# On Dynamics and Voltage Control of the Modular Multilevel Converter

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## Keywords

« Converter control », « Multilevel converters », « Modulation strategy », « Emerging topology », « VSC »

## Abstract

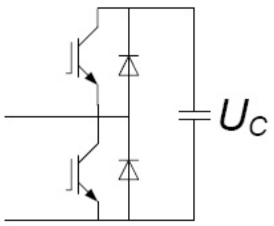
This paper discusses the impact of modulation on stability issues of the Modular Multilevel Converter (M2C). The main idea is to describe the operation of this converter system mathematically, and suggest a control method that offers stable operation in the whole operation range. A possible approach is to assume a continuous model, where all the modules in each arm are represented by variable voltage sources, and as a result, all pulse width modulation effects are disregarded. After simulating this model and testing different control methods, useful conclusions on the operation of the M2C have been extracted. The control methods are then implemented on a model with discrete half-bridge modules, in order to compare the results and to validate the continuous model approach. When assuring that this model functions as expected, the goal of this paper is to conclude into a self-stabilizing voltage controller. A controller is proposed, which eliminates circulating currents between the phase legs and balances the arm voltages regardless of the imposed alteranting current.

## Introduction

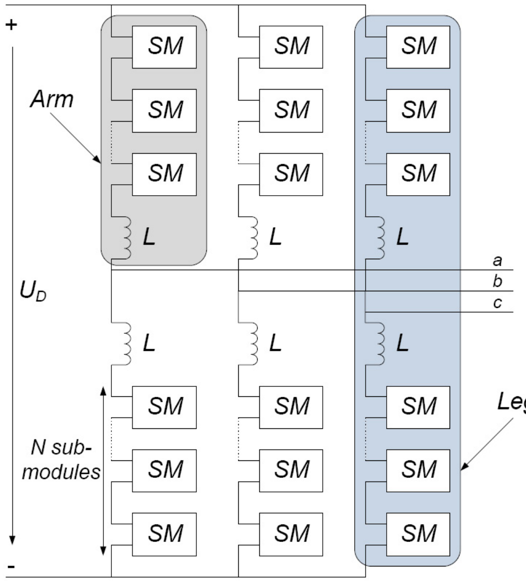
Modular multilevel converters seem to have great potential in energy conversion in the near future. High-power applications, such as dc interconnections, dc power grids, and off-shore wind power generation are in need of accurate power flow control and high-efficiency power conversion in order to reduce both their operating costs and their environmental impact.

The multilevel concept provides lower losses due to a considerably lower switching frequency compared to a 2-level equivalent. Apart from the lower switching frequency, the quality of the output voltage waveform is higher. This implies that smaller and simpler harmonic filters are required. To take this thought one step further, a modular multilevel converter provides simplicity of design and control, as well as scalability to various voltage and/or power levels. Moreover, the concept of a modular converter has the potential to improve the reliability, as a faulty module can be bypassed without significantly affecting the operation of the whole circuit.

A great advance in the design of modular topologies was made when Marquardt and Lesnicar suggested their design of the Modular Multilevel Converter (M2C) in 2002 [1] - [4], a design being investigated by several research teams lately [5]. In fact the basic circuit, using voltage sources instead of capacitors, was already proposed by Alesina et al. in 1981 [6]. The basic component of the M2C, here called a "submodule", is a simple half bridge with a capacitor bank, as shown in Figure 1a. Each phase leg of the converter has 2 arms, each one constituted by a number of series connected submodules. The number of the output steps depends straightly on the number of modules available in each arm. There is also a small inductor in each phase in order to take up the voltage difference, which is produced when a module is switched in or out. A simple schematic of this converter with $N$ modules per arm is shown in Figure 1b. By applying a simple modulation scheme, the converter operates as expected when viewed from the output, but a closer look inside the converter reveals high currents circulating between the phase legs. The existence of these currents implies that the submodules have to be rated for a higher current, but the most important effect is the energy transfer between the arms. Unless controlled, they can lead to instabilities as the energy stored inside the converter will not be balanced. The purpose of this paper is to investigate modulation strategies that can be applied to this type of converter and to suggest an effective way to control and balance the arm voltages.




(a)




(b)

Figure 1: One submodule (a), Schematic of a three-phase M2C (b)

## Description of the M2C concept and operation principle

Let the converter consist of $N$ submodules per arm (or half leg). In general, each arm is controlled by an insertion index $n(t)$, which is defined such that $n(t) = 0$ means that all $N$ modules in the arm are bypassed, while $n(t) = 1$ means that all $N$ modules in the arm are inserted. In a sampled system, $n(t)$ is sensed in the beginning of the interval $k$, during which the number of inserted modules will vary between $floor(Nn_k)$ and $ceil(Nn_k)$ with a switching instant determined by $frac(Nn_k)$.

Define $C^{arm}$ to be the capacitance of the series-connection of the capacitances $C$ of each and all of the submodules in the arm as in (1).

$$C^{arm} = \frac{C}{N} \tag{1}$$

The effective capacitance for the series-connection of the inserted capacitors then becomes as in (2)

$$C^{m} = \frac{C^{arm}}{n(t)} \tag{2}$$

where the superscript $m$ denotes the number of the arm.

Equal voltage sharing among the capacitors of each arm is ensured by a selection mechanism of the inserted or bypassed modules during each sampling period [4]. This process is based on the direction of the corresponding arm current. In case this current flows to the direction that charges the capacitors of the arm, the module that has the capacitor with the lowest charge will be selected to be inserted, or the module with the capacitor having the highest charge will be selected to be bypassed, depending on the requested action. The opposite strategy is followed in case the current flows to the direction that discharges the capacitors of the arm. In that case, if the requested action is to insert a module, the one with the highest charge will be selected, otherwise, if it is requested to bypass a module, then the selected will be the one with the lowest charge. Using large enough capacitors, this strategy will share the total dc link voltage equally among all modules inside each arm, ensuring that no module capacitor will be overcharged or totally discharged.

## General analysis of the arm current

Although it is possible to simulate the M2C using a full representation of all the individual submodules in the arms, such a procedure tends to be fairly complicated. It also produces a vast amount of output data, which can only be interpreted after reduction into average values for each arm. A simpler way is to consider a continuous model of the converter, wherein the switching frequency is assumed to be high relative the frequency of the produced output voltage. Additionally, it is assumed that the resolution in the output voltage is small compared to the amplitude of the output voltage. The first assumption thus means infinite switching frequency and the second an infinite number of modules in each arm. In this study it is assumed that the selection mechanism described above is used and that it provides equal voltage sharing among the submodules inside each arm. It is, therefore, possible to focus on the total energy inside the converter and the energy balance between the arms.

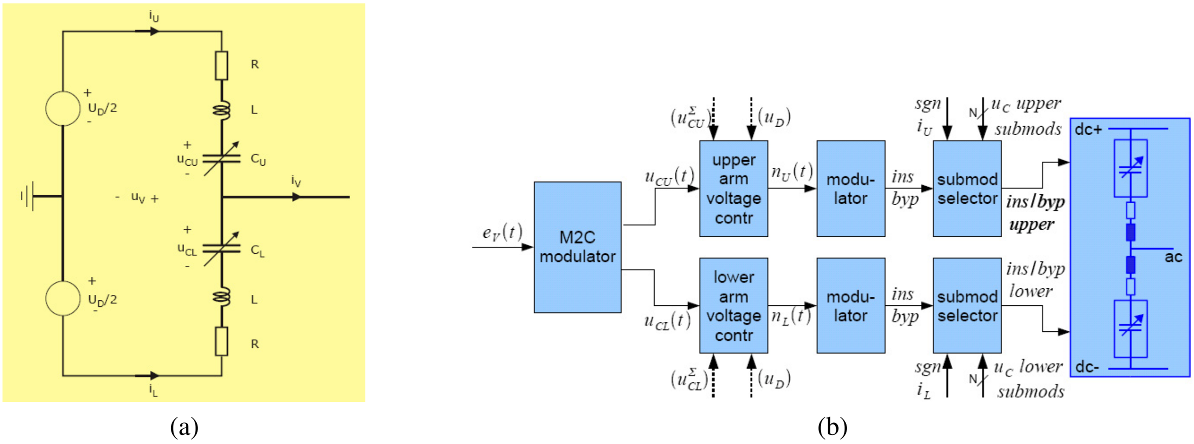



Figure 2: Continuous equivalent of a phase leg (a), Structure of suggested M2C leg control system (b)

Using this continuous model, it is possible to represent the arms as variable capacitances, or variable voltage sources, with their values depending on the insertion index $n(t)$, which is now a continuous variable. Let the available voltage (sum of all capacitor voltages) inside one arm be $u_C^{\Sigma}(t)$. Then, the voltage inserted by the arm $m$ is given by (3).

$$u_C^m = n(t)u_C^{\Sigma}(t) \tag{3}$$

When a charging current, $i(t)$ flows through the arm, where the effective capacitance $C^m$ is connected, then the total capacitor voltage in that arm will increase according to (4).

$$\frac{du_C^{\Sigma}(t)}{dt} = \frac{i(t)}{C^m} \tag{4}$$

Naming the upper and lower arm currents $i_U$, and $i_L$, their sum will constitute the output current $i_V$. However, in addition, a difference current $i_{diff}$ circulates through the phase leg and the dc link (or parallel phase legs). Accordingly,

$$\left.\begin{array}{l} i_U = \frac{i_V}{2} + i_{diff} \\ i_L = \frac{i_V}{2} - i_{diff} \end{array}\right\} \begin{array}{l} i_V = i_U + i_L \\ i_{diff} = \frac{i_U - i_L}{2} \end{array} \tag{5}$$

If $n_U$, $n_L$ are the insertion indices corresponding to the upper and lower arms, then from Eqs. (2) and (4) it is derived that

$$\frac{du_{CU}^{\Sigma}(t)}{dt} = \frac{n_U i_U}{C^{arm}} \tag{6}$$

$$\frac{du_{CL}^{\Sigma}(t)}{dt} = -\frac{n_L i_L}{C^{arm}} \tag{7}$$

Introducing the resistance $R$ and the inductance $L$ of each arm according to the circuit in Figure 2a, it is found that

$$\frac{u_D}{2} - Ri_U - L\frac{di_U}{dt} - n_U u_{CU}^{\Sigma} = u_V \tag{8}$$

$$-\frac{u_D}{2} - Ri_L - L\frac{di_L}{dt} + n_L u_{CL}^{\Sigma} = u_V \tag{9}$$

Subtracting Eq. (9) from (8) and after substituting the currents derived in (5) the following system of differential equations is obtained:

$$\frac{d}{dt}\begin{bmatrix} i_{diff} \\ u_{CU}^{\Sigma} \\ u_{CL}^{\Sigma} \end{bmatrix} = \begin{bmatrix} -\frac{R}{L} & -\frac{n_U}{2} & -\frac{n_L}{2} \\ \frac{n_U}{C^{arm}} & 0 & 0 \\ \frac{n_L}{C^{arm}} & 0 & 0 \end{bmatrix} \begin{bmatrix} i_{diff} \\ u_{CU}^{\Sigma} \\ u_{CL}^{\Sigma} \end{bmatrix} + \begin{bmatrix} \frac{u_D}{2} \\ \frac{n_U i_V}{2C^{arm}} \\ -\frac{n_L i_V}{2C^{arm}} \end{bmatrix} \tag{10}$$

As four of the elements in the 3-by-3 matrix are time variables, the system is non-linear.

## Direct Modulation, using a sinusoidally varying insertion index

A first approach of modulation can be an ideal open-loop modulation, where a sinusoidal reference is applied to each arm. Naming $\omega_N$ the angular frequency of the output voltage, the reference signal is

$$m(t) = \hat{m}cos(\omega_N t) \tag{11}$$

The ideal terminal voltage is given by Eq. (12).

$$u_V(t) = \frac{u_D}{2}m(t) \tag{12}$$

A stiff alternating current source $i_V(t)$ is assumed as the output load. In order to model different kinds of loads, the phase angle $\varphi$ of the load current can be chosen arbitrarily. Accordingly,

$$i_V(t) = \hat{i}_V cos(\omega_N t + \varphi) \tag{13}$$

The reference signal is translated by the modulator into insertion indices for each converter arm. Thus, the insertion indices for the upper and lower arm, respectively, are given by (14) and (15).

$$n_U(t) = \frac{1 - m(t)}{2} \tag{14}$$

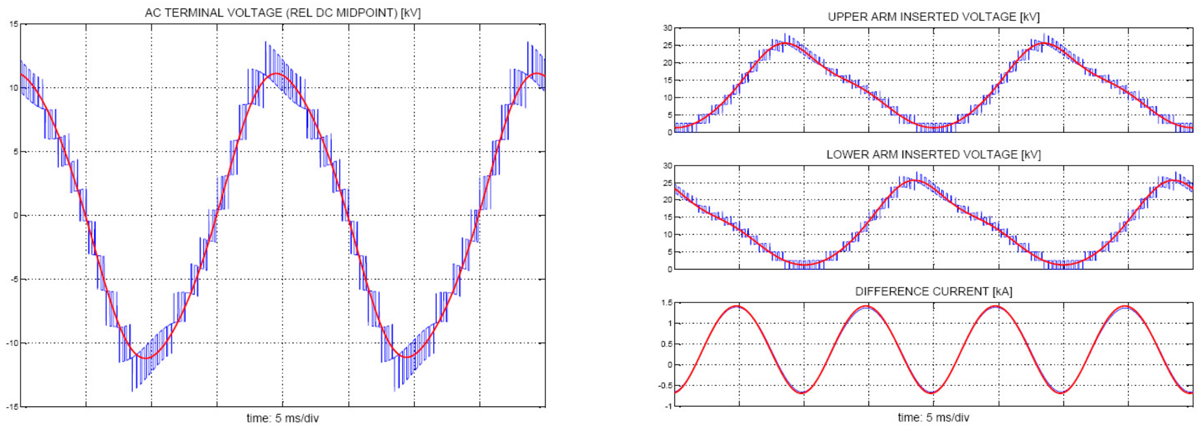
$$n_L(t) = \frac{1 + m(t)}{2} \tag{15}$$

The system given in Eq. (10) can be solved numerically, using the parameters described in Table I. The solution is then compared to the simulation of a discrete M2C with 10 modules per arm, using the simulation software PSCAD-EMTDC. A comparison between the resulting output voltages is shown in Figure 3a, while arm voltages and difference currents are shown in Figure 3b.

The results show that the circulating current contains a high $2^{nd}$ harmonic component that increases the RMS value of the current, thus increasing the losses. Apart from increasing the rating of the components used in the M2C, this current, if left uncontrolled, may create unbalance and disturbances at transients.

Table I: Direct modulation, Simulation parameters

| Rated Power | Line-line voltage | Phase current | Frequency | Direct voltage |
|---|---|---|---|---|
| 30 MVA | 13.8 kV rms | 1255 A rms | 50 Hz | 25 kV |
| Modules per arm | Module capacitance | Arm inductance | Arm resistance | Load |
| 10 | 5 mF | 3 mH | 0.1 Ω | 1 p.u., cosφ=1 |




(a) (b)

Figure 3: Comparison between results obtained using the continuous model and a PSCAD model with discrete submodules implemented, ac terminal voltages (a), inserted arm voltages and difference current $i_{diff}$ (b)

# Modulation and control using voltage references

## Energy dynamics and general circuit analysis

Another principle for modulation is based on the premise that the total capacitor voltage in each arm is being measured continuously. Then, the inserted voltage in the arm can be controlled arbitrarily by manipulating the insertion indices $n_U(t)$ and $n_L(t)$ according to Eq. (3). Hence, it is found that

$$u_V = \frac{u_{CL} - u_{CU}}{2} - \frac{R}{2} i_V - \frac{L}{2}\frac{di_V}{dt} \tag{16}$$

$$L\frac{di_{diff}}{dt} + Ri_{diff} = \frac{u_D}{2} - \frac{u_{CL} + u_{CU}}{2} \tag{17}$$

From Eqs. (16) and (17), the following conclusions can be drawn:

- The alternating voltage depends only on the alternating current $i_V$ and the difference between the arm voltages $u_{CL}$ and $u_{CU}$.
- The arm voltage difference acts as an inner alternating voltage in the converter and the inductance $L$ along with the resistance $R$, form a fix, passive inner impedance for the alternating current.
- The difference current, $i_{diff}$, only depends on the dc link voltage and the sum of the arm voltages. Therefore, subtracting the same voltage contributions from both arms will not affect the ac side quantities, but will impact the difference current instead.

Thus, it is reasonable to define the arm voltages, using also a difference voltage contribution, $u_{diff}$, for control purposes:

$$u_{CU} = \frac{u_D}{2} - e_V - u_{diff} \tag{18}$$

$$u_{CL} = \frac{u_D}{2} + e_V - u_{diff} \tag{19}$$

The inner alternating voltage described above is represented by the term $e_V$. Using Eqs. (18) and (19), equation (17) can then be written in a way that expresses the dependence of the difference current on the difference voltage. Accordingly,

$$L\frac{di_{diff}}{dt}+Ri_{diff}=u_{diff} \tag{20}$$

Assuming that the voltage, and consequently the energy stored in each arm, is equally shared among the modules, the total energies stored in the upper and lower arms are given by:

$$W_{CU}^{\Sigma}=N\left[\frac{C}{2}\left(\frac{u_{CU}^{\Sigma}}{N}\right)^2\right]=\frac{C}{2N}\left(u_{CU}^{\Sigma}\right)^2=\frac{C^{arm}}{2}\left(u_{CU}^{\Sigma}\right)^2 \tag{21}$$

$$W_{CL}^{\Sigma}=N\left[\frac{C}{2}\left(\frac{u_{CL}^{\Sigma}}{N}\right)^2\right]=\frac{C}{2N}\left(u_{CL}^{\Sigma}\right)^2=\frac{C^{arm}}{2}\left(u_{CL}^{\Sigma}\right)^2 \tag{22}$$

According to the circuit in Figure 2a, the stored energy in each arm deviates as shown in Eqs. (23) and (24).

$$\frac{dW_{CU}^{\Sigma}}{dt}=i_U u_{CU}=\left(\frac{i_V}{2}+i_{diff}\right)\left(\frac{u_D}{2}-e_V-u_{diff}\right) \tag{23}$$

$$\frac{dW_{CL}^{\Sigma}}{dt}=-i_L u_{CL}=\left(\frac{i_V}{2}+i_{diff}\right)\left(\frac{u_D}{2}+e_V-u_{diff}\right) \tag{24}$$

## Impact of the difference current

The total capacitor energy in the leg and the difference (unbalance) energy between the upper and the lower arms are defined as:

$$W_C^{\Sigma}=W_{CU}^{\Sigma}+W_{CL}^{\Sigma} \tag{25}$$

$$W_C^{\Delta}=W_{CU}^{\Sigma}-W_{CL}^{\Sigma} \tag{26}$$

Differentiating and inserting (23) and (24) yields

$$\frac{dW_C^{\Sigma}}{dt}=\left(u_D-2u_{diff}\right)i_{diff}-e_V i_V \tag{27}$$

$$\frac{dW_C^{\Delta}}{dt}=-2e_V i_{diff}+\left(\frac{u_D}{2}-u_{diff}\right)i_V \tag{28}$$

The Eqs. (27) and (28) show that $i_{diff}$ plays a crucial role for controlling the total capacitor energies in the converter arms. The dc component in $i_{diff}$ multiplies with the dc link voltage $u_D$ to balance the power delivered to the ac side plus the losses caused by the difference current itself (note that $i_{diff}$ and $u_{diff}$ always have the same frequency components according to (20)). On the other hand the dc component in $i_{diff}$ has no impact on the energy balance between the upper and the lower arm as long as there are no dc components in $e_V$ or $i_V$. Therefore, the dc component of the difference current can be used to control the total energy level (or voltage level) stored in the capacitors in each converter leg.

On the other hand, the fundamental frequency component in $i_{diff}$, which has the same frequency as the alternating output voltage $u_V$, impacts on the distribution of the capacitor energies between the upper and lower arms in a converter leg. Let the created emf in the converter and the difference current contain a component with the same frequency and phase, such that

$$e_V = \hat{e}_V cos\left(\omega t + \varphi\right) \tag{29}$$

$$i_{diff} = I_{diff} + \hat{i}_{diff} cos\left(\omega t + \varphi\right) \tag{30}$$

In such a case, a dc component occurs in the product $e_V i_{diff}$ and forces the energy difference $W_C^\Delta$ to change. A similar effect is created by the product of the corresponding difference voltage and the alternating current. However, (at high output voltage amplitude and for reasonably small arm inductance values) this component is smaller than the one discussed earlier. Therefore, in a first approach it can be neglected.

## Suggested modulation and control system

The circuit diagram shows that the arm currents must contain at least a fundamental frequency component and a dc component such that:

- the sum of the fundamental frequency components in the upper and the lower arm currents, $\tilde{i}_U$ and $\tilde{i}_V$ respectively, equals the alternating output current.
- the product of the difference current and the dc link voltage provides the power that is transferred to the ac side.

Ideally, only these two current components exist in the converter

$$\tilde{i}_U = \tilde{i}_L = \frac{i_V}{2} \tag{31}$$

$$i_{diff} = i_{diff,dc} \tag{32}$$

This operating condition also implies the minimum RMS value for the arm currents. It was shown earlier that with direct modulation the arm currents also contain a substantial second harmonic component.

In this paper a modulation principle is proposed, which eliminates the harmonics in the arm currents and controls the capacitor energies in the arms under various loading conditions. The modulation concept assumes that measured values for the total capacitor voltages (sum of all submodule capacitor voltages) in the arms are available at any time.

At first glance it appears simple to modulate the converter if the measured total voltages are available, just by using (3). It can be argued that $u_{diff}$ may be neglected if the arm impedances $(R, L)$ are small and then $u_{CU}$ and $u_{CL}$ can be calculated from the desired inner emf, $e_V$ and the dc link voltage $u_D$. The modulation indices can then be obtained from (33) and (34).

$$n_U(t) = \frac{u_{CU}(t)}{u_{CU}^{\Sigma}(t)} \tag{33}$$

$$n_L(t) = \frac{u_{CL}(t)}{u_{CL}^{\Sigma}(t)} \tag{34}$$

However, tests have shown that the capacitor voltages become unstable if this modulation approach is adopted without any further stabilization. This section describes the mentioned method with two stabilizing controllers added, as illustrated in Figure 4a.

When the total capacitor voltages are measured the corresponding capacitor energies $W_{CU}^{\Sigma}$ and $W_{CL}^{\Sigma}$ can be obtained from the Eqs. (21) and (22).

From the obtained values for the energies stored in the upper and lower arm, the total and unbalance energies, $W_C^{\Sigma}$ and $W_C^{\Delta}$, are derived and brought to a total energy controller and a balance controller respectively. The total energy controller is used to regulate the total energy in the converter leg to a desired reference level, $W_C^{\Sigma,ref}$, which typically corresponds to the energy in the two arms in the converter leg when they are both charged to the dc link voltage level $u_D$. This controller uses a PI controller to determine an additional dc component, which is added to $u_{diff}$. The integral part of the controller is necessary in order to eliminate static errors in the average energy level for various active output power from the ac circuit. Without the integral part the average value of the capacitor voltage drops when active power is supplied form the dc side to the ac side and increases when power is transferred from the ac side to the dc side. *Note that the reference for the total capacitor energy can be freely selected in run-time with this control method.*

The balance controller determines the amplitude for an ac component that is added to the difference current, $i_{diff}$, which aims to cancel the energy difference between the upper and lower arms. Two problems are related to this difference current component. The first is that it must be provided in phase with the generated inner emf $e_V$ in the converter. The second problem is that the difference current is not directly controlled, but it must be manipulated through the difference voltage $u_{diff}$. According to Eq. (20) the phase of $u_{diff}$ must lead that of $i_{diff}$ by the phase angle of the arm impedance $R_{arm} + jX_{arm}$. At high frequencies the reactance in the impedance is dominating and the corresponding phase advance is close to $\pi/2$, but for lower frequencies the phase shift is smaller. It shall further be noted that the measured unbalance energy $W_C^{\Delta}$ in steady state contains a considerable fundamental frequency component, specifically if the submodule capacitors have low capacitance. For the control, however, only the average of the unbalance energy is relevant. Therefore, a filter that extracts the average components of $W_C^{\Delta}$ is necessary. In its simplest form this filter may be just a lowpass filter with a low cutoff frequency. In the simulations a filter time constant of 100 ms has been used. In Figure 4a, a reference for the unbalance energy controller has been indicated, which in normal operating conditions is set to zero.

## Simulation results

### Steady-state operation

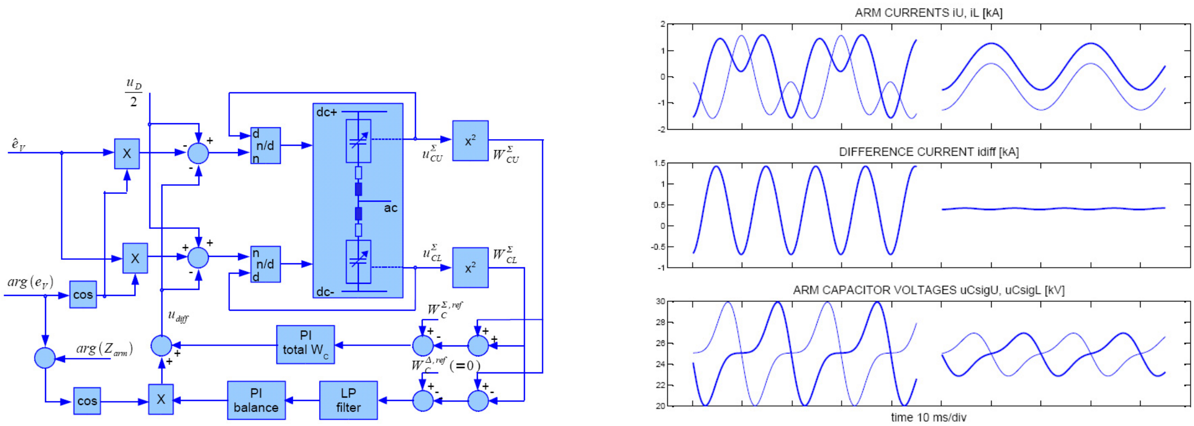
The proposed modulation and control concept has been simulated with Simulink. Figure 4b shows the compared waveforms obtained by a direct modulation and the proposed method.

Figure 4b shows that the second order harmonics in the arm currents have been totally eliminated such that the arm current contains exclusively a dc component and half the ac fundamental frequency component. As a result the difference current only comprise a dc component corresponding to the active load that is delivered to the ac side. This dc current theoretically is $30MW/25kV$ distributed among the three phases.

The significant second harmonic current in the difference current also contributes to the second harmonic voltage component in the capacitor voltages of the arm. The contribution is eliminated in the proposed modulation method. On the other hand, a second harmonic voltage still exists as a result of the multiplication of the insertion index and the fundamental frequency component in the arm current. It should be noted that the second order harmonic component in the capacitor voltage of the arm is not always a disadvantage; for load currents at certain power factors it may expand the available voltage range, which otherwise may be limited by the rippple of the capacitor voltage of the arm.

### Dynamical control of the arm capacitor voltages

For various reasons, it is very important to control the capacitor voltages of the arms. First, it is critical to provide stability under all loading conditions, as it directly determines the electrical stress on the semiconductors and capacitors in the submodules. Another justification is that the possibility to control

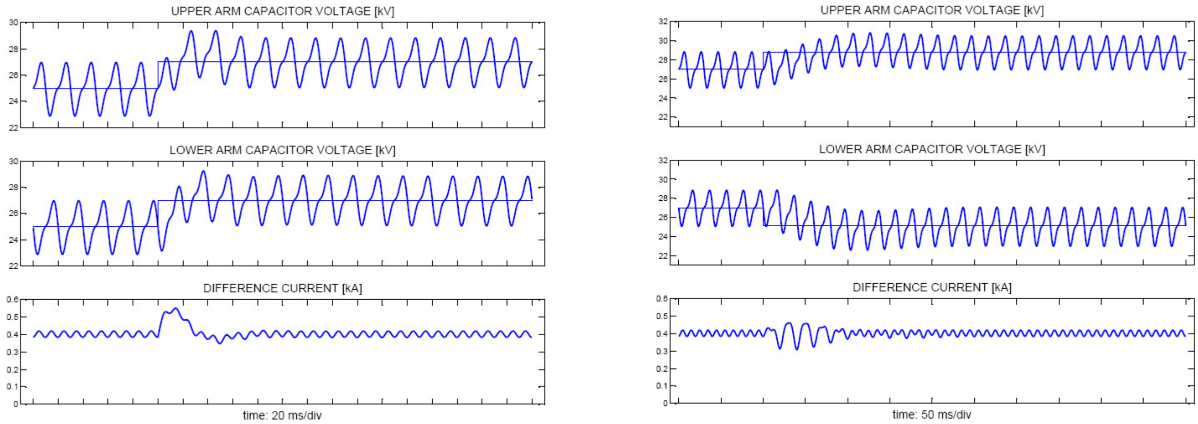



(a) (b)

Figure 4: Outline of the proposed control and modulation method (a), Comparison of steady-state waveforms (b)

the arm capacitor voltage in run-time adds flexibility in the application of the converter, which can be advantageously utilized in many ways. One such approach is briefly described later in this paper.

In the theoretical description of the M2C it was mentioned that the modulation only impacts on the derivative of the stored energy in the arm capacitors and that, accordingly, the average energy level can be freely selected and controlled by a regulator. Figures 5a and 5b show simulation results when step changes are executed for the energy controllers. In Figure 5a the reference for the average energy level in the converter is changed, while in Figure 5b the reference for the arm voltage unbalance is changed (the latter naturally is zero in normal operation and it is made non-zero here only for demonstration purposes).




(a) (b)

Figure 5: Response to step change of energy references, symmetrical(a), asymmetrical (intentionally created unbalance for demonstration) (b)

It is to be noted that the symmetrical change of the arm capacitor voltages is performed by a temporary dc component in the difference current, while the unsymmetrical change is executed by a temporary fundamental frequency component.

### Adaptation of the arm capacitor voltages to the amplitude of the ac output voltage

The capacitor voltages of the arm $u_{CU}^{\Sigma}$, $u_{CL}^{\Sigma}$ must be kept sufficiently high to produce the maximum required arm voltages $u_{CU}$, $u_{CL}$. When the maximum ac output voltage, $e_V = u_D/2$, is required the inserted arm voltages reaches the same level as the dc link, but sometimes the output voltage may be restricted to a lower value, $\hat{e}_{Vmax}$. Then the maximum inserted arm voltage is only

$$u_{CUmax} = u_{CLmax} = \frac{u_D}{2} + \hat{e}_V \tag{35}$$

and accordingly the capacitor voltage of the arm can be reduced to this value. Such actions serves several purposes. It allows a larger ripple in the submodule capacitor voltages without causing too high peak voltages, which might be desired e.g. when a motor operates at low speed. Further, the converter losses decrease approximately proportional to the submodule capacitor voltages. Figure 6 shows simulation results for such a scenario, when the ac output voltage has been restricted to 50% of the rated voltage in Table I. The frequency of the output voltage also has been reduced to 25 Hz.

## Conclusion

In this paper a method to modulate and stabilize an M2C has been presented. It has been shown that the difference current is an intermediate quantity which can be controlled independently of the ac output voltage with a direct influence on the dynamics of the energy stored in the capacitors in the arms of the converter. In this application, after the use of the energy controllers, the difference current contains only a dc component in steady state. For stabilizing the capacitor voltages, the dc component is estimated from the outcome of the total energy (voltage) controller and the fundamental frequency component is used by the energy balance controller to mitigate unbalanced capacitor voltages in the arms. Sometimes it would, however, also be of interest to induce a second order harmonic component in the difference current on purpose, as a mean to minimize the dc capacitor voltage variation and to impact on the maximum voltage that can be produced at various load currents.

Simulations further have shown that the difference current is a flexible tool that can be used to determine the characteristics and limitations of the M2C.

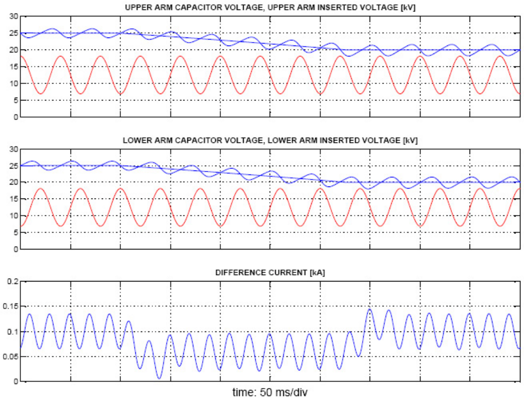



Figure 6: Arm capacitor voltage reduction in low load conditions

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