Digital control of a pulsed current source converter with low reactions on mains

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Abstract
To satisfy the demands of more stringent norms dealing with the electromagnetic compatibility of power electronic equipment one has to be increasingly concerned about the reactions on mains. In an ideal case the power supply should be loaded with sinusoidal current and voltage. For this a closed loop digital control of a pulsed current source converter is presented. The calculation of the pulsed current is done on-line. The system consisting of mains, converter and closed loop control is completely dealt with in a time-discrete state-space.
On the basis of simulation results the steady-state and the dynamic operation of the resulting system is shown.

1. - INTRODUCTION

A growing amount of converters use switchable power semiconductors (e.g. GTO, IGBT) to make use of the advantages concerning the steady-state and the dynamic behaviour of pulsed converters compared with diode- or thyristor bridges.
During steady-state operation the load is less stressed by harmonic distortion. This leads to higher efficiency. The increased dynamic is advantageous during quick load variations.
Using pulse converters on the mains-side leads to a circuitry with little system perturbation.
Latest developments in the field of power electronics and microelectronics enable us to employ known pulsed converter concepts at increasing pulse frequency combined with a digital control.
The control can be classified into open and closed loop controls [4]. The advantages of open loop controls are their good steady-state behaviour and their simplicity. The disadvantage is their poor dynamic quality. Closed loop controls combine good steady-state and dynamic quality. Because of the required intensive calculations closed loop controls are only realizable with reasonable hardware expense for a limited pulse frequency.

In this paper a new closed loop digital control for a pulsed current source converter on the line side is presented. The control ensures an optimal steady-state and dynamic behaviour of the converter at a limited pulse frequency. In addition mains is loaded with sinusoidal current and voltage.

2. - PULSED CURRENT SOURCE CONVERTER

Because it is not possible to draw pulsed currents out of the inductive mains capacitors have to be added (Figure 1). Together with the reactance a resonant circuit with very low damping is formed. An excitation of the resonant circuit has to be avoided with regard to low reactions on mains. For this several concepts are found in literature.
The classic open loop control generates the pulses by comparing a sinusoidal reference curve with a triangular curve with higher frequency. This is a subharmonic concept. By using a switching frequency higher than the resonant frequency a stationary excitation of the resonant circuit is prevented. An excitation during dynamic processes cannot be prevented. An improved dynamic performance can be achieved by optimizing the pulse-

patterns [2]. But the amount of calculation makes an on-line calculation impossible. An additional disadvantage of the above mentioned open-loop control concepts is the impossibility to directly infect the line currents because of the capacitors. This can be avoided by adding two additional power-semiconductors [5]. The operation of the circuit in Figure 1 with a closed loop control shows none of the mentioned disadvantages. But in the literature we mainly find non-linear concepts [1]. Newer approaches control a on-line calculated pulse pattern to cope with dynamic processes [6]. A state-space control is used.

Now a concept is presented that completely describes the system mains, converter and closed loop control in a time-discrete state-space. The purpose is to control a sinusoidal line current at variable power factor. The results can also be transferred from the mains-side to the load-side quite easily.

3. - CONTROL MODE AND CONCEPTS

The circuit in Figure 1 can be described by the following sixth order system of differential equations (1). The equations are coupled by the three capacitor voltages U_{C1} to U_{C3}.

$$L\cdot\frac{\mathrm{d}}{\mathrm{d}t}\begin{pmatrix}I_{N1}(t)\\I_{N2}(t)\\I_{N3}(t)\end{pmatrix}+R\cdot\begin{pmatrix}I_{N1}(t)\\I_{N2}(t)\\I_{N3}(t)\end{pmatrix}=$$

$$\begin{pmatrix}U_{N1}(t)\\U_{N2}(t)\\U_{N3}(t)\end{pmatrix}+\frac{1}{3}\cdot\begin{pmatrix}-2&1&1\\1&-2&1\\1&1&-2\end{pmatrix}\cdot\begin{pmatrix}U_{C1}(t)\\U_{C2}(t)\\U_{C3}(t)\end{pmatrix}$$

$$\frac{\mathrm{d}}{\mathrm{d}t}\begin{pmatrix}U_{C1}(t)\\U_{C2}(t)\\U_{C3}(t)\end{pmatrix}=\frac{1}{C}\cdot\left(\begin{pmatrix}I_{N1}(t)\\I_{N2}(t)\\I_{N3}(t)\end{pmatrix}-\begin{pmatrix}I_{P1}(t)\\I_{P2}(t)\\I_{P3}(t)\end{pmatrix}\right)\tag{1}$$

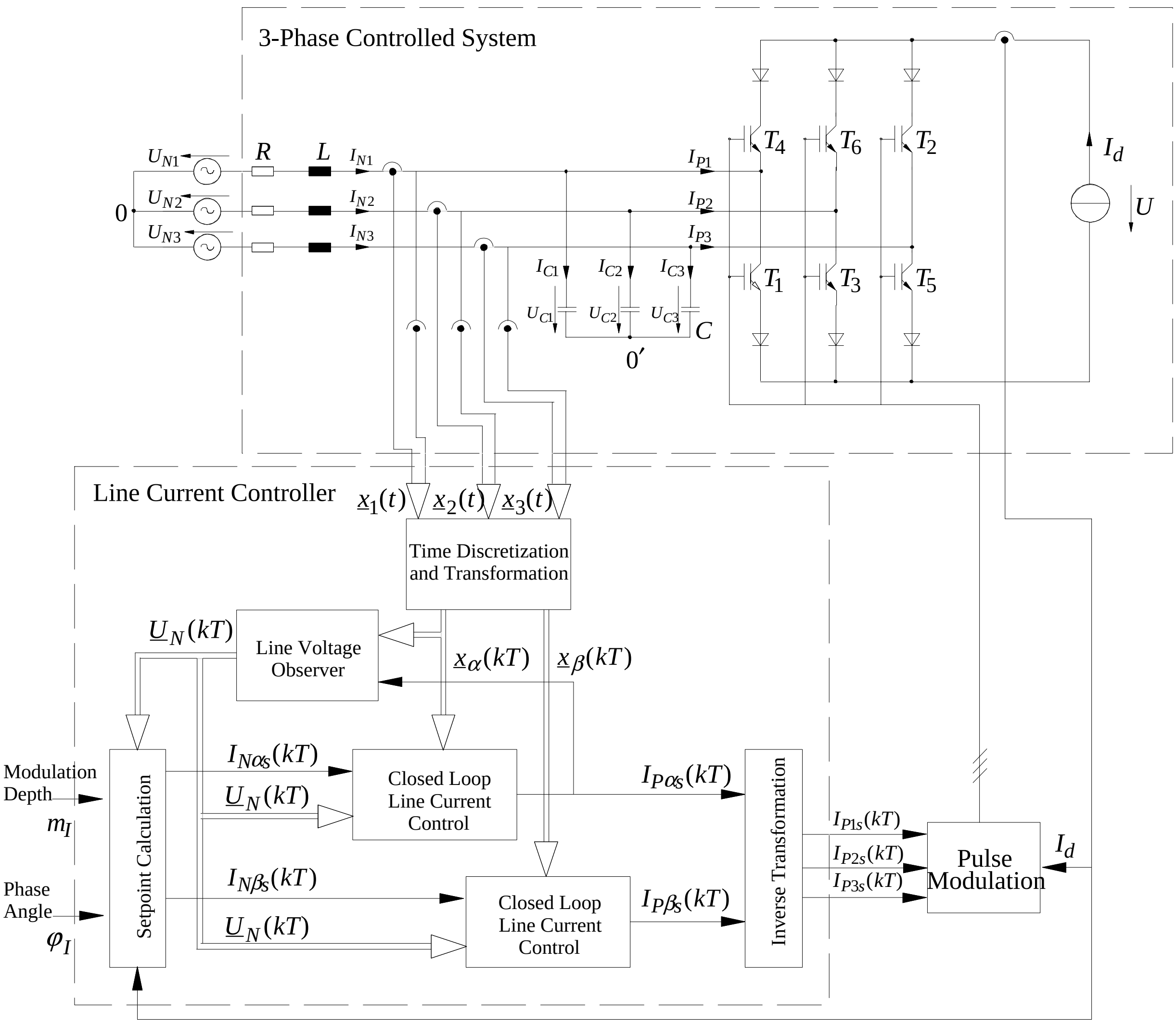

Figure 1 : System overview of the line current control of the 3-phase pulsed current source converter

With not connected star point and balanced circuit the six dependent equations can be converted into four independent equations using the αβ0-transformation. The advantage of the αβ0-transformation is that the resulting components are real numbers. The 0-component vanishes because of the balanced circuit. Doing this for (1) with a following time-discretization and a rearrangement of the resulting terms we obtain (2).

The abbreviation $\underline{x}(k\cdot T)=\begin{pmatrix} U_C(k\cdot T)\\ I_N(k\cdot T)\end{pmatrix}$ is used.

$$\begin{pmatrix}\underline{x}_\alpha((k+1)\cdot \mathrm{T})\\ \underline{x}_\beta((x+1)\cdot \mathrm{T})\end{pmatrix}=\begin{pmatrix}\underline{\underline{A}} & \underline{\underline{0}}\\ \underline{\underline{0}} & \underline{\underline{A}}\end{pmatrix}\cdot\begin{pmatrix}\underline{x}_\alpha(k\cdot \mathrm{T})\\ \underline{x}_\beta(k\cdot \mathrm{T})\end{pmatrix}+ \begin{pmatrix}\underline{b} & \underline{0}\\ \underline{0} & \underline{b}\end{pmatrix}\cdot\begin{pmatrix}U_{N\alpha}(k\cdot \mathrm{T})\\ U_{N\beta}(k\cdot \mathrm{T})\end{pmatrix}+ \begin{pmatrix}\underline{c} & \underline{0}\\ \underline{0} & \underline{c}\end{pmatrix}\cdot\begin{pmatrix}I_{P\alpha}(k\cdot \mathrm{T})\\ I_{P\beta}(k\cdot \mathrm{T})\end{pmatrix} \quad (2)$$

The coefficients $\underline{\underline{A}}$, $\underline{b}$ and $\underline{c}$ result to (3).

$$\underline{\underline{A}}=\begin{pmatrix}a_{11} & a_{12}\\ a_{21} & a_{22}\end{pmatrix}$$
$$\underline{b}=\begin{pmatrix}1-a_{11}\\ -a_{21}\end{pmatrix} \text{ and } \underline{c}=\begin{pmatrix}R\cdot(a_{11}-1)-a_{12}\\ 1-a_{11}\end{pmatrix} \text{ with :} \quad (3)$$

$$a_{11}=\mathrm{e}^{-\delta\cdot \mathrm{T}}\cdot\left(\cos(\omega\cdot T)+\frac{\delta}{\omega}\cdot\sin(\omega\cdot T)\right)$$
$$\approx 1-\frac{\omega'^2}{2}\cdot T^2$$
$$a_{12}=\frac{1}{\omega\cdot C}\cdot \mathrm{e}^{-\delta\cdot \mathrm{T}}\cdot\sin(\omega\cdot T)$$
$$\approx \frac{1}{C}\cdot T-\frac{R\cdot\omega'^2}{2}\cdot T^2$$
$$a_{21}=-\frac{1}{\omega\cdot L}\cdot \mathrm{e}^{-\delta\cdot \mathrm{T}}\cdot\sin(\omega\cdot T)$$
$$\approx -\frac{1}{L}\cdot T+\frac{R}{2\cdot L^2}\cdot T^2$$
$$a_{22}=\mathrm{e}^{-\delta\cdot \mathrm{T}}\cdot\left(\cos(\omega\cdot T)-\frac{\delta}{\omega}\cdot\sin(\omega\cdot T)\right)$$
$$\approx 1-\frac{R}{L}\cdot T+\left(\frac{R^2}{2\cdot L^2}-\frac{\omega'^2}{2}\right)\cdot T^2$$

The following abbreviation are used :

$$\delta=\frac{R}{2\cdot L}\ ,\quad \omega=\sqrt{\frac{1}{L\cdot C}-\delta^2}\ ,\quad \omega'=\sqrt{\frac{1}{L\cdot C}}$$

The approximations in (3) result from a series expansion aborted after the square term. If δ is assumed to zero, what is permitted on the mains-side, one gets still more compact coefficients.

It can be concluded from (2) that the 3-phase controlled system can be devided into two 1-phase controlled systems. Because of this they will be discussed first. Figure 2 depicts the state-space description. Adding two orthogonal controls of a 1-phase controlled system results in a control of the 3 phase system as already shown in Figure 1.

The state variable $\underline{x}$ consists of the measurable quantities line current I_N and capacitor voltage U_C. The not measurable line voltage U_N acts as a disturbance. Output of the line current control is the manipulated variable I_{Ps} which is applied onto the controlled system after pulse modulation.

The inevitable line voltage observer can be composed of a reduced order observer using the state variable as input for a sinusoidal model with known line frequency $\omega_1 = 2\pi\cdot f_1$. In this paper the observer will not be discussed in detail.

The setpoint is calculated using the time independent setpoint values modulation depth m_I, phase angle φ_I and DC link current I_d (4).

$$I_{Ns}(k\cdot T)=m_I\cdot I_d\cdot\sin\left(\arctan\left(\frac{U_{N\alpha}(k\cdot T)}{U_{N\beta}(k\cdot T)}\right)-\varphi_I\right) \quad (4)$$

Within the framework [3] forming the basis of this paper several control modes were analysed. They are summarized in Figure 2. The explicit calculation is a dead-beat control. It tries to adjust the line current within one sampling interval T. Because the manipulating variable is limited to the DC-link current, a large setpoint stepchange leads to a prolongation of the settling time. To remedy this disadvantage, a state space control with different methods of compensating disturbances is realized. The dynamic quality of the control can then be influenced by assigning the poles of the characteristic polynom. The tested non-linear control modes have problems coping with the low damping of the controlled system.

In this paper only the state space control with prefilter and with a PI-controller are compared because they show the best results.

Figure 3 shows the state space control with prefilter. First the influence of the disturbing line voltage is ignored. After z-transformation this leads to the second order transfer function (5).

$$\frac{\underline{x}(z)}{I_{Ns}(z)}=\left(z\cdot\underline{\underline{I}}-\underline{\underline{A}}+\underline{c}\cdot\underline{k}^T\right)\cdot V\cdot\underline{c} \quad (5)$$

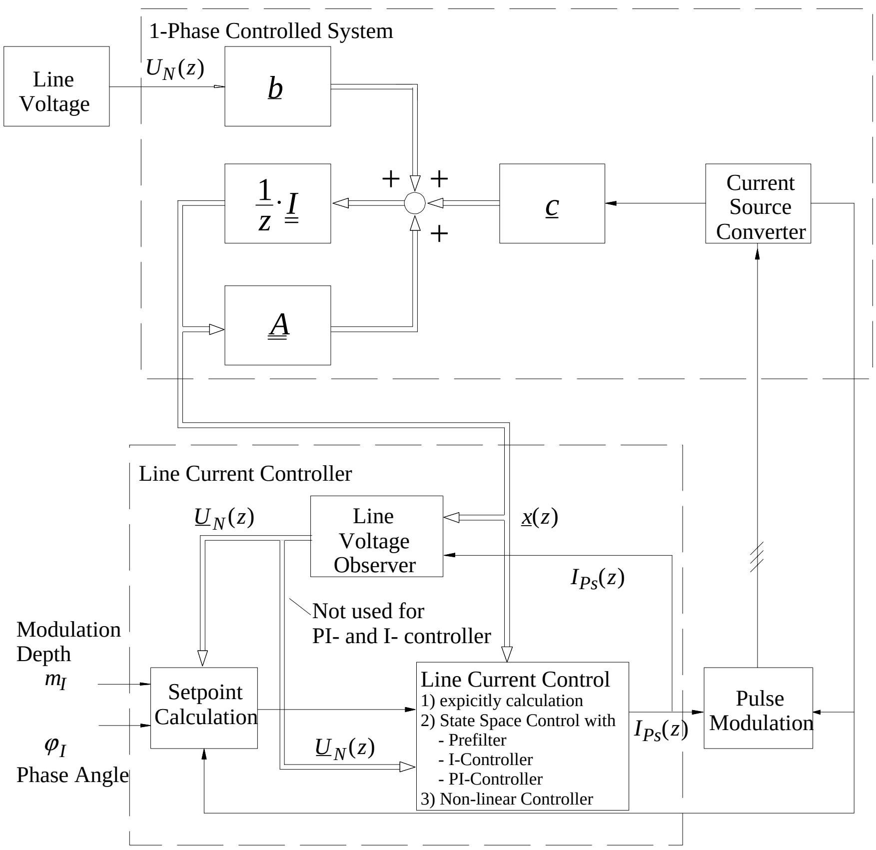

Figure 2 : State-space description of the 1-phase controlled system

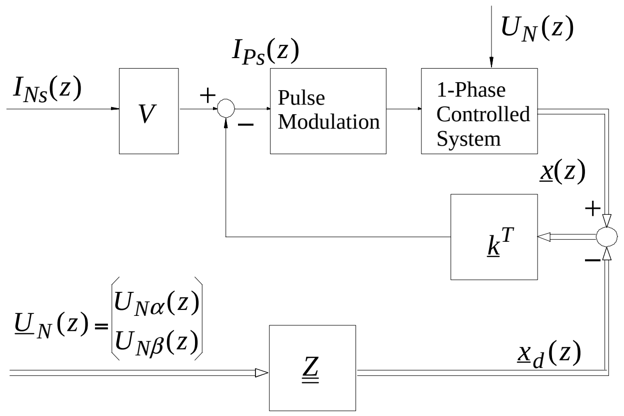

Figure 3 : State space control with prefilter

The poles are assigned according to the desired control dynamic by assigning the elements of the feedback vector $\underline{k}$. The prefilter V is necessary to eliminate the static system deviation. It can be determined using the final-value theorem on (5). The effects of the line voltage onto the state vector can be compensated by using (6) with the observed line voltage as input.

$$\underline{\underline{Z}} = \begin{pmatrix} \frac{-\sin(\varphi_N)}{\omega_1 \cdot C \cdot Z_N} & \frac{\cos(\varphi_N)}{\omega_1 \cdot C \cdot Z_N} \\ \frac{\cos(\varphi_N)}{Z_N} & \frac{\sin(\varphi_N)}{Z_N} \end{pmatrix} \approx \begin{pmatrix} 1 & 0 \\ 0 & -\omega_1 \cdot C \end{pmatrix} \qquad (6)$$

The following abbreviations are used :

$$Z_N = \sqrt{R^2 + \left(\omega_1 L - \frac{1}{\omega_1 C}\right)^2}, \quad \varphi_N = \arctan\left(\frac{\omega_1 LC - 1}{\omega_1 RC}\right)$$

and $\underline{U}_N(k \cdot T) = \begin{pmatrix} U_{N\alpha}(k \cdot T) \\ U_{N\beta}(k \cdot T) \end{pmatrix}$

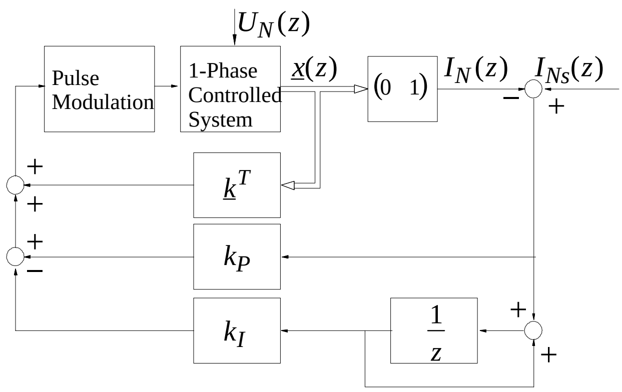

Figure 4 : State space control with PI-controller

The manipulated variable can be calculated with (7).

$$I_{Ps}(kT) = -\underline{k}^T \cdot \left(\underline{x}(kT) - \underline{\underline{Z}} \cdot \underline{U}_N(kT)\right) + V \cdot I_{Ns}(kT) \qquad (7)$$

Figure 4 shows the state space control with PI-controller. Now the influence of the disturbing line volt-

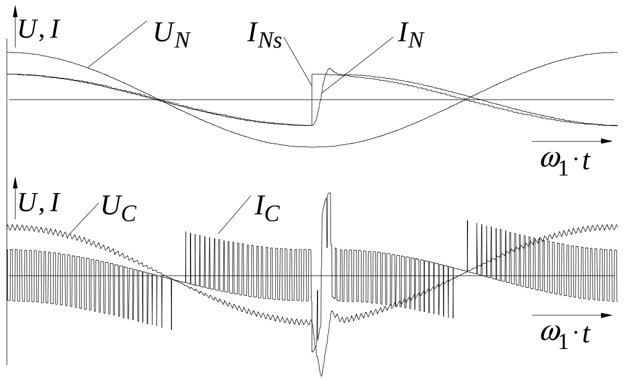

Figure 5 : Simulation results, 1-phase line current control with prefilter,
$m_I = 100\%$ **and** $\varphi_I = 0° \rightarrow 180°$

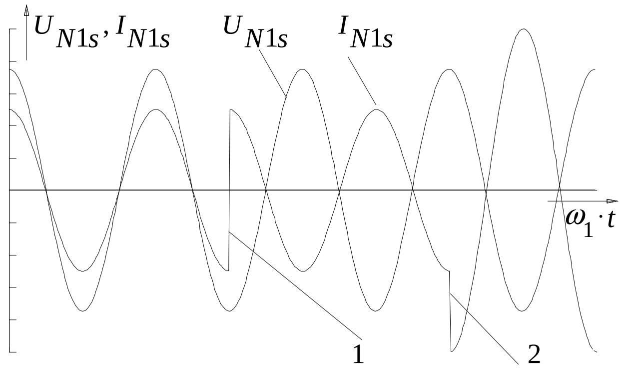

Figure 6 : 3-phase control, setpoint step change
1) $m_I = 50\%$ and $\varphi_I = 0° \rightarrow 180°$
2) $\varphi_I = 180°$ and $m_I = 50\% \rightarrow 100\%$

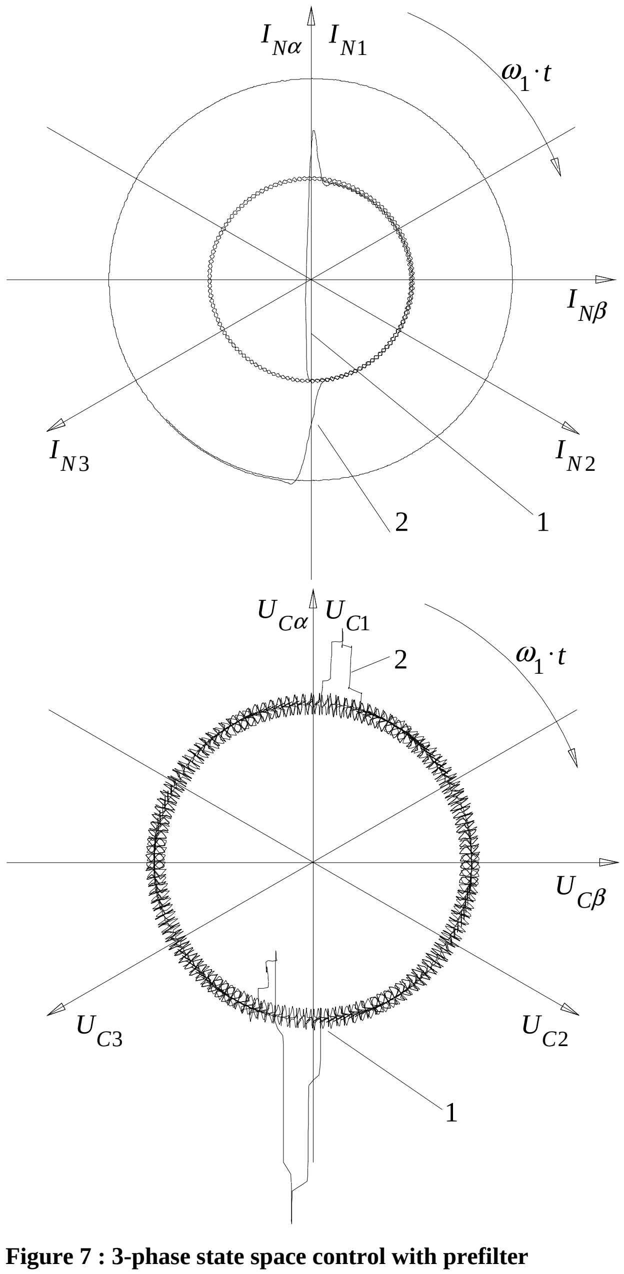

Figure 7 : 3-phase state space control with prefilter

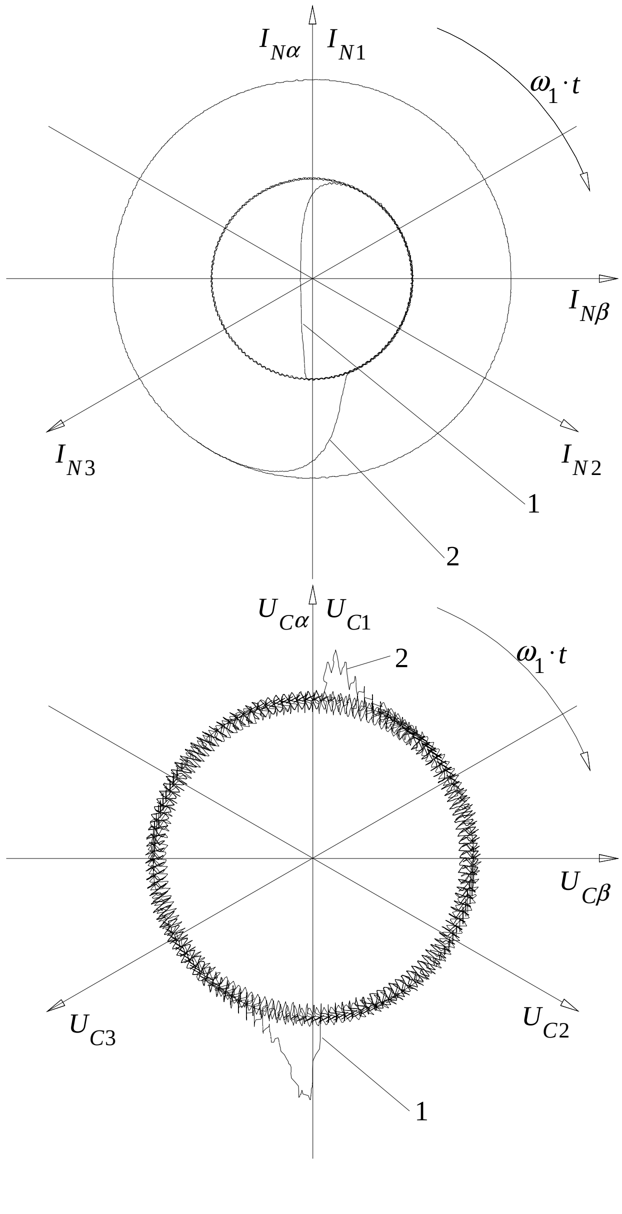

Figure 8 : 3-phase state space control with PI-controller

age is compensated using an additional PI-controller. Its I-term increases the system order to three (8).

$$\begin{pmatrix} \underline{x}((k+1)\cdot T) \\ q((k+1)\cdot T) \end{pmatrix} = \begin{pmatrix} \underline{\underline{A}} & \underline{0} \\ \underline{0} \; -1 & 1 \end{pmatrix} \cdot \begin{pmatrix} \underline{x}(kT) \\ q(kT) \end{pmatrix} + \begin{pmatrix} \underline{c} \\ 0 \end{pmatrix} \cdot I_P(kT) + \begin{pmatrix} \underline{0} \\ 1 \end{pmatrix} \cdot I_{Ns}(kT) \tag{8}$$

The manipulated variable can be calculated with (9).

$$I_{Ps}(kT) = \underline{k}^T \cdot \underline{x}(kT) + k_P \cdot \left(q((k+1)\cdot T) - q(kT) \right) - k_I \cdot q(kT) \tag{9}$$

We choose $k_P = -1$ for a quick step response and calculate $\underline{k}$ and k_I after assigning the poles of the characteristic polynom according our demands.

Now we have to combine two orthogonal controls of a 1-phase controlled system as already mentioned. To do this only the setpoint calculation and the line voltage observer have to adapted according to (10).

$$I_{N\alpha s}(k\cdot T) = I_{Ns}(k\cdot T)$$

$$I_{N\beta s}(k\cdot T) = m_I \cdot I_d \cdot \cos\left(\arctan\left(\frac{U_{N\alpha}(k\cdot T)}{U_{N\beta}(k\cdot T)} \right) - \varphi_I \right)$$

$$\underline{\underline{Z}}_\alpha = \underline{\underline{Z}} \tag{10}$$

$$\underline{\underline{Z}}_\beta = \begin{pmatrix} \frac{-\cos(\varphi_N)}{\omega_1 \cdot C \cdot Z_N} & \frac{-\sin(\varphi_N)}{\omega_1 \cdot C \cdot Z_N} \\ \frac{-\sin(\varphi_N)}{Z_N} & \frac{\cos(\varphi_N)}{Z_N} \end{pmatrix} \approx \begin{pmatrix} 0 & 1 \\ \omega_1 \cdot C & 0 \end{pmatrix}$$

The pulse modulation, especially the minimization of the switching frequency to 2/3 of the sampling frequency, will not be discussed in this paper.

4. - SIMULATION RESULTS

The simulations were done using the networksimulation software NETASIM with the parameters in Table 1.

First results of the 1-phase control are presented. Figure 5 shows a reversing at the worst angle. The state space control with prefilter was used. Line voltage and current are both sinusoidal.

The results of the 3-phase control are presented using the state space control with prefilter and with PI-controller. Two setpoint step changes according to Figure 6 are examined. The simulation results of the line current and the capacitor voltage are presented in Figure 7 and Figure 8 using a state space depiction. The two mentioned setpoint step changes are marked as 1 and 2. They mainly differ in their dynamic behaviour. The I-term of the PI-controller reduces the overswing. But it also increases the phase shift during steady-state operation.

Controlled System	Controller
$R = 1\,\mathrm{m}\Omega$	$\frac{1}{T} = 6400\,\mathrm{Hz}$
$L = 1\,\mathrm{mH}$	
$C = 63{,}3\,\mu\mathrm{F}$	$f_D = 2500\,\mathrm{Hz}$ (poles)
$U_N = 230\,\mathrm{V}$	$D_R = \frac{1}{\sqrt{2}}$ (poles)
$f_1 = 50\,\mathrm{Hz}$	
$I_d = 50\,\mathrm{A}$	$\tau_R = 400\,\mu\mathrm{s}$ (poles)

Table 1 Simulation parameters

5. - CONCLUSIONS

A new closed loop digital control for pulsed current source converters has been presented. Despite the low pulse frequency sinusoidal line current and voltage with low reactions on mains can be obtained. A comparison of all mentioned control modes can be taken from [3].

Because the orthogonal currents are controlled separately the presented controls are also suitable for the load side, i.e. vector control of AC-machines.

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