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CAPACITOR VOLTAGE BALANCING IN MODULAR MULTILEVEL CONVERTERS

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Abstract

A universal capacitor voltage control method for converters built from series connected modules is presented. It fully exploits both the circulating currents and the common-mode voltage without affecting the phase current control. The controllability of the capacitor voltages in various such converters is investigated. It is found that the nonzero branch currents and terminal voltages are necessary for capacitor voltage balancing.

1 Introduction

A number of power converter structures built from series connections (branches) of identical two pole power electronic building blocks (PEBBs; see Figure 1) are known: AC-DC converters with two or three phases (Figure 2 a) [1], three phase AC to two phase AC converters (also a) [2][3], STATCOMs in delta (b) and wye (c) configuration [4] and three phase AC-AC converters resembling full- or sparse matrix converters (d) (e) [5][6]. They share the attraction of comparatively simple voltage scaling by means of varying the number of PEBBs in each branch. The number of voltage levels and the apparent switching frequency is proportional to the number of PEBBs. The many voltage levels lead to a very low THD and small voltage steps that render filters or reinforced machine insulation dispensable. On the other hand, a relatively large amount of energy has to be stored in the PEBB capacitors as the PEBBs are essentially single phase converters.

Each PEBB contains a capacitor, IGBTs and the associated gate drive-, communication-, control- and protection electronics. The current commutation loops are encapsulated in the PEBBs and the branch currents are continuous in MMCs, which distinguishes them from conventional voltage source and matrix converters. Optimisation for low leakage inductance is not necessary beyond the PEBBs which facilitates modular designs.

The voltages in all those capacitors must be controlled. They can diverge due to control imperfections, transients and component variations. The PEBB capacitor voltage control problem has three parts: equalisation of the capacitor voltages within branches, equalisation of the capacitor voltages among branches and control of the overall energy stored in the converter. This paper deals with the equalisation among branches.

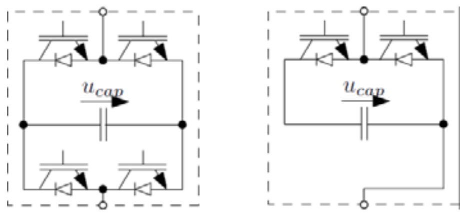

Figure 1: Typical power circuits of PEBBs used in converters built from series connected modules. Depending on the application bipolar PEBBs with three output voltage levels ($u_{cap}, 0, -u_{cap}$; shown left) or unipolar PEBBs with two output voltage levels ($u_{cap}, 0$) are preferable.

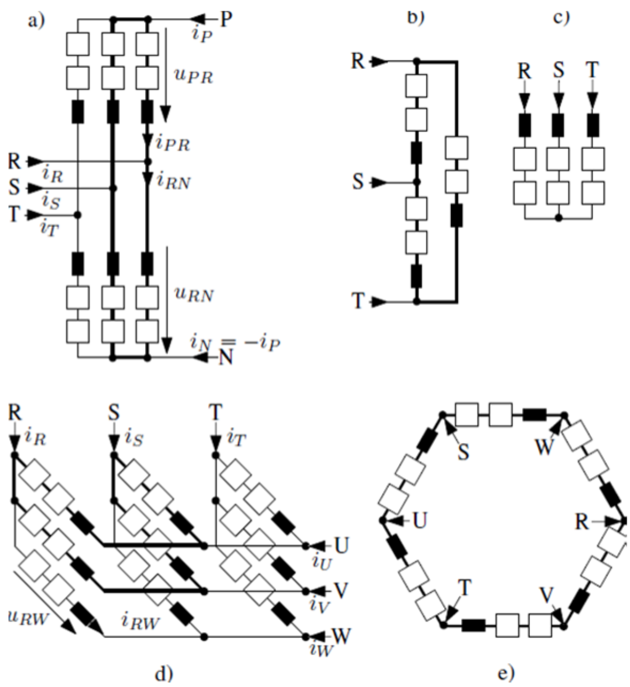

Figure 2: Known converter topologies built from series connected modules. The boxes are either type of PEBB shown in Figure 1, the number of series connected PEBBs is variable. Examples for loops within the converters are indicated by bold lines.

2 Capacitor voltage balancing within branches

PEBBS in the same branch are series connected and will thus always be exposed to the same current. Balancing within branches must thus be accomplished by altering the output voltages of the PEBBs, ideally without altering the output voltage of the branch. Marquardt proposed to select an appropriate cell for each switching event as it is triggered by a central modulator such that the capacitor balancing within branches is achieved [1]. The central generation of all switching commands poses demanding requirements on the communication links to the many switches. These requirements are relaxed if the modulators are implemented as close as possible to the switches. The voltage balancing algorithm proposed by Marquardt cannot be applied if the switching is triggered by distributed modulators however. With distributed modulators capacitor voltage balancing within branches is achieved by individually adjusting the voltage references of the PEBBs [7]. Another alternative for capacitor voltage balancing within branches is to use phase shifts on the switching patterns of individual PEBBs [4].

Changes of the PEBB output voltages can only have an effect on the capacitor voltages if current is flowing through the PEBB. When there are no phase currents, circulating currents in the loops of branches within the topologies must be introduced for this purpose.

3 Capacitor voltage balancing among branches

The voltage references across branches are determined by the DC link and phase voltage references and the terminal current controllers [5][6][12]. Only the common-mode or the neutral point voltage may be altered for capacitor voltage balancing [4][8].

The loops found in most topologies permit the introduction of circulating currents which flow within the converter without passing through the load or grid (examples for such loops are indicated by bold lines in Figure 2). Circulating currents will occur without further control action if the capacitor voltage ripple is not compensated for in the modulation and the branch currents are not controlled. Converter transient performance can be improved by actively controlling the circulating currents to perform branch balancing [7][9][10][11].

If capacitor voltage balancing among branches is achieved exclusively by means of circulating currents and common-mode voltages adverse effects on the phase and DC current controllers can be avoided.

3.1 Topology independent feed-back law

A capacitor voltage balancing controller that can be applied to all the topologies shown in Figure 2 is derived subsequently.

The average capacitor voltages in a branch correlate with the sum of the stored energies in the capacitors e_b. The derivative of the branch energy is the branch power p_b, which in turn is the product of the branch current i_b and the branch voltage u_b.

$$\dot{e}_b = p_b = i_b u_b \tag{1}$$

If the stored energy in one branch is lower than in the others, it must be charged. This can be accomplished with a simple proportional feedback law from branch energy deviation Δe_b to branch power deviation Δp_b:

$$\Delta p_b = -\mathrm{K}\Delta \mathrm{e_b} \tag{2}$$

The branch power itself is not a manipulated variable, but it can be adjusted via branch current deviations Δi_b or branch voltage deviations Δu_b.

$$\dot{e}_b = \mathrm{p_b} + \Delta p_b = (\mathrm{i_b} + \Delta \mathrm{i_b})(\mathrm{u_b} + \Delta \mathrm{u_b}) \tag{3}$$

$$\Delta \dot{e}_b = \Delta p_b = \Delta \mathrm{i_b} u_b + i_b \Delta u_b + \Delta \mathrm{i_b} \Delta u_b \tag{4}$$

Subsequently it will be assumed that $\Delta \mathrm{i_b} \ll i_b$ and $\Delta u_b \ll u_b$ and therefore equation (4) can be linearized to

$$\Delta \dot{e}_b = \Delta p_b \approx \Delta \mathrm{i_b} u_b + i_b \Delta u_b \tag{5}$$

A feedback of the energy shall be introduced. Feedback linearization can be achieved with

$$\Delta \mathrm{i_b} = \frac{-K_i \Delta e_b}{u_b}, \qquad \Delta \mathrm{u_b} = \frac{-K_u \Delta e_b}{i_b} \tag{6}$$

This feedback linearization will lead to infinite control action if the denominator is zero. Since the branch currents are coupled the infinite $\Delta \mathrm{i_b}$ of a branch with $u_b = 0$ will upset the capacitor voltage of the other branches which will in general have a $u_b \neq 0$. This is avoided by weighing the control action of equations (6) with the square of the denominator, resulting in:

$$\Delta \mathrm{i_b} = -K_i \Delta e_b u_b, \qquad \Delta \mathrm{u_b} = -K_u \Delta e_b i_b \tag{7}$$

The branch current or voltage deviations calculated so far can alter the phase voltages and currents when they are applied to the references. Thus the vector of branch current and voltage deviations shall be mapped into a subspace of the branch currents or voltage respectively which is orthogonal to the currents and voltages that should not be altered. These constraints are formulated in matrix form:

$$\Delta \vec{\imath}_{terminals} = T_i \Delta \vec{\imath}_b = \vec{0} \tag{8}$$

$$\Delta \vec{u}_{line-line} = T_u \Delta \vec{u}_b = \vec{0} \tag{9}$$

T must not be of full rank (i.e. the branch currents must not be fully determined by the terminal currents) in order to have any degrees of freedom left for control. The kernel of a matrix is the subspace that is mapped to the zero vector by the matrix. Deviation vectors that are in the kernel of the T matrix will thus not violate the constraint (8) and (9) respectively. An arbitrary vector can be orthogonally projected into the kernel using an orthonormal basis of the kernel B:

$$\vec{y} = B^T B \vec{x} \tag{10}$$

The orthogonality of the projection implies that the difference between the original vector $\vec{x}$ and the projected deviation vector $\vec{y}$ will have the smallest 2-norm of all vectors that satisfy the constraints (i.e. it is optimal in the least-squares sense). Such a basis can be obtained from the singular value decomposition of the constraint matrix:

$$\mathrm{T} = \mathrm{USV}^{\mathrm{T}} \tag{11}$$

Where U and V are unitary matrices and S is diagonal. The diagonal elements of S are the singular values σ. The column vectors of V with a singular value of zero are mapped to the zero vector by T. Together these column vectors $V_{\sigma=0}$ are an orthonormal basis of the kernel of T. The deviation vectors (7) are projected into the kernel of T using (10) ($B = V_{\sigma=0}$) and thereby decoupled from the phase current control:

$$\Delta\vec{\imath}_b = -D_i \mathrm{K_i}(\Delta\vec{e}_b \circ \vec{u}_b), \qquad D_i = V_{\sigma=0}V_{\sigma=0}^{\mathrm{T}}(T_i) \tag{12}$$

$$\Delta\vec{u}_b = -D_u \mathrm{K_u}(\Delta\vec{e}_b \circ \vec{\imath}_b), \qquad D_u = V_{\sigma=0}V_{\sigma=0}^{\mathrm{T}}(T_u) \tag{13}$$

where $\circ$ is the entry-wise product. The decoupling matrices depend only on the constraints and can thus be precomputed offline. Functions to numerically calculate singular value decompositions are provided by common mathematics software tools.

The decoupling matrices are an elegant and universal solution that is well suited for analysis. They are not the most efficient approach to implement the balancing control for a specific topology however. For instance the matrix D_u can be replaced by a mean in most cases since the common-mode voltage u_{cm} will be the only degree of freedom of the branch voltages left for $\Delta\vec{u}_b$:

$$\mathrm{u_{cm}} = \frac{1}{N}\sum_i s_i u_{b,i} \tag{14}$$

s_i is either 1 or -1 according to the topology and the count arrows.

Similarly, more efficient equivalent implementations of D_i can be found for particular topologies.

3.2 Feasibility of capacitor voltage balancing

It is of interest to know under which conditions regarding the terminal voltage and current references the branches can be fully balanced. A necessary and sufficient condition for this is asymptotic Lyapunov stability of the system given by equations (3), (12) and (13):

$$\dot{\vec{\mathrm{e}}}_{\mathrm{b}} = \left(\vec{\imath}_b - D_i\mathrm{K_i}(\vec{\mathrm{u}}_{\mathrm{b}} \circ \vec{e}_b)\right) \circ \left(\vec{\mathrm{u}}_{\mathrm{b}} - D_u\mathrm{K_u}(\vec{\mathrm{\imath}}_{\mathrm{b}} \circ \vec{e}_b)\right) \tag{15}$$

The system $\dot{\vec{\mathrm{e}}}_{\mathrm{b}} = \vec{\mathrm{u}}_{\mathrm{b}} \circ \vec{\mathrm{\imath}}_{\mathrm{b}}$ must be stable before introducing Δu_b and Δi_b. The derivation of stable operating states is the subject of references [5][6] and [12].

The branch voltage references $\vec{\mathrm{u}}_{\mathrm{b}}$ are calculated from the terminal voltages. The terminal voltages consist of a positive sequence voltage (U_1 for port 1, U_2 for port 2) and a common mode voltage between the two ports U_{cm}. Due to symmetry it is irrelevant which is port one and which is port two.

Circulating currents must be introduced for voltage balancing within branches when the phase currents are zero as noted in section 2. Circulating currents with distinct frequencies are introduced for each degree of freedom in the branch currents that is left by the constraint matrix T_i in order to gain the maximum degrees of freedom for branch balancing:

$$\vec{\imath}_b = I_c V_{\sigma=0}\begin{pmatrix}\sin\omega_{c1}t\\ \sin\omega_{c2}t\\ \vdots\end{pmatrix} \tag{16}$$

Instead of formally proving Lyapunov asymptotic stability for the system it was found to be more efficient to numerically integrate the equations from random initial conditions and check for convergence. An exemplary solution is shown in Figure 3.

Table 1 shows in which operating points the topologies of Figure 2 can be balanced with circulating currents and common mode voltage alone, that is, without altering the terminal voltages and currents. It can be seen that for all topologies balancing only works when a nonzero voltage is applied at all terminals. The wye STATCOM (c) cannot be balanced at all without terminal currents for the obvious reasons that it has no loops for circulating currents.

	$U_{cm} = 0$	$U_{cm} \neq 0$
$U_1 = 0, U_2 = 0$		
$U_1 \neq 0, U_2 = 0$	b	
$U_1 = 0, U_2 \neq 0$	*d, e*	*a, d, e*
$U_1 \neq 0, U_2 \neq 0.\ f_1 \neq f_2$	a, d, e	a, d, e
$U_1 \neq 0, U_2 \neq 0, f_1 = f_2$	a, d	a, d

Table 1: Controllability of the branch energies in the topologies shown in Figure 2 using common-mode voltage and circulating currents for different operating conditions distinguished by the voltage magnitudes (U_1, U_2) and frequencies (f_1, f_2) at the two ports of the topologies and the magnitude U_{cm} and frequency f_{cm} of the common mode voltage between the two ports. In all cases $f_{cm} \neq f_1$, $f_{cm} \neq f_2, I_1 = I_2 = 0$ and circulating currents at frequencies distinct from those of all other currents and voltages are injected. Cases that are controllable without altering phase currents or voltages are indicated with the letters of the topology in Figure 2. Cases that become controllable when the terminal currents on port 2 are also altered for the purpose of branch energy balancing are indicated in *italics*. For STATCOMs only the cases with $U_2 = 0$ are considered as they do not have a second port. For the other topologies it does not matter which is port one and which is port two.

In motor drive applications the converter may be operated with zero motor voltage and current for extended periods of time. Thus balancing must also be possible in the case $U_1 =$

0, $U_2 \neq 0$ where the motor is connected to port one. For this case the results in Table 1 show that the converters cannot be balanced unless the constraints (8) are relaxed to admit variations in the currents on port two of the converter for the purpose of branch energy balancing. These adjustments are different from the ones needed to control the overall energy stored in the converter as they cannot be restricted to cause only low frequency symmetric positive sequence current deviations. To this end the decoupling matrix D_{i1} is calculated with a constraint matrix that does not put constraints on the terminal currents of port two. A weighted sum D_{ic} of D_i and D_{i1} allows using a smaller gain for balancing with terminal currents than for balancing with circulating currents in order to reduce the impact on the terminal currents. The scaling preserves the unity spectral radius of D_{ic} which is relevant for the controller gain calculation shown in the next section.

$$\mathrm{D_{ic}} = \frac{D_i + \kappa D_{i1}}{1+\kappa}, \qquad 0 < \kappa \tag{17}$$

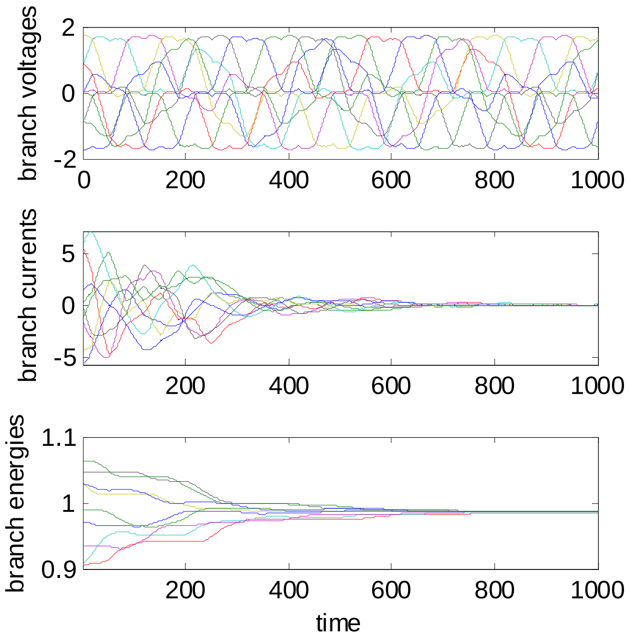

Figure 3: Results of numerical integration of equation (15) for topology d of Figure 2. The convergence of the branch energies is evident. The units are arbitrary.

3.3 Controller gains

Values for the controller gains K_i and K_u which will lead to stable system behaviour can be determined using a robust design approach, i.e. by designing the controller for a worst-case (i.e. maximum gain) plant derived from equation (15) using the linearization shown in equation (5). The maximum gain of D_i and D_u corresponds to their spectral radius and is unity because V is a unitary matrix. The simplified worst case plant is thus:

$$\dot{e}_b = -\left(K_i U_{b,max}^2 + K_u I_{b,max}^2\right) e_b \tag{18}$$

This system is stable for any positive gains K_i and K_u. In practice the gains are limited by the overall dead time of the control loop T_d. According to the bode stability criterion the open loop gain has to be smaller than unity for frequencies where the phase is equal to uneven multiples of π. The integral type plant at hand has a phase of $-\pi/2$. The dead time introduces a phase of $-T_d\omega$. With a targeted phase margin of $\pi/4$ at the crossover frequency ω_c the bode stability criterion becomes

$$\frac{\pi}{2} + \mathrm{T}_d \omega_c + \frac{\pi}{4} \leq \pi \tag{19}$$

The crossover frequency of the control loop must thus satisfy:

$$\omega_c \leq \frac{\pi}{4T_d} \tag{20}$$

The controller gains must be chosen such that the loop gain is unity at ω_c:

$$\frac{\mathrm{K_i U_{b,max}^2} + K_u I_{b,max}^2}{\omega_c} = 1 \tag{21}$$

With equality of the summands in equation (21) the gains are:

$$\mathrm{K_i} \leq \frac{\pi}{\mathrm{8T_d U_{b,max}^2}}, \qquad \mathrm{K_u} \leq \frac{\pi}{\mathrm{8T_d I_{b,max}^2}} \tag{22}$$

Gain-scheduling based on $U_{b,max}$ and $I_{b,max}$ may be employed to improve transient performance when $U_{b,max}$ and $I_{b,max}$ are far from their worst-case values.

4 Conclusions

The capacitor voltages of all converter topologies built from series connected modules considered in this paper but the wye STATCOM can be fully balanced with the presented method without distorting the terminal currents and voltages in most cases. When no terminal currents are present and voltage is absent on one port of the converter terminal current deviations for balancing must be admitted on the port where voltage is present.

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