage controller. The open-loop transfer function with the controller from Section III-C and the DC plant (27) is

\[
l_{dc}(s) = k_{dc}(s) \cdot G_{dc}(s) = \frac{K_{p,dc}s + K_{i,dc}}{s(sC_{dc} - 1)} .
\]
are substantial. It could not be tuned with the analytical difference between the linearized and the full dynamic model behavior.

The linearized power step has a second-order behavior. This allows it to model was developed. It is seen in the results, that the active control model (blue dotted curve) does not have the same fast. It takes about 100 ms to reach the new steady state. This leads to a response of the reactive power controller which is similar to the one of the Power Factory Model. The parameters for the PI controllers were tuned heuristically. This method allows us to tune a VSC-HVDC link to a desired step response. This is not only valid for two level PWM converters. The switching harmonics are neglected and, with the appropriate parameters, the behavior of a modular multilevel converter can be reproduced.

Fig. 5. Active power step response of different VSC-HVDC models. The red solid curve is the response of the ABB HVDC Light model in power factory. The green solid curve is the response of the introduced nonlinear VSC-HVDC model and the blue dotted line is the response of the linearized control model.

Fig. 6. Reactive power step response of different VSC-HVDC models. The red solid curve is the response of the ABB HVDC Light model in power factory, the green solid curve is the response of the introduced nonlinear VSC-HVDC model and the blue dotted line is the response of the linearized control model.

Factory model does not have a second-order response (red solid curve), as is seen in Fig. 6. The response time is very fast. It takes about 100 ms to reach the new steady state. This results in limiters and saturations being activated. The linear control model (blue dotted curve) does not have the same response as the nonlinear dynamic model (green solid curve). The parameters for the PI controllers were tuned heuristically. However, for the reactive power this is not true. The difference between the linearized and the full dynamic model is substantial. It could not be tuned with the analytical method. However the analytical method gave a starting point for the heuristic method. The mismatch is explained by the fast response, because the limits of the controllers reach saturation. For a more efficient tuning method, a nonlinear controller tuning should be applied. This would assure us that the stability of the converter is guaranteed.

This method allows us to tune a VSC-HVDC link to a desired step response. This is not only valid for two level PWM converters. The switching harmonics are neglected and, with the appropriate parameters, the behavior of a modular multilevel converter can be reproduced.

ACKNOWLEDGMENT

We acknowledge the support of swisselectric research and ABB for the project Power System Performance Enhancement by Use of Voltage Source Converter Based HVDC in the ENTSO-E RG Continental Europe System.

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