A Constant Output Current Three-Phase Diode Bridge Employing a Novel "Electronic Smoothing Inductor"

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Abstract — The paper presents a novel extension of a conventional three-phase diode bridge rectifier with capacitive smoothing of the output voltage which improves the power factor at the AC input and which reduces the ripple current stress on the smoothing capacitor. The basic concept of the proposed system is the arrangement of an active element between the output of the diode bridge and the smoothing capacitor which is controlled in a way that it "emulates" an ideal smoothing inductor. With this the input currents of the diode bridge showing usually high peak amplitudes are transformed to a 120° rectangular shape which ideally result in a total power factor of 0.955. The active element mentioned before is realized by a low-voltage switch-mode converter stage of small power rating as compared to the output power of the rectifier.
Starting from a discussion of the drawbacks of three-phase diode rectifiers with capacitive smoothing, the paper describes basic concept, stationary operation and dimensioning guidelines, design of the control and the transient behavior of the proposed system. Finally, measurements taken from a laboratory model are presented.

1 Introduction

The widely applied three-phase diode bridge rectifier (with a single smoothing inductor located at the DC side as shown in **Fig.1** (a) or with three inductors at the AC side) provides a very cost-efficient and reliable solution for feeding the DC link of power electronic converters (e.g., for drive systems or switch-mode power supplies) from the AC mains. The essential drawback of this system, however, is the significant harmonic stress on the AC mains caused by the largely non-sinusoidal line current shape (cf. Fig.1 (b)) which results (despite of the relatively good displacement factor $\cos\varphi_1$ due to the operation without phase control) in a poor total power factor λ. Furthermore, the current harmonics also lead to a significant ripple current stress on the DC link capacitor located at the output side of the rectifier. As shown in **Fig.2** both drawbacks mentioned before can be reduced by

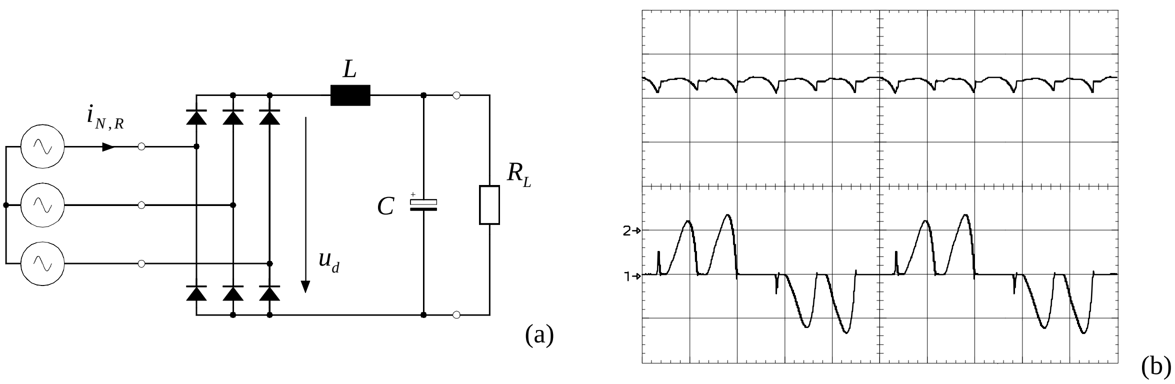


Fig.1: Mains current and output voltage time-behavior of a three-phase diode bridge rectifier. (a): power circuit; (b): mains current $i_{N,R}$ (ch.1: 20A/div) and output voltage u_d (ch.2: 150V/div), measured power factor $\lambda = 0.75$; parameters: mains voltage 400V (rms, line-to-line), $L \approx 1.5$mH, $P \approx 5$kW, 2 ms/div.

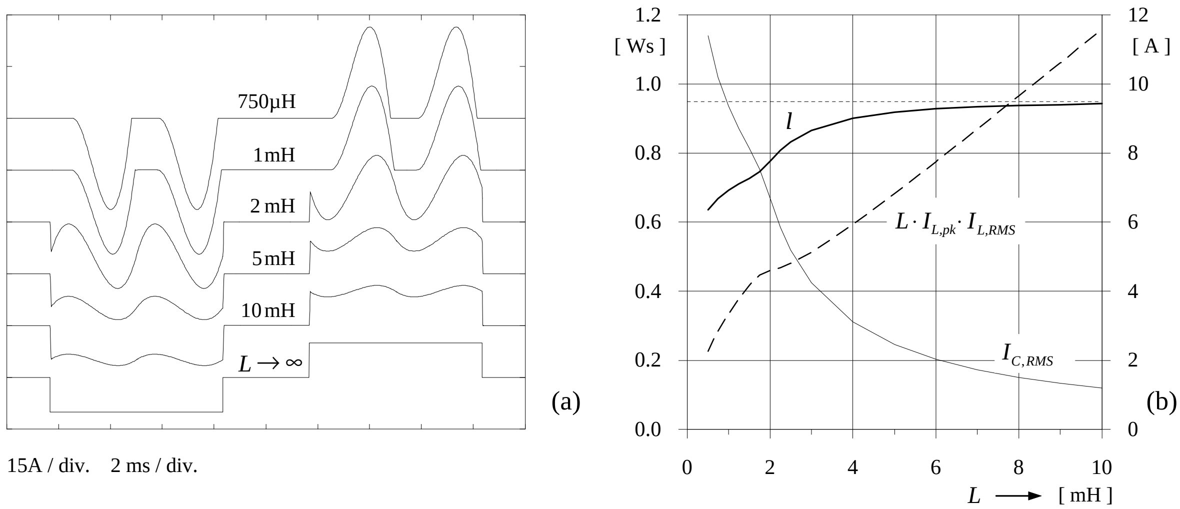


Fig.2: (a): Mains current shapes for different values of the smoothing inductance L; (b): total power factor l and current stress of the smoothing capacitor in dependency on L; furthermore, the product $L \cdot I_{L,pk} \cdot I_{L,RMS}$ which corresponds to the rated power (size) of the smoothing inductor is shown by the dashed curve; parameters: mains voltage 400V (rms, line-to-line), $P \approx 5$kW.

increasing the smoothing inductance L (for $L \to \infty$, i.e. pure 120° rectangular current shape, the total power factor becomes $\lambda = \frac{3}{\pi} = 0.955$ and $I_{C,rms}$ decreases to zero). However, this would significantly impair the power density of the rectifier due to the bulky smoothing inductor. Furthermore, the increased L also worsens the dynamic behavior of the DC link voltage because the characteristic impedance $Z^2 = L/C$ is increased as compared to the load resistance R_L which reduces the damping of the LC smoothing filter. This problem is additionally intensified in case of the supply of non-linear loads, e.g., constant power loads showing a negative small-signal input impedance.

2 "Active Smoothing" – Basic Operating Principle

The basic approach of the system presented in this paper is the replacement of the passive smoothing inductor L by a small power electronic unit (cf. **Fig.3**) which in the stationary case compensates the voltage ripple u_{AC} of the diode bridge and guarantees a well damped dynamic behavior by proper control. Ideally, this system acts as a pure current sink to the diode bridge as this is also the case if a step-up DC/DC converter is connected in series to the output of the B6 circuit as has been proposed in [1] (cf. **Fig.4**). As shown in [2] such a system is characterized by a very high utilization of the active and passive components as compared to other active three-phase rectifier circuits. Nevertheless the step-up converter has to be designed concerning the total output (DC link) voltage and the total output current. Therefore, for rectifier systems connected to the 400V three-phase mains

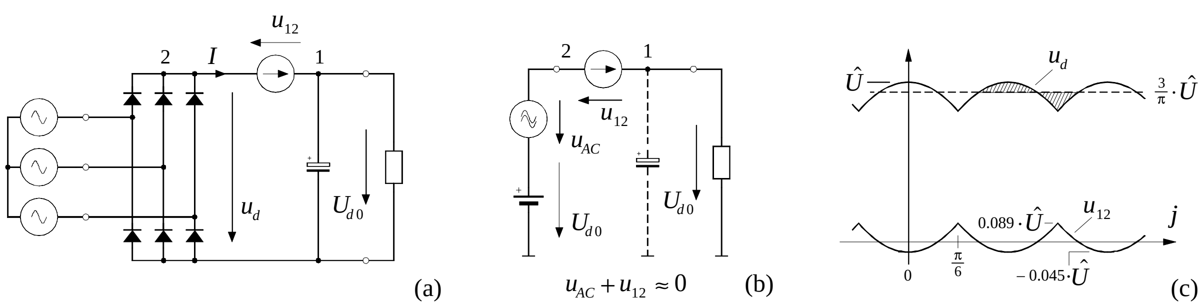


Fig.3: Replacement of the smoothing inductor by an active system u_{12} which compensates the output voltage ripple u_{AC} of the diode bridge rectifier.

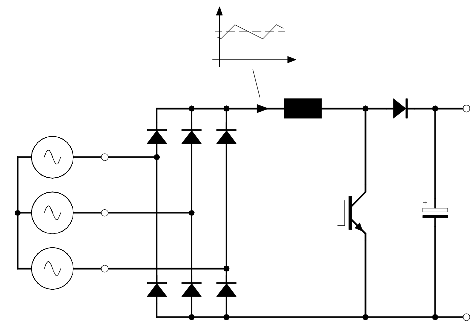


Fig.4: Active rectifier based on a series connection of a conventional three-phase diode bridge with a step-up converter located at the DC side [1].

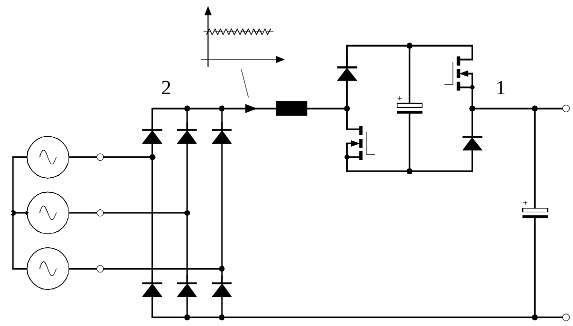


Fig.5: Active smoothing according to Fig.3 by application of a small switch-mode converter emulating a large smoothing inductor.

(rms line-to-line voltage) with resulting DC link voltage levels of > 600V turn-off power semiconductor devices with a blocking capability of $\geq$1000V (IGBTs) have to be used. Due to the limited switching frequency of such devices (< 10...25kHz) there remains a substantial filtering effort.

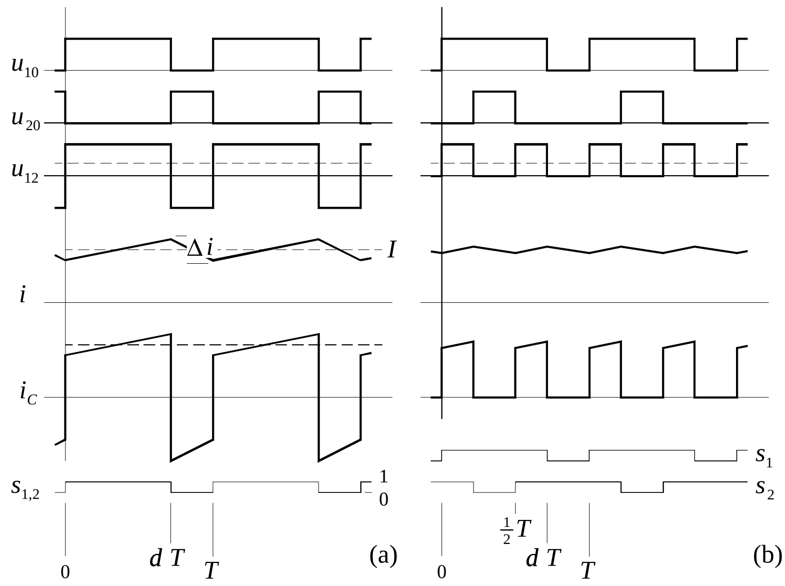
On the contrary the power circuit of the active smoothing system proposed here according to Fig.3 (realized by the switch-mode topology of **Fig.5**) only has to be designed regarding the voltage ripple of the B6 rectifier (i.e., ≈10% of the total DC output voltage). Therefore, majority carrier based power semiconductors (modern low-voltage high-current MOSFETs and Schottky diodes) are applicable which allow switching frequencies $f_S = \frac{1}{T} >$ 100kHz and lead to a significantly improved power density. The voltage u_{12} is generated by a switch-mode topology which results from the well-known 4Q-H-bridge where two transistors and two free-wheeling diodes can be omitted due to the unidirectional current flow given by the rectifier diodes. In the ideal case there would be a power flow across the switch-mode stage showing a frequency of $6 f_N$ (f_N ... mains frequency) with a zero average value. This results from the ideally pure DC current $i = I$ (for $f_S \rightarrow \infty$) in connection with the pure AC voltage u_{12}. Therefore, neglecting the losses of the power semiconductors, the DC link voltage U_C of the switch-mode stage could be provided by a capacitor and no auxiliary power supply would be necessary. This can be compared to a passive smoothing inductor which is – if its losses are neglected – also a pure energy storage element. The proposed active system realizes this energy storage behavior by "transformation" into the DC link capacitor of the switch-mode stage. However, the energy storage by electrolytic capacitors (usually applied in DC voltage link systems) shows a very high specific energy density as compared to the magnetic energy storage capability of coils, which leads to a high power density of the active smoothing system.

3 Stationary Operation – Dimensioning Guidelines

In the following the voltage and current stresses on the active and passive components of the switching stage shall be calculated in order to give guidelines for the dimensioning of the system. First, the DC link voltage U_C has to be chosen. According to Fig.3 (c) we get $U_C \geq 0.089 \cdot \hat{U} = \hat{U} \cdot (3/p - \sqrt{3}/2)$ for guaranteeing controllability ($\hat{U}$ ... peak value of the diode bridge output voltage). Considering a sufficient voltage margin for a practical realization U_C is set to about 10...15% of $\hat{U}$, i.e. 60...90V for applications at a 400V mains ($\hat{U}$ = 566V). The voltage across the active smoothing stage

$$u_{12} = U_{d0} - u_d = U_{d0} - u_d = \tfrac{3}{\pi}\hat{U} - \hat{U}\cos(\omega_N t) \tag{1}$$

(valid for the angle (time) interval $-\frac{\pi}{6} \ldots +\frac{\pi}{6}$ with $U_{d0} = u_{d,avg} = \frac{3}{\pi}\hat{U}$ and ω_N ... mains angular frequency) has to be generated as local average value (averaging within the switching period T) by the switching stage via pulse width modulation with the duty cycle $\delta = 0 \ldots 1$ (cf. **Fig.6**):



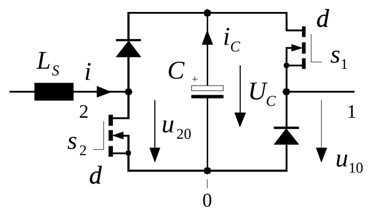


Fig.6: Voltage and current wave shapes of the switch-mode stage for (a) two-level control and (b) three-level control (control signals s_1 and s_2 phase-shifted by $T/2$).

$$u_{10} = U_C \cdot \delta \qquad u_{20} = U_C \cdot [1-\delta] \qquad u_{12} = u_{10} - u_{20} = U_C \cdot (2\delta - 1)\,. \tag{2}$$

The combination of Eqs.(1) and (2) gives the duty cycle for the transistor δ and $1-\delta$ for the diode

$$\delta(\omega_N t) = \frac{1}{2} - \frac{1}{2}\frac{\hat{U}}{U_C}\left[\cos(\omega_N t) - \frac{3}{\pi}\right] \qquad 1-\delta(\omega_N t) = \frac{1}{2} + \frac{1}{2}\frac{\hat{U}}{U_C}\left[\cos(\omega_N t) - \frac{3}{\pi}\right] \tag{3}$$

which are required as weighting functions for the calculation of the avg- and rms-values of the transistor current

$$I_{T,avg} = \frac{3}{\pi}\int_{-\pi/6}^{+\pi/6} I \cdot \delta(\omega_N t)\, d\omega_N t = \frac{1}{2} I \qquad I_{T,rms} = \sqrt{\frac{3}{\pi}\int_{-\pi/6}^{+\pi/6} I \cdot \delta(\omega_N t)\, d\omega_N t} = \frac{1}{\sqrt{2}} I\,. \tag{4}$$

These equations show that the current stress is independent of the ratio $U_C/\hat{U}$ which can be explained by the fact that due to the pure AC characteristic of u_{12} the average duty ratio is equal to $\frac{1}{2}$. Therefore, the current stress of the diodes becomes equal to the values of the transistors (omitting the detailed calculation here): $I_{D,avg} = I_{T,avg}$ and $I_{D,rms} = I_{T,rms}$. The ripple in the smoothing inductor L_S and the current stress on the DC link capacitor C depends on the operating mode of the pulse width modulator. If both MOSFETs are gated by the same control signal $s_1 = s_2$ the voltage u_{12} shows the values $+U_C$ and $-U_C$ (2-level control, cf. **Fig.6** (a)), the capacitor current results to $I_{C,rms} = I$ and is not dependent on $U_C/\hat{U}$ (contribution of current ripple Δi is neglected; cf. i and i_C in Fig.6 (a)). It is preferable, however, to shift the control signals by $\frac{1}{2}T$ to get 3-level control (Fig.6 (b) shown for $\delta > \frac{1}{2}$). There, due to the tighter voltage approximation ($+U_C$, 0, $-U_C$) and the doubling of the effective switching frequency the maximum current ripple ΔI is reduced to $\frac{1}{4}$ as compared to 2-level control:

$$\text{2-level control:} \quad \Delta I = \frac{U_C}{2 L_S f_S} \quad \text{at} \quad \delta = \frac{1}{2} \qquad \text{3-level control:} \quad \Delta I = \frac{U_C}{8 L_S f_S} \quad \text{at } \delta = \frac{1}{4}, \frac{3}{4}\,. \tag{5}$$

Furthermore, also the current stress on C is reduced, but it now depends on the ratio $U_C/\hat{U}$. Omitting the detailed calculation, I_C results to

$$I_{C,rms} = I \cdot \sqrt{\frac{\hat{U}}{U_C}} \cdot \sqrt{\tfrac{36}{\pi^2}\left[\sqrt{\tfrac{\pi^2}{9} - 1} - \arctan\sqrt{\tfrac{\pi^2}{9} - 1}\right]} = 0.186 \cdot I \cdot \sqrt{\frac{\hat{U}}{U_C}}\,. \tag{4}$$

4 Loss Compensation – Control Concept

A calculation of the losses based on the relationships given in the previous section results for the switch-mode stage on total in less than 1% of the rectifier output power. Without any counter-measures these losses would have to be covered by an auxiliary power supply feeding the DC link voltage U_C. However, by application of a proper control it is possible to cover the losses out of the power flow through the electronic smoothing stage. The control concept shown in **Fig.7** gives one possibility for a realization where a "virtual" resistor R located between 1–2 is emulated by the switch-mode stage. The power "dissipated" by R (resulting from the pure AC voltage u_{12}) is fed into the DC link and compensates the losses. The value of R has to be adapted by a voltage control loop for U_C which is not shown in Fig.7. Alternatively, it would also be possible to achieve loss compensation by a small DC component of u_{12}.

The basic current control of the system shown in Fig.7 is realized by a simple hysteresis controller with the reference value $i^{*} = i_O + u_{21}/R$ (output current pre-control) which leads to a first order response of u_{12} (DC components of u_{12} will be suppressed according to the time constant RC_O). The simulation results given in **Fig.8** demonstrate the time behavior of a system for the following parameters: $U_{d0} \approx 540\text{V}$ (400V mains), $I = 10\text{A}$ ($P \approx$ 5.4kW), $\Delta I = 2\text{A}$, $L_S = 200\mu\text{H}$, $U_C = 70\text{V}$, $f_{S,\max} \approx 100\text{kHz}$ for different values of R (20Ω for the first fundamental period, 200Ω for the second period). The current $I_{C,avg}$ (averaging within a 60°-interval of the fundamental period) shows the resulting power flow into the DC link capacitor C for compensating the losses.

Basically, Fig.7 would also suggest to measure the current through the output capacitor C_O instead of measuring i and i_O. However, the drawback of a "perfect" pre-control by i_O is that it counteracts the physical reason for providing C_O, i.e., the energy storage capability. If, e.g., the rectifier feeds a PWM drive, i_O has rectangular shape with the pulse frequency of the drive inverter. In this case a proper low-pass filter in the i_O-path of Fig.7 is required that i is controlled to the local (pulse frequency) average value $i_{O,avg}$.

The essential drawback of the hysteresis current control, the 2-level behavior as described in the section before, can be avoided by the application of a fixed frequency, phase shifted pulse width modulator-based current controller. The much lower current ripple ΔI (cf. Fig.6) showing twice the frequency f_S will significantly ease the dimensioning of the EMI filter located at the mains terminals or at the output of the diode bridge, respectively (not shown in Fig.7). (Remark: The transition from 2-level control to 3-level control will reduce the remaining voltage ripple at the EMI filter capacitor by a factor of 8.) For a practical realization the phase shifted gate control signals s_1 and s_2 will be generated, e.g., according to **Fig.9** (a). Due to the pre-control by i_O the current control can be performed with the P-type controller R instead of the switching (variable frequency) hysteresis controller in case of the 2-level control. A possibility for the loss compensation (i.e., the control of the DC link voltage U_C) is the addition of an offset reference current I_P (defined as output signal of the U_C-controller) as

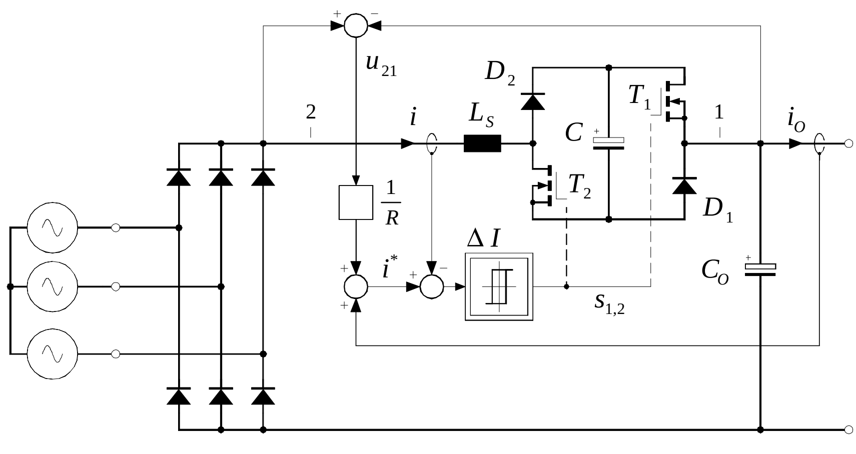


Fig.7: Basic control concept (2-level hysteresis current controller) with output current pre-control and loss compensation (feedback path u_{21}/R).

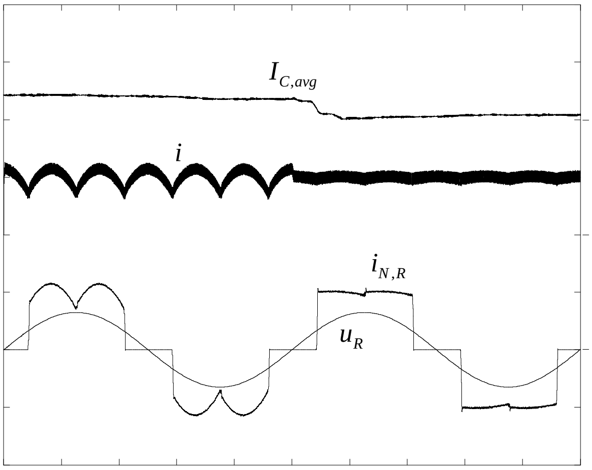


Fig.8: Simulation of the active smoothing concept (2-level hysteresis current control according to Fig.7) for different gain values R (see text); 500V/div, 10A/div ($i_{N,R}$), 2A/div ($I_{C,avg}$), 4ms/div.

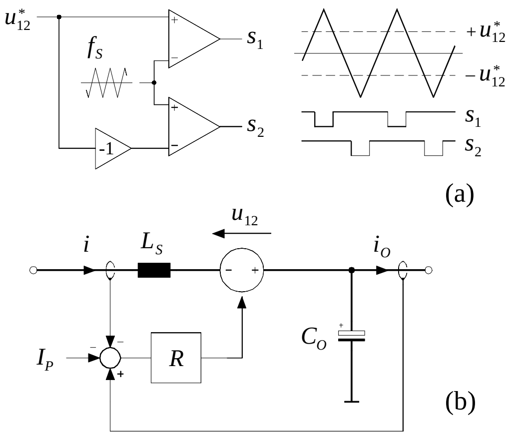


Fig.9: Generation of control signals for 3-level PWM control based on phase shifted pulse width modulators (a) and control oriented equivalent circuit diagram (b).

shown in the control oriented equivalent circuit diagram Fig.9 (b). This results in a small DC component $I_P \cdot R$ of the voltage u_{12} which leads to a stationary power flow of $P = I \cdot I_P \cdot R$ compensating the losses of the switch-mode stage. Figure **10** presents the diode bridge input and output currents taken from a laboratory model of an active smoothing rectifier with loss compensation (cf. first fundamental period of Fig.8). The measured total power factor $\lambda = 0.95$ comes close to the theoretically maximum of $\frac{3}{\pi} = 0.955$.

Finally, the start-up of the system should be discussed briefly. As is the case of conventional diode bridge rectifiers, also the proposed concept requires a resistor to limit the start-up inrush currents at $t = t_0$ (**Fig.11**). However, now there exists a series connection of the main DC link capacitor C_O, the DC link capacitor C and the diodes $D_{1,2}$ of the switch-mode stage (cf. Fig.7). Due to its higher capacity, a lower voltage appears across C but usually would rise over the nominal value U_C. For avoiding this, the PWM stage is turned on with 1:1 duty ratio at $t = t_1$ when U_C reaches its nominal value. Because of $u_{12,avg} = 0$ the charging of C is stopped and C_O is charged to the value U_{d0}; the control loops of the active smoothing can be activated subsequently.

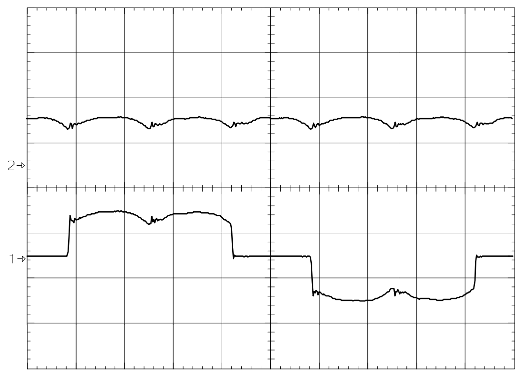


Fig.10: Mains current $i_{N,R}$ (ch.1) and diode bridge output current i (ch.2) taken from a laboratory model of an active smoothing rectifier, power factor $\lambda = 0.95$; parameters: mains rms voltage 220V (line-to-line), 5A/div, 2ms/ div.

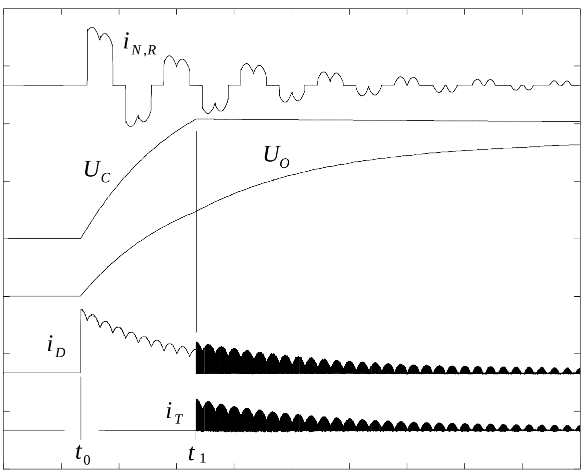


Fig.11: Simulation of the start-up (see text). Scales: 15A/ div; 200V/div (U_O), 30V/div (U_C), 15ms/div.

6 Conclusions – Future Developments

The proposed concept improves the mains behavior of three-phase diode bridge rectifiers based on the replacement of the passive smoothing inductor by an active switch-mode stage. The rated dimensioning power of this switch-mode converter, which can be realized by application of modern high-current low-voltage MOSFET devices, is relatively small (≈10%) as compared to the total rectifier output power. The applicable semiconductor technology leads to a system of very high total power density and efficiency. Furthermore, besides the improvement of the total power factor at the mains, the active smoothing also results in a lower current stress of the output (DC link) capacitor. The losses of the switch-mode stage (which are < 0.5 ... 1% of the total output power) can be compensated by proper control. Due to this, only a very small (or even actually no) auxiliary power supply would be necessary.

The basic idea of the described active smoothing is the compensation of the relatively small output voltage ripple of a three-phase diode bridge rectifier by a low-voltage switch-mode stage. As the example of **Fig.12** shows, this concept also can be applied advantageously for three-phase switch-mode power supplies. With this topology, an additional advantage appears due to the fact that the ripple compensation stage ("auxiliary converter", consisting of T and D) also acts as a step-down converter. Therefore, the primary isolation stage ("main converter" $T_1 \ldots T_4$) can be equipped with transistors with lower blocking capability. For systems connected to the 400V mains ($U_{d0} \approx 540$ V) it would be possible to use modern 600V MOSFET devices (CoolMOS etc.) for the main converter (e.g., $U_P \approx 450$ V, $U_A \approx 250$ V). In the case at hand the output voltage control is performed via the auxiliary converter. The main converter, which is current-fed by the auxiliary stage, is operated at full (1:1) duty cycle, L_S acts as (reflected) output smoothing inductor. Contrary to the active smoothing concept analyzed in the previous sections, here the "low-voltage switch-mode stage" (i.e., the auxiliary converter) has non-zero average power due to the step-down behavior ($U_P < U_{d0}$). Depending on the input voltage level and ripple amplitude the power delivered by the auxiliary converter is about 10...20% of the power of the main converter. It should be noted that there exists a specific ratio of the main and auxiliary converter power levels defined by the voltage relationship (the output current of the diode bridge rectifier feeds both converters). Therefore, it is hardly possible to feed independent loads by the two converters. A detailed analysis of this switch-mode power supply topology, which seems to be advantageous if the input and output voltages are specified within a relatively narrow region, will be presented in future paper.

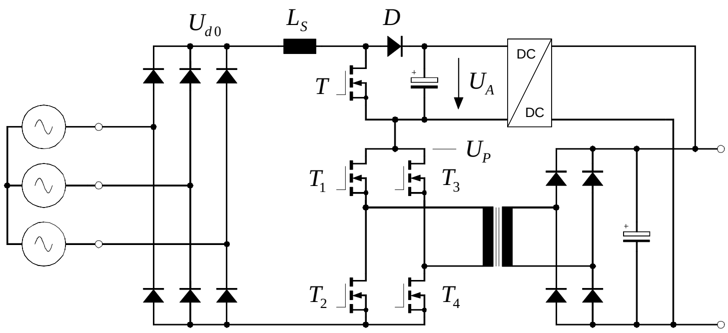


Fig.12: Three-phase switch-mode power supply based on the active smoothing concept.

References

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