

Status of the Techniques of Three-Phase Rectifier Systems with Low Effects on the Mains

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Abstract Based on a comprehensive study of the literature concepts of three-phase rectifier systems with low effects on the mains are classified. Such systems are unidirectional and bidirectional self-commutated converters with impressed output voltage or output current and line-commutated rectification in connection with passive and active filtering. Selected circuit concepts are analyzed concerning the operational behavior and the obtainable quality of the mains current. Furthermore, an evaluation of the rectifier concepts concerning utilization of the power semiconductors, rated power of the inductive and capacitive components and of the realization effort in general is given. Finally, problems of a practical application of high switching frequency PWM rectifier systems and topics of further research are discussed.

1 Introduction

Due to guidelines, recommendations [1] and standards (IEEE-519 and IEC-555, [2]) giving a limit to strive for concerning the harmonic stress on the mains by power electronic systems, converter concepts with low effects on the mains gain more and more importance [3].

The technical conditions for developing of self-commutated PWM rectifiers with low effects on the mains have been established by the availability of turn-off power semiconductors with low switching power loss (power MOSFETs, IGBTs and GTOs) and of digital control systems (modern micro controllers or digital signal processors) with low cycle times and/or high dynamic quality.

Power converter systems have highly varying topology and complexity, depending on the application area and the required operational behavior. The basic requirements being determining for the selection of the circuit concept of a PWM rectifier from a technical point of view can be defined as, e.g.,

- approximately sinusoidal current consumption,
- resistive fundamental mains behavior,
- fixed reference value control of the output voltage,
- high power density.

Besides reducing the effects on the mains, this also gives (as compared to conventional line-commutated rectifier systems) the possibility of a highly dynamic control of the power flow, the generation of an output voltage being constant independently of the mains voltage and the load. Furthermore, a reduction of the rated power and the weight of the rectifier unit can be achieved. There, the constant output

voltage feature also gives the possibility of an adjustment to a wide input voltage region and a maximum utilization of the rated power of a converter connected in series on the DC side for two-stage rectifier systems.

The aim of this paper is to give a classification of the circuit concepts of three-phase PWM rectifier systems as given in literature and to give an overview over the technical possibilities and the most important trends of the presently very dynamic developments in the area of self-commutated rectifier systems. Also a comparative evaluation of selected circuit concepts will be given. The main attention will be paid to PWM converters (with low and medium power) to be operated from the public low-voltage mains. This means that area which seems to be most likely affected (due to its high application diversity) by future stricter standards concerning effects on the mains of power converter systems. Finally, we will point out trends of further technical developments of three-phase rectifier circuits and problems which can be expected in connection with extended application of this type of rectifier systems.

2 Classification of Three-Phase Rectifier Systems with Low Effects on the Mains

Figure 1 shows a classification of three-phase rectifier systems for low and medium power. Based on the number of pertinent publications the main focus of the development is now in the area of self-commutated and hybrid circuit concepts. (The latter denotes the series and/or parallel connection of a line-commutated and a self-commutated converter.) In connection with hybrid systems we also have to point out the future high importance of active filters (besides the concepts discussed here) in the area of energy distribution. A more detailed discussion of this topic has to be omitted however, for the sake of brevity. Due to the much broader application of pulse rectifier systems with impressed output voltage (voltage DC link pulse rectifier systems) as compared to current DC link pulse converters (e.g., for electrical drive systems) also current DC link pulse rectifiers are omitted from the considerations here.

For voltage DC link pulse rectifiers we have to distinguish between unidirectional circuit concepts (limited to rectifier operation) and bidirectional circuit concepts (rectifier and inverter function). Bidirectional two-level, three-level and multi-level converters are known from the area of high-power traction [4]. There the control concepts used for inverter operation and/or energy supply for the motors can basically

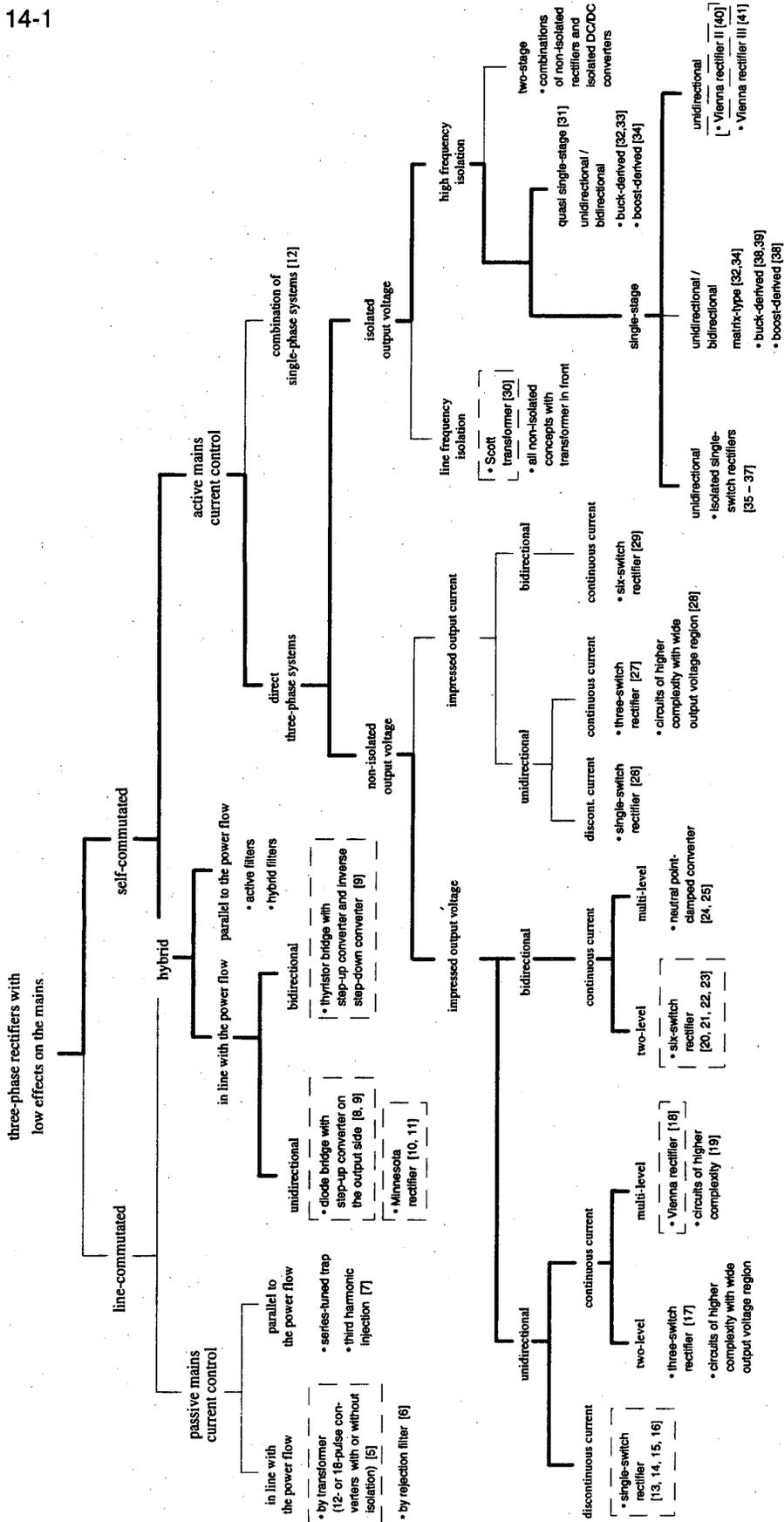


Fig. 1: Classification of three-phase rectifier circuits with low effects on the mains. The circuit concepts discussed in greater detail in this paper are denoted by hatched frames. For the other groups references are given. The primary trends of development are marked by thicker connecting lines.

also be used for the rectifier circuits. There, the minimization of the ripple of the motor torque corresponds to the requirement of a limitation of low-frequency mains current harmonics and/or the minimization of the rated power of filter circuits possibly required.

The rectifier circuits can be designed in unidirectional version, if according to the specific application, no energy feedback from the DC side into the mains is required. This is the case, e.g., in drive systems with low dynamics (fan drives, air conditioners, etc.), UPS (uninterruptable power supplies, e.g., for telecommunications and medical electronics), battery chargers (electric vehicles), electric power supply for process technology (welding units, laser and plasma power supplies) and power supply for higher power switch mode amplifiers (measurement and test equipment). This makes possible a significant reduction of the circuit complexity as compared to bidirectional systems and has led to the derivation of a multitude of different circuit concepts.

In section 3 the basic function and the mains behavior of in each case one example of the main groups of unidirectional PWM rectifier systems are treated. There, we distinguish concerning the time behavior of the input current, i.e., cases with continuous and discontinuous current. Line-commutated converters with low effects on the mains (with a pulse number of 12 or 18) are preferred for feeding of energy to systems of high power and where high reliability (and high efficiency) are required, as well as little emphasis is put on power density and mains current quality and if no highly dynamic control of the output voltage is required. If no isolation between AC and DC sides is required, the high pulse number (e.g., 12 or 18) can be achieved by a combination of two or three partial rectifier systems and phase-shifting auto-transformer circuits in connection with interphase reactors on the DC side or by application of inductances connected in series on the mains side and with a three-phase interphase transformer on the mains side. The basic function of an uncontrolled twelve-pulse (diode) rectifier system with three-phase interphase transformer and capacitive smoothing will be described in section 3.7.

Basically, by connecting a mains frequency transformer in series at the input a galvanic isolation of the output voltage can be achieved for each PWM rectifier system. However, such systems show a low power density due to the high volume and the high weight of a low frequency transformer. A very elegant realization of a mains-frequency isolated system has been proposed in [30] (cf. section 3.8.1) employing two boost power factor correctors with common DC output voltage on the secondary.

rectifier systems with high-frequency isolation of the output circuit in general show a two-stage structure. There the partial tasks (i) rectification and (ii) shaping of the input current are realized by the input stage and the tasks (iii) high-frequency isolation, (iv) matching of the input and output voltage levels and (v) tight, highly dynamic control of the output voltage are handled by a DC/DC converter output stage. Alternatively, the power conversion also can be achieved without DC link energy storage resulting in quasi-single-stage energy conversion [31–34] or by single-stage topologies according to [35–41]. For single-stage converter systems all partial tasks mentioned are realized by a single converter stage. Advantages are a lower realization effort and a potential for higher efficiency. Drawbacks are a higher complexity of the system control, especially for guaranteeing a symmetrical magnetization of the high-frequency

transformer which is an integral part of the converter topology as well as a higher voltage and current stress on the valves (as compared to two-stage power conversion) for wide input or output voltage range. In order to show the basic principle of operation of a single-stage system a single-stage boost-type (due to inductors connected in series on the mains side) PWM rectifier system which has been introduced in the literature as *VIENNA Rectifier II* [40] will be discussed briefly in section 3.8.2.

3 Circuit Concepts, Operation and Mains Behavior

Before we go into the details of the different circuit concepts we want to briefly present the functional principles of the pulse rectifier circuits in the following. As already mentioned here, we have to distinguish basically between systems with continuous and with discontinuous input current shape.

For systems with continuous input current shape one can describe the current generation on the AC side by the equivalent circuit shown in Fig.2(a). There, the phase quantities are summarized by space vectors \underline{u}_N (mains voltage space vector) and \underline{u}_U (space vector of the rectifier input voltage). The space vector \underline{u}_U^* being formed in the average over a pulse period at the input of the rectifier system is set by a vector control quantity \underline{u} regarding phase and value (relative to the mains voltage) in such a way that the voltage drop $\underline{u}_N - \underline{u}_U^*$ across the series inductors L on the mains side impresses the current i_N which is required by an output voltage control loop. The actual rectifier input voltage \underline{u}_U contains (besides its fundamental \underline{U}_U^*) harmonics with switching frequency which result in corresponding mains current harmonics. That means that the low frequency current harmonics being characteristic for conventional line-commutated rectifier circuits are shifted to higher frequency regions by the pulse width modulation. With this the advantage of reducing the effects on the mains is opposed by the danger of the occurrence of EMI (cf. VDE 0871, frequency range: 10 kHz... 30 MHz). The equivalent circuit according to Fig.2 remains basically valid also for the circuits according to Figs.9–11. However, the possibility of influencing the voltage generation by a control input is lacking if the circuit shown in Fig.9 is realized without turn-off power semiconductors.

A noticeable deviation of the basic circuit function from the previous considerations is only given for the circuits according to Figs.3 and 5. There, the current flow is impressed by a DC/DC converter at the output of a three-phase diode bridge. Therefore, a block current shape results as being characteristic for uncontrolled three-phase bridge circuits with smoothing inductor on the DC side. A substantial reduction of the low-frequency mains current harmonics (occurring in this case) can be obtained by the circuit shown in Fig.4 by feedback of a current harmonic with three times the mains frequency to the AC terminals of the rectifier bridge.

The circuits corresponding to the group *Voltage DC link Power Converters with Discontinuous Input Current* according to the classification of Fig.1 have typically only one turn-off valve. The peak values of the rectifier input phase currents occurring within the pulse periods are guided there directly by the corresponding mains phase voltages along sinusoidal envelopes. There, the amplitude of the envelope (and,

therefore, the power flow to the DC side) can be defined by a control input. Rectifier systems of this group have intrinsic resistive fundamental mains behavior [42]. Based on the direct impression of the input current by the input voltage the control part of Fig.2(a) has to be replaced concerning its basic function by a current source which is influenced by the mains input voltage and a scalar control quantity s (turn-on time of the power transistor) as shown in Fig.2(b). In the

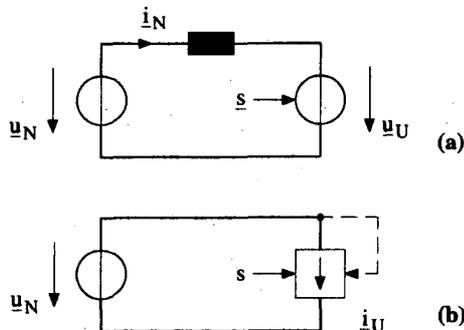


Fig.2: Space vector equivalent circuit of the system part on the AC side of three-phase rectifier systems with continuous (cf. (a)) and discontinuous (cf. (b)) shape of the input current.

following we want to treat briefly the principle of operation and the mains behavior of selected rectifier circuits which are characteristic for the different groups of the classification according to Fig.1. The component stresses and the rated power of the passive components (which essentially determine the dimensioning and possible application area) and/or a comparative evaluation of the systems is given in section 4. Regarding a detailed description of the systems we would like to refer to the references compiled in section 6.

3.1 Three-phase Diode Bridge with Step-Up Converter Connected in Series

The basic thought of the circuit [8] shown in Fig.3 is to replace the smoothing reactor on the DC side of a conventional uncontrolled rectifier circuit by an active circuit part, i.e. a step-up DC/DC converter. This results in a substantial reduction of the rated power of the smoothing inductance. Furthermore, the possibility of controlling the output DC voltage (lying above the peak value of the mains line-to-line voltage) is given. The mains behavior of the circuit is not essentially influenced as compared to purely passive realization. There, the amplitudes of low-frequency harmonics being characteristic for the current with block shape occur. However, the circuit being applied, e.g., for realizing the input stage of telecommunication power supply modules shows a relative high power factor and is characterized in general by a very high utilization of the power components (cf. Tab.1).

3.2 Third-Harmonic Modulated (Minnesota) Rectifier

A substantial reduction of the effects on the mains of the circuit shown in Fig.3 can be achieved by splitting up the

circuit part on the DC side and insertion of a zigzag auto-transformer according to Fig.4 [10]. There, the output current of the diode bridge is modulated with three times the mains frequency. The current component formed in such a manner is fed back to the inputs of the diode bridge via the transformer circuit (based on a basic idea given in [7]) representing a low impedance for zero-sequence quantities. The injection device (transformer) also could be replaced by a wye-delta transformer of higher rated power [11] or a filter circuit tuned to three times the mains frequency [43].

By controlling the magnitude and the phase of the third harmonic current the low-frequency mains current harmonics of the circuit (developed by Prof. Mohan at the University of Minnesota and introduced in the literature as *Minnesota Rectifier*) can be limited to values $\frac{1}{I_1} \hat{I}_3 < 0.03$ (cf. Tab.1). The concept is insensitive to line voltage unbalances and can be applied also for twelve-pulse arrangements (there, 9th harmonic currents are circulated, cf. Fig.4 in [44]) and inverter operation [45].

Remark: We would like to point out that according to [46] a low frequency distortion of the mains current can be avoided ideally by modifying the shape of the current fed back to the diode bridge inputs (cf. Fig.5 in [46]). A similar concept also has been analyzed in [47].

3.3 Three-Phase Thyristor Bridge with Series Step-Up and Inverse Step-Down Converter

By replacing the diode bridge by a thyristor bridge and insertion of a thyristor which inverts the output voltage polarity in connection with a blocking diode (cf. Fig.5) the circuit shown in Fig.3 can be extended to bidirectional power flow [9]. For energy feed back into the mains, the feedback thyristor is turned on and the thyristor bridge is operated in the inverter mode. The DC side circuit part is switched over to a step-down DC/DC converter structure thereby (where the output voltage acts in a polarity opposed to that of the rectifier operation); this makes possible a control of the power flow into the mains. As disadvantage of this circuit which is applied in the area of drives (regenerative braking) we have to mention the relatively low mains current quality especially in the inverter mode. As advantage we have to point out the cost which is lower than for conventional bidirectional (four-quadrant) rectifier systems (cf. Fig.8).

Concerning further combinations of mains commutated and a forced commutated converters for realizing a bidirectional AC/DC converter we would like to refer to Fig.1 in [48] for the sake of brevity. In case a high distortion of the mains current can be tolerated a realization of a bidirectional converter in principal can be achieved without using explicit inductive components as shown in [48]. The basic principle of operation of this system which has been introduced as *Loss-Free Braking Resistor* in the literature is comparable to a diode rectifier relying on the current limiting property of the inner mains impedance. The energy is fed back into the mains in a discontinuous manner with six times the mains frequency in the vicinity of the maxima of the mains line-to-line voltage.

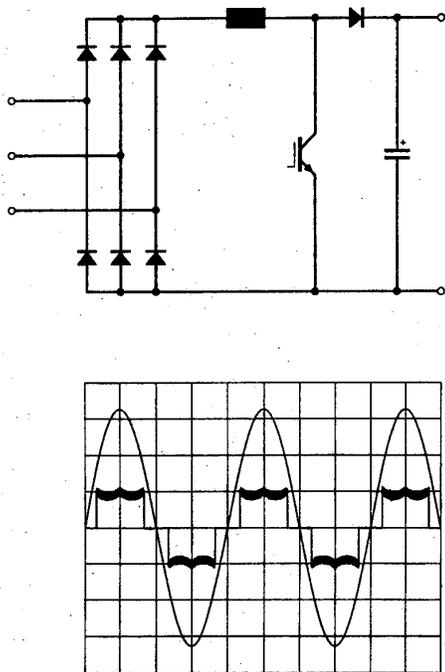


Fig.3: Basic structure of the power circuit and mains current shape of a unidirectional hybrid pulse rectifier system formed by combination of a three-phase diode bridge and a DC/DC step-up converter. Scales: 100V/div, 20A/div, 5ms/div.

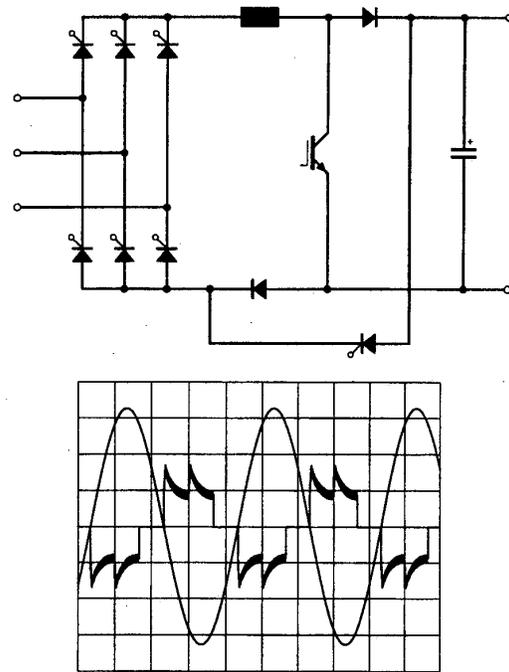


Fig.5: Basic structure of the power circuit and mains current shape (shown for energy feedback into the mains) of a bidirectional hybrid pulse rectifier system formed by combination of a three-phase thyristor bridge and a DC/DC step-up (for rectifier operation) / step-down (for energy feedback and/or inverter operation) converter. Scales: 100V/div, 20A/div, 5ms/div.

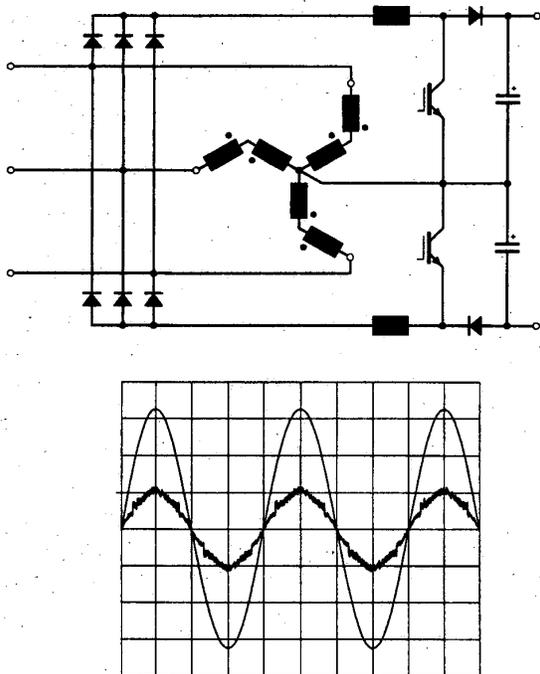


Fig.4: Basic structure of the power circuit and mains current shape of a three-phase third-harmonic modulated (*Minnesota*) rectifier (unidirectional hybrid pulse rectifier system with current impression by zigzag injection transformer). Scales: 100V/div, 20A/div, 5ms/div.

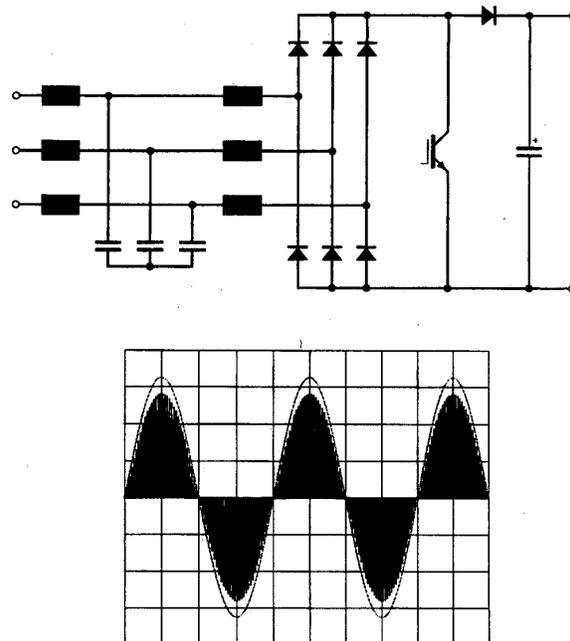


Fig.6: Basic structure of the power circuit and mains current shape of a single-switch discontinuous-mode boost rectifier (unidirectional voltage DC link pulse rectifier system) with discontinuous input current. Scales: 100V/div, 20A/div, 5ms/div.

3.4 Unidirectional Single-Switch Discontinuous-Mode Boost Rectifier

If the smoothing reactor of the circuit shown in Fig.3 is shifted to the AC side and split up to the three-phases there results the circuit concept shown in Fig.6 which is to be called unidirectional single-switch discontinuous-mode voltage DC link pulse rectifier system [13] due to the mains current shape and the commutation of the diode bridge with pulse frequency. By the power transistor the DC side of the diode bridge is shorted periodically with constant pulse frequency (lying substantially above the mains frequency). Thereby, in each phase at the end of an on-interval a current value is obtained which is proportional to the instantaneous value of the relevant phase voltage. (This means resistive fundamental mains behavior.) After turning off the power transistor there follows the demagnetization of the input inductors via the smoothing capacitor on the output side. The power flow is influenced by the length of the on-intervals being constant over the mains period in the simplest case.

The discontinuous input current of the circuit shows low-frequency spectral components [15] with relative amplitudes being dependent on the value of the voltage transfer ratio (output voltage related to the amplitude of the mains line-to-line voltage). A high mains current quality is connected to a high voltage transfer ratio and/or a high voltage stress on the valves. Further disadvantages of the circuit concept are the high peak current stress on the valves and a high filtering effort for guaranteeing electromagnetic compatibility.

Despite of these disadvantages this circuit is discussed presently as suitable for different industrial applications due to its low complexity and simple control. There the focus is on the reduction of the amplitudes of the low-frequency harmonics of the mains current by modulation of the turn-on time of the power transistor with six times the mains frequency (cf. [15] and 6th harmonic injection [49], [50]) and on the reduction of the transistor turn-off losses by soft-switching [?]. Furthermore, we would like to point out the reduction of the high amplitudes of switching frequency harmonics of the input current by staggered switching of two systems connected in parallel [51] as known from single-phase power factor correction [52], [53].

3.5 Three-Phase/Switch/Level PWM (Vienna) Rectifier

In the ideal case avoiding the occurrence of low-frequency mains current harmonics for unidirectional rectification can be achieved only by insertion of a turn-off semiconductor into each phase leg of a three-phase diode bridge. A circuit (developed at the Technical University Vienna, [18]) being especially advantageous concerning the stresses on the devices and the level of the current harmonics with switching frequency is shown in Fig.7. The control of the mains phase currents being sinusoidal in the average over a pulse period is achieved (as described in connection with Fig.2) by appropriate pulsing of the power transistors (pulse width modulation). When a transistor is turned on, the corresponding phase is connected to the output voltage center point, causing a rise of the associated phase current. Turning off the transistor leads to conduction of the associated diode in the upper or lower bridge half (depending on the direction of the current flow) and, therefore, to a reduction

of the phase current. As a result the possibility of a sinusoidal mains current control is given (in the phase angle region $\varphi = 30^\circ \text{cap.} \dots 90^\circ \text{ind.}$).

Due to the inclusion of the center point of the output voltage into the system function the bridge legs of the circuit show three-level characteristic. As opposed to two-level converter circuits, the harmonic level of the mains current is significantly reduced and/or input inductors of lower rated power can be used for equal rms value of the mains current harmonics with switching frequency. Furthermore, the switching losses are cut in half (all valves ideally are stressed by only half of the output voltage in blocking state and/or power MOSFETs with a blocking capability of only $V_{DSS} = 500\text{V}$ (characterized by low on-resistance) can be employed for realizing a system with 700V DC output voltage resulting in high efficiency of the system.

The relatively high assembly effort (which could possibly be seen as disadvantage) can be avoided in the lower power region (10...20kW) by application of a power module (IXYS VUM25-05 and IXYS VUM85-05A) containing one bridge leg of the circuit. The sensing effort required for phase current control can be minimized by reconstructing the instantaneous shape of the phase currents based on center point current measurement and integration of the differences of the mains phase voltages and the corresponding converter input voltages.

A low-frequency version (higher volume) version of the system has been proposed in [54], [55]. There, the switches are operated with six times the mains frequency. Due to low switching losses and the low transistor conduction times (each transistor is conducting only within 60° intervals following the zero crossings of the corresponding phase voltage) this circuit achieves an outstanding high efficiency but is characterized by relatively high amplitudes of low frequency mains current harmonics occurring for low-load condition.

Low switching losses also could be achieved by extending the operation of the high switching frequency circuit to soft-switching. A corresponding topology has been proposed in [56]. For the sake of brevity we, however, would like to refer to [38] for a detailed description of this advantageous system modification.

3.6 Four-Quadrant DC Link PWM Converter

There exist no differences concerning the basic function regarding the circuit shown in Fig.8 [20] as compared to the circuit concept given in Fig.7. However, the former circuit shows a higher realization effort (number of switches) and a two-level characteristic of the bridge legs resulting in higher voltage stress on the valves and higher amplitudes of harmonics with switching frequency of the voltage occurring at the input side of the bridge circuit. (The first harmonic of the rectifier input voltage with switching frequency has typically about a level of 0.3 of the output DC voltage). This disadvantage is opposed by the possibility of inverting the power flow, i.e. feeding of energy supplied by an inverter connected in series into the mains which is often required, especially in the drives area (e.g., for lowering the load in crane or elevator drives or for centrifugal drives and/or braking of large inertias).

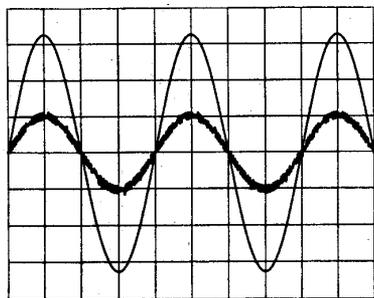
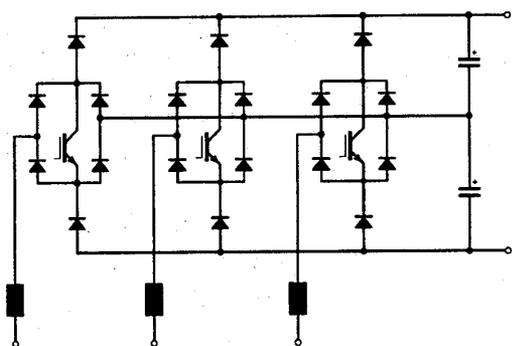


Fig.7: Basic structure of the power circuit and mains current shape of a unidirectional three-level pulse rectifier system (VIENNA Rectifier). Scales: 100V/div, 20A/div, 5ms/div.

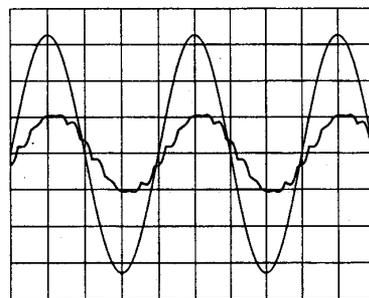
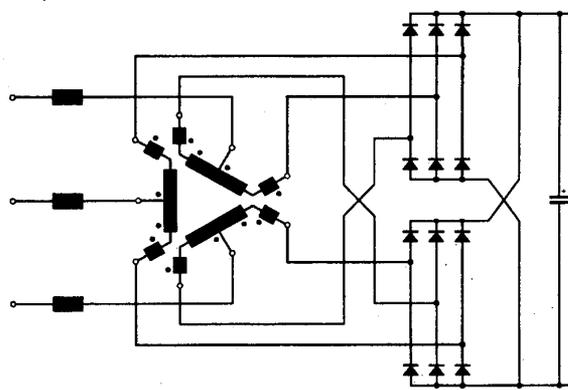


Fig.9: Basic structure of the power circuit and mains current shape of a unidirectional line-commutated 12-pulse rectifier system with interphase transformer on the mains side. Scales: 100V/div, 20A/div, 5ms/div.

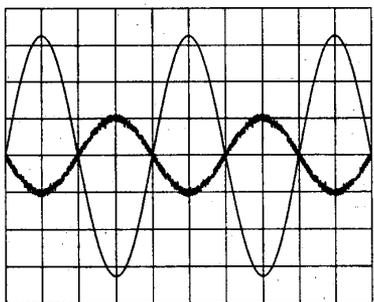
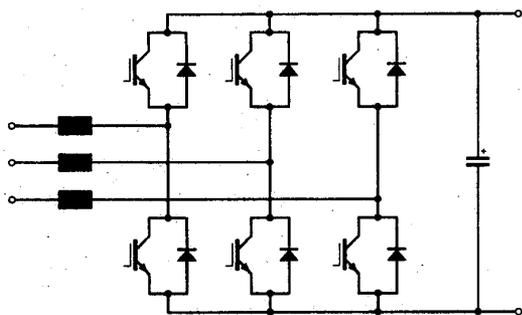


Fig.8: Basic structure of the power circuit and mains current shape (for inverter operation) of a four-quadrant voltage DC link pulse rectifier system (bidirectional six-switch voltage source rectifier). Scales: 100V/div, 20A/div, 5ms/div.

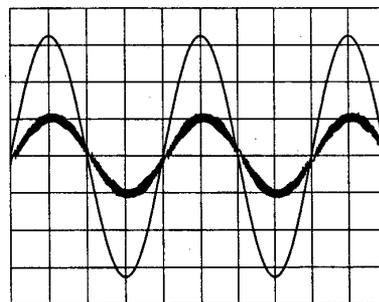
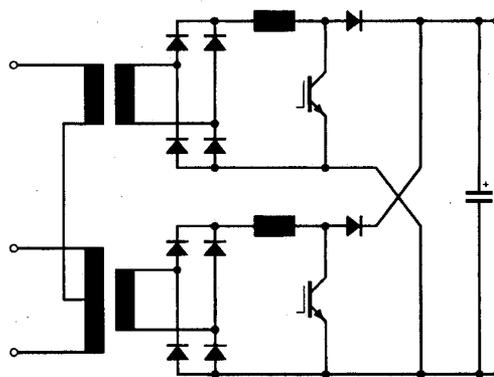


Fig.10: Basic structure of the power circuit and mains current shape of a mains-frequency isolated Scott-transformer-based unidirectional two-switch pulse rectifier system. Scales: 100V/div, 20A/div, 5ms/div.

Bidirectional three-level PWM rectifier systems corresponding to Fig.7 concerning the number of the levels used for voltage generation are presently applied only in the high power region due to the high realization effort [4].

For controlling bidirectional three-phase PWM rectifier systems discontinuous modulation [57] (two-phase modulation) can be applied advantageously. There, in a cyclic changing manner always one phase is clamped to the positive or negative output voltage rail within $\frac{\pi}{3}$ or $\frac{\pi}{6}$ -wide intervals of the fundamental period. Due to the missing connection between output voltage center point and mains voltage star point the mains current can be shaped sinusoidally by pulsing of always only two bridge legs. For two-phase modulation and ohmic fundamental mains behavior the effective switching frequency can be doubled as compared to continuous modulation for equal thermal stress on the power transistors. This allows a significant reduction of the rated power of the inductors connected in series on the mains side required for reaching a given harmonic level of the input current.

3.7 Twelve-Pulse Rectifier with Interphase Transformer on the Mains Side

If (as opposed to conventional six-pulse rectifiers) only a reduction of the effects on the mains and no controllability of the output voltage (or high power density) is required, the circuit shown in Fig.9 [5] offers an interesting realization possibility of a unidirectional rectifier system. The control of the voltage formation at the input of the three-phase diode bridge (which is realized for self-commutated pulse rectifier systems by turn-off power semiconductors) is achieved here by an interphase (auto-)transformer on the mains side. It transforms the six-pulse AC voltage systems at the AC sides of two three-phase diode bridges connected in parallel on the DC side into a twelve-pulse AC system. The interphase transformer has to be realized using three single-phase cores or a five-limb core because of zero-sequence voltages and should have small leakage reactances.

Concerning the input voltage formation the basic function of the interphase transformer is equivalent to a symmetric inductive voltage divider inserted between two voltage systems of 6-pulse staircase shape having a phase displacement of 30° el. This fundamental phase displacement constitutes a phase shift of π for the 6th harmonic components. Therefore, the voltage available at the center tap only shows harmonics of 12th and higher order besides the fundamental resulting in harmonics of the system input current. The current-forming effect of the harmonics of the 12-pulse AC system is limited (in analogy to pulse rectifier systems) by series inductors on the mains side. Therefore, due to smoothing of the current, also the decrease of the amplitudes of the mains current harmonics with increasing frequency is higher in comparison to conventional twelve-pulse rectification. A variant of the circuit showing 18-pulse characteristic is proposed in [58].

A similar concept being characterized by impressed output current and being applicable also for 18-pulse rectification has been proposed in [59].

3.8 Isolated Converter Topologies

As already discussed in section 2 a galvanic isolation of the output voltage can be achieved by a mains frequency of a high-frequency transformer (cf. Fig.1). In the following for the sake of brevity only a representative circuit of each class showing low realization effort will be treated. Concerning a detailed discussion and evaluation under consideration of single-switch single-stage SEPIC-type [35], flyback [36] and Cuk-type [37] converter systems we would like to refer to a publication being currently in preparation.

3.8.1 Scott Transformer Isolated Two-Switch Boost Power Factor Corrector

By space vector transformation three linear dependent phase quantities are mapped into a complex phasor and/or a two-phase system (real and imaginary component of the phasor) Accordingly, for symmetric current consumption can be achieved for a three-wire system by using only two single-phase boost-type power factor correction systems with common output capacitor and two Scott-connected single-phase transformers. This transformer arrangement transforms the orthogonal sinusoidal current consumption of the power factor correctors into a symmetric three-phase current system. Alternatively, a three-phase magnetic core with a winding arrangement according to Leblanc could be employed [30].

For a turns ratio of the Scott-transformer primaries $\frac{n_{12}}{n_{11}} = \frac{2}{\sqrt{3}}$ (n_{12} denotes the total number of turns of the center tapped transformer primary) and equal number of turns n_2 of the transformer secondaries equal input voltages result for both power factor correctors; furthermore, then advantageously a symmetric three-phase primary current system is formed for equal peak value of the boost converter input currents. Considering the transformer stray inductance a low harmonic level of the mains current can be achieved by connecting filter capacitors in parallel to the secondaries.

As compared to the conventional combination of three isolated single-phase PWM rectifier systems for realizing a three-phase AC/DC converter the system depicted in Fig.10 shows a lower control and realization effort. However, this advantages are possibly partly compensated by the requirement of a special transformer arrangement resulting in a higher construction effort and realization costs as compared to a conventional three-phase transformer.

3.8.2 VIENNA Rectifier II

Conventional single-stage three-phase PWM rectifier systems are characterized by an three-phase/single-phase AC/AC matrix converter topology of the primary power circuit [38] and/or a relatively high realization effort. A novel single-stage system showing a significant lower complexity has been proposed in [40] (cf. Fig.11) and will be discussed briefly in the following.

Dependent on the converter switching state the input phase currents are guided via the four-quadrant switches at the input or the associated free-wheeling diodes, the transistors of the bridge leg and the high-frequency transformer primary. Due to the secondary current balancing the primary current ampere turns the output voltage is applied to the secondary via the output diode bridge. Therefore, on the primary a voltage being equivalent to the DC output voltage of the

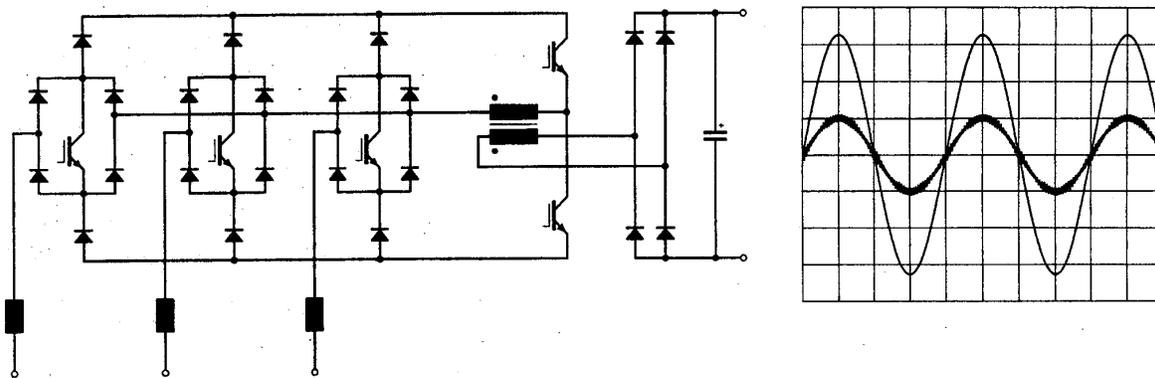


Fig.11: Basic structure of the power circuit and mains current and mains current shape of a single-stage high-frequency isolated boost-type pulse rectifier system (VIENNA Rectifier II). Scales: 100V/div, 20A/div, 5ms/div.

circuit shown in Fig.1 concerning the control of the input current. Concerning a detailed discussion of the basic principle of operation we have to refer to section 3 in [40] for the sake of brevity. Due to operating the transformer with impressed input current, an overvoltage limitation circuit has to be provided at the primary. With respect to high efficiency the limitation power is delivered to the output by a low-power isolated DC/DC converter. This DC/DC converter also can be used for precharging the output capacitor during system start-up and for supplying the control electronics.

We would like to point out that for a dual exchange of the energy storage elements at the input and output of the PWM rectifier system shown in Fig.11 (i.e., for replacing the input inductances by input capacitors and replacing the output capacitor by an output inductor) a buck-type PWM rectifier system as proposed in [41] (VIENNA Rectifier III) can be derived. There the high frequency transformer is operated voltage-fed, therefore, the overvoltage clamping circuit on the primary can be omitted. Furthermore, no precharging of the output capacitor has to be performed in connection with the start-up of the system.

4 Comparative Evaluation

In the literature in connection with the discussion of novel converter topologies in most cases no comparative evaluation to known circuit concepts, i.e. a comparison of advantages and drawbacks concerning realization effort, stresses on the components etc. is given.

Therefore, in the following in extension to [11] and [37] and with reference to papers on the comparison of DC/DC converter systems [60–62] a first evaluation of important rectifier topologies discussed in section 3 is given. For the sake of brevity there only non-isolated converter systems are considered.

4.1 Evaluation Criteria, Characteristic Quantities, Definitions

With respect to the calculation effort for evaluating the converter systems only integral characteristic quantities and/or the sum of component voltage or current mean and rms values, apparent switching powers etc. are considered. The

characteristic quantities are calculated for a given set of operating parameters (cf. section 4.2) and can be used for a first evaluation of a converter topology concerning

1. complexity (number of components)
2. utilization of the components (silicon utilization [37])
3. voltage and current stresses on the components (conduction and switching losses)
4. aspects of system technology (control complexity, system behavior for phase loss, etc.).

Of course there specific advantages of a circuit concept if, e.g., wide input voltage range or soft-switching is required cannot be considered. In case the characteristic quantities do not clearly identify an advantageous converter topology a detailed dimensioning of the most promising systems and a calculation of their realization costs etc. has to be performed.

4.1.1 Characteristic Quantities, Definitions

Transistor Utilization

$$\mu_T = \frac{P_O}{\sum_n u_{T,\max,n} i_{T,\max,n}} \quad (1)$$

μ_T gives the output power delivered per volt-ampere of the total stress $\sum_n u_{T,\max,n} i_{T,\max,n}$ on the power transistors (cf. *component load factor* in [60] and *silicon utilization* in [37]). Analogous to the total power factor in AC circuits an ideal figure would be $\mu_T = 1$. $u_{T,\max,n}$ denotes the maximum transistor blocking voltage under neglect of switching overvoltages, $i_{T,\max,n}$ denotes the peak value of the current in power transistor n . The possibility of a bidirectional energy conversion given for the system shown in Fig.8 could be considered by doubling μ_T in a more general evaluation of converter topologies.

Diode Utilization

$$\mu_D = \frac{P_O}{\sum_n u_{D,\max,n} i_{D,\max,n}} \quad (2)$$

(cf. Eq.(1)).

Total Semiconductor Utilization

$$\mu_{D,g} = \frac{1}{\frac{1}{\mu_{D,1}} + \frac{1}{\mu_{D,2}}} \quad (3)$$

($\mu_{D,1}$ and $\mu_{D,2}$ denote, e.g., the utilization given for the line-frequency and fast-recovery diodes of a rectifier circuit.

Transistor Conduction Losses

$$p_T = \frac{1}{I_O} \sum_n I_{T,avg,n} \quad (4)$$

I_O denotes the mean value of the load current. Eq.(4) in a first approximation characterizes the conduction losses for constant (current-independent) forward voltage drop of the power transistors. For power semiconductors with ohmic on-characteristic the current mean value would have to be replaced by the current rms value.

Diode Conduction Losses

$$p_D = \frac{1}{I_O} \sum_n I_{D,avg,n} \quad (5)$$

(cf. Eq.(4)).

Transistor Switching Losses

$$s_P = \frac{1}{k_P} \frac{1}{P_O} \sum_n |i_T|_{avg,n} U_{T,off,n} \quad (6)$$

Under assumption of constant pulse frequency, current proportional switching loss and constant turn-off voltage the average value of the turn-on/turn-off current is determining for the switching losses. The avoidance of turn-on losses (cf. section 3.4) or the reduction of the turn-on/turn-off voltage to half the output voltage value for the rectifier topologies discussed in sections 3.2 and 3.5 is considered by a factor $k_P = 2$.

Rated Power of Inductors and Transformers

In order to make possible the calculation of a total rated power of all inductive components (inductors and transformer) inductors are characterized by the rated power of an equivalent transformer [62]

$$s_{T,eq} = \frac{1}{P_O} 2.2 f_N L I_{rms} I_{max} k_B \quad (7)$$

For high-frequency components with ferrite cores (e.g. input inductors of systems 3.4, 3.5, and 3.6) there a maximum allowable magnetization being lower by a factor of $k_B = 5$ is assumed as compared to low-frequency components. This corresponds to a ferrite saturation flux density of typically 0.3 T, and an allowable flux-density of 1.5 T in the magnetic core of a mains frequency transformer.

In general we have for the rated power of a transformer

$$s_T = \frac{1}{P_O} \sum_i S_i \quad (8)$$

where S_i denotes the apparent power of a partial winding.

Electrolytic Capacitor Current Stress

$$i_C = \frac{1}{k_{ESR}} \frac{I_{C,rms}}{I_O} \quad (9)$$

Eq.(9) characterizes the current stress on the output electrolytic capacitor being determining for the dimensioning and the capacitor service life. For capacitor current stresses with frequencies higher than 10 kHz the reduction of the equivalent series resistance (ESR) with increasing frequency is considered by a factor $k_{ESR} = 1.5$. Due to the high output voltage level ($U_O = 700V$) for all systems a series

connection of 2 capacitors and/or the same total number of electrolytic capacitors has to be employed.

Power Factor

For the definition of the power factor we have

$$\lambda = \frac{P_O}{\sum \sqrt{3} U_{N,rms} I_{N,rms}} \quad (10)$$

With the exception of system 3.7 the input current rms value $I_{N,rms}$ also contains components with switching frequency besides the mains current fundamental. For the system discussed in section 3.4 a single-stage LC-filter is considered in order to limit the input current ripple according to Eq.(11). For additional filtering on one hand the harmonic level of the mains current would be decreased resulting in a marginal reduction of $I_{N,rms}$, however, on the other hand the mains current consumption would increase due to the increasing reactive power of the filter capacitors.

Remark: Besides the stresses on the components obviously the realization costs would be of special interest in connection with a comparative evaluation of the systems. As described in [69] such a comparison has been performed based on cost functions for the major components in connection with selecting a suitable voltage level for the electrical power plant of future internal combustion engine vehicles. However, due to its complexity a corresponding evaluation of three-phase PWM rectifier topologies is beyond the scope of this paper.

4.2 Assumptions

The characteristic quantities defined in the foregoing section are calculated by digital simulation for stationary operation and the following operating parameters:

$$\begin{aligned} P_O &= 10 \text{ kW} \\ U_N &= 400 \text{ V}_{rms} \text{ (line - to - line)} \\ f_N &= 50 \text{ Hz} \\ U_O &= 700 \text{ V} \\ f_P &= 25 \text{ kHz.} \end{aligned}$$

(P_O denotes the output power, U_N and f_N are the rms value of the mains line-to-line voltage and the mains frequency, U_O denotes the output voltage and f_P defines the average switching frequency within a mains period). For the uncontrolled rectifier circuit discussed in section 3.7 the output voltage resulting for P_O is $U_O = 482.5V$.

For the self-commutated rectifier systems an equal average absolute value of the input current ripple

$$\text{avg}\{|\Delta i_{N,P}|\} = 1 \text{ A} \quad (11)$$

is set by proper dimensioning of the input inductors. For a purely triangular shape of the ripple this corresponds to a ripple amplitude of $\pm 2A$ and/or in the case at hand to $\pm 10\%$ of the mains current fundamental. (The rms value of the input current ripple being different for the different systems doesn't represent an important quantity). Due to defining a switching frequency and a average input current ripple being equal for all systems advantages of a system concerning the harmonic level of the input current are expressed by a low rated power of the ripple-limiting inductors.

Remark: Besides Eq.(11) and/or the ripple average time behavior a basis for comparison also could be defined in the

frequency domain (e.g. based on the amplitudes of the harmonics with switching frequency [63]). However, this results in a considerably higher complexity of the comparative evaluation of the different systems due to the higher calculation effort and the different distribution of the harmonics in the input current spectra.

Furthermore, the simulations performed for determining the characteristic current peak, average and rms values are based on the following assumption:

- purely sinusoidal symmetric mains voltage system
- no inner mains impedance
- ideal constant output voltage
- ideal components (no forward voltage drop and no switching losses of the power semiconductors, no copper losses, no hysteresis losses and no eddy current losses of the inductors and transformers, no equivalent series resistance of the electrolytic capacitors, etc.)
- neglectation of the switching frequency ripple for calculation of the component peak current values.

4.3 Simulation Results

The simulation results are compiled in Tab.1 and will be discussed in the following section.

4.4 Discussion

The system according to section 4.1 (which is used as a basis for comparison in [11]) shows a high utilization of the power transistor and of the output diode and low transistor and diode conduction losses due to employing only a single device of each type and due to the operation of the DC/DC boost converter stage with about constant input voltage. Therefore, the only drawback consists in the low-frequency distortion of the input current.

For system discussed in section 3.4 the high peak value of the discontinuous input current results in lower transistor utilization and higher rms current stress on the output capacitor as compared to system 3.1. However, the transistor switching losses of both systems are about equal due to discontinuous output diode current shape and/or the avoidance of turn-on losses of the power transistor. In order to reach a continuous shape of the input current having a average ripple amplitude according to Eq.(11), one has to connect a LC-filter in series on the mains side. This results in an increase of the rated power of the inductive components as given in brackets in Tab.1. However, the inductance values are still considerably lower than for system 3.1.

Again a disadvantage is given by the low frequency distortion of the mains current which only could be reduced by selecting a higher output voltage resulting in a higher blocking voltage stress on the valves. Under consideration of the low complexity of the system control and of the possibility of a staggered operation of two systems of half rated output power connected in parallel (resulting in a continuous shape of the input current) system 3.1 and system 3.4 show about equal performance for realizing PWM rectifier systems of up to 10kW output power.

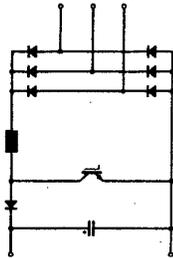
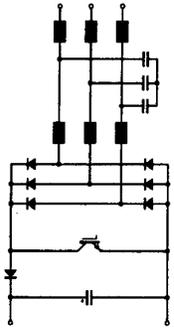
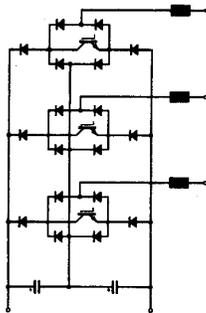
A purely sinusoidal input current shape is only avoided for the system according to sections 3.5. As compared to system 3.1 there the utilization of the power transistors μ_T is

lower due to the sinusoidal (i.e. not constant) current shape and the higher number of transistors. However, as compared to system 3.6 being comparable concerning the mains current quality μ_T shows a able high value. The main reason is the low blocking voltage stress on the power semiconductors which is defined by half the output voltage and the utilization of the power transistors for conducting the current within the positive and negative input current half period. Furthermore, the total transistor losses remain limited to low values due to the low turn-off voltage. Of further importance is the three-level characteristic of the bridge legs which results in a significant reduction of the rated power of the mains-side inductances for a given average input current ripple amplitude as compared to system 3.6 showing a two-level characteristic of the bridge legs. Therefore, in case a high mains current quality is required system 3.5 represents an ideal solution for realizing a unidirectional PWM rectifier system.

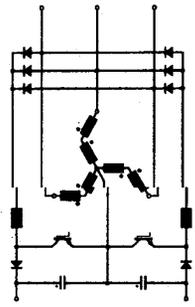
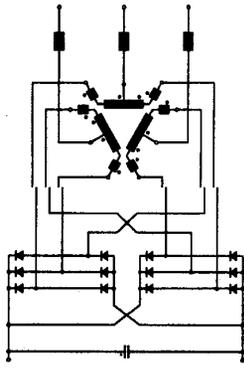
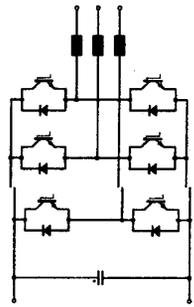
Of special interest is also a comparison of the system according to sections 3.2 and 3.7. Both systems show about equal requirements concerning the rated power of the inductive components and equal mains current quality. For system 3.7 the rms current stress is limited to very low values and the mains current doesn't contain any switching frequency harmonics, therefore, the only remarkable disadvantage of the systems consists in the dependency of the output voltage on the mains voltage and on the load. However, in case system 3.2 is fed via a delta-wye-step-down transformer the magnetic injection device can be omitted (then the third harmonic current can be injected into the neutral point of the wye-connected secondary) and the realization effort is limited to power semiconductors of low blocking capability (half the output voltage) and a control of low complexity. Accordingly, circuit 3.2 represents an extremely attractive solution guaranteeing optimal utilization of a DC/DC converter stage connected in series due to the controlled output voltage.

Besides the calculated characteristic quantities in course of the selection a PWM rectifier topology attention should also be paid to the items listed in the following:

- three-phase PWM rectifier systems in general show a common-mode voltage with switching frequency between the output voltage center point and the mains star point. Therefore, a common-mode filter is required for limiting conducted common-mode emissions
- for assessing the filtering effort required for the limitation of conducted differential-mode emissions the amplitudes of the harmonics with switching frequency of the rectifier input voltage should be considered. The rms value of the mains current ripple is of very limited importance with evaluating the performance of a system with respect to harmonic regulations.
- the reduction of electromagnetic interference and the increase of the rectifier efficiency by extending the system operation to soft switching should be opposed to the resulting increase of the system complexity and the realization effort
- buck-type converter systems should be paid more attention in the analysis and comparison of three-phase PWM rectifier systems
- single-stage converter systems only show significant advantages concerning power density and realization costs over two-stage systems for limited input or output voltage range

topology	section 3.1	section 3.4	section 3.5
characteristic value			
number of transistors	1	1	3
transistor utilization	0.70	0.26	0.47
transistor conduction losses	0.31	0.32	0.78
transistor switching losses	1.30	1.23	1.36
number of diodes			
line frequency diodes	6	-	12
fast recovery diodes	1	7	6
diode utilization			
total	0.32	0.28	0.36
line frequency diodes	0.48	-	0.57
fast recovery diodes	1.00	0.28	1.00
diode conduction losses	3.62	3.61	5.55
number of single- or three-phase inductive components	1	3 (6)	3
total rated power of inductive components	0.03	0.0024 (0.0066)	0.015
line frequency ind. components	-	-	-
switching frequency ind. comp.	0.03	0.0024 (0.0066)	0.015
output capacitor current stress	0.37	0.75	0.36
displacement factor	1.0	1.0	1.0
total harmonic distortion of the mains current	0.32	0.2	0.082
total power factor	0.953	0.981	0.997
rated low frequency mains current harmonics			
\hat{i}_5	0.17	0.190	0.01
\hat{i}_7	0.17	0.016	0.00
\hat{i}_{11}	0.09	0.014	0.00
\hat{i}_{13}	0.09	-	0.00
direct mains phase current control	yes	no	yes
output voltage control	yes	yes	yes
robust to utility voltage unbalances	yes	yes	yes
sensing effort	medium	low	medium
control complexity	low	low	medium

Tab.1: Comparative evaluation of three-phase PWM rectifier systems.

topology	section 3.2	section 3.7	section 3.6
			
characteristic value			
number of transistors	2	0	6
transistor utilization	0.45	-	0.117
transistor conduction losses	0.39	-	0.5
transistor switching losses	1.19	-	2.76
number of diodes			
line frequency diodes	6	12	-
fast recovery diodes	2	-	6
diode utilization			
total	0.34	0.427	0.40
line frequency diodes	0.52	0.427	-
fast recovery diodes	1.00	-	0.40
diode conduction losses	4.4	2.9	2.46
number of single- or three-phase inductive components	3	4	3
total rated power of inductive components	0.232	0.206	0.032
line frequency ind. components	0.2	0.206	-
switching frequency ind. comp.	0.032	-	0.032
output capacitor current stress	1.25	0.03	0.39
displacement factor	1.0	0.963	1.0
total harmonic distortion of the mains current	0.072	0.077	0.081
total power factor	0.997	0.961	0.997
rated low frequency mains current harmonics			
\hat{i}_5	0.028	0.028	0.01
\hat{i}_7	0.012	0.018	0.01
\hat{i}_{11}	0.016	0.061	-
\hat{i}_{13}	0.015	0.040	-
direct mains phase current control	yes	no	yes
output voltage control	yes	no	yes
robust to utility voltage unbalances	yes	no	yes
sensing effort	medium	-	high
control complexity	medium	-	high

Tab.1: Comparative evaluation of three-phase PWM rectifier systems (continued).

- a sinusoidal input current shape is believed to gain increasing importance for power supply manufacturers in connection with selling their product in a competitive market also if not required by explicit harmonic standards.

5 Future Trends

In conclusion, we want to mention briefly major areas of future developments of pulse rectifier systems evolving from the increasing number of contributions of different authors at the relevant scientific conferences. (These two topics can be seen besides the integration of power semiconductor and control functions into intelligent power modules and the development and analysis of new pulse rectifier topologies.)

Concerning the effects on the mains with switching frequency caused by pulse rectifier systems it is tried to obtain a wider distribution of harmonic power (resulting due to the operating principle) and/or a reduction of the amplitudes of single frequency components by varying the pulse frequency (e.g., random pulse width modulation [64]). In this area a clarification of the standpoints would be advantageous also in connection with the standards being in preparation regarding the admissible harmonics stress on the mains; this should make clear if as much as possible even distribution of the harmonic power within a frequency band (and, therefore, minimum harmonic stress on the mains for discrete frequencies) should be preferred as compared to a concentration of the harmonic power around multiples of the mains frequency.

For dimensioning of high-frequency pulse rectifier systems one has to pay special attention (besides to dimensioning of the power semiconductors) to the design of an input filter to be provided (if required) for the reduction of the load on the mains by current harmonics and/or for avoiding of EMI. By the pulse rectifier systems a constant power flow is impressed in the stationary case (according to a negative differential input impedance of the system). This results in a reduction of the filter damping. In most cases the input filter (being realized as single stage filter in the simplest case) has to be included into the design of the system control. There, for stability basically a filter output impedance is required being lower than the input impedance of the rectifier system and/or the parallel resonance of the filter being present for the rectifier system has to be sufficiently damped (actively or passively) and/or it has to be positioned above the corner frequency of the output voltage control. Thereby, also an excitation of the filter by low-frequency harmonics of the mains voltage (series resonance) is avoided. In any case, the pulse frequency of the rectifier system has to be set sufficiently above the resonance frequency of the filter in order to obtain sufficient damping of the current harmonics with switching frequency. However, also in this case a resonance condition can result by periodic load changes. In connection with the problem of proper selection of the pulse frequency we also have to point out systems with time-variable switching frequencies (e.g., given for direct impression of the input currents by hysteresis control [65] or by specific frequency-modulation or random modulation [64, 66]). In general, one has to emphasize that the optimum dimensioning of the mains filter with regard to rated power, weight and costs constitutes an important problem of the practical application of pulse rectifier systems. This should be paid more attention in the scientific literature in the future (besides

developing of new circuit topologies).

Due to the limited life time of electrolytic capacitors (limited by technological reasons) the replacement of the output (DC-link) capacitors by foil capacitors with long utilization periods is discussed for pulse rectifier systems with high switching frequencies [67]. However, there results a closer coupling of the load-side converter with the mains-side pulse rectifier system; this means that also the load behavior influences the effects on the mains and that a relatively unclear mix between load-side and rectifier-side harmonic components results. Assuming a wider application in the future of pulse converter systems with low effects on the mains and/or of power electronic energy converters, in general, we have to point out the deterioration of the damping conditions in low-voltage supply systems to be expected due to the constant power characteristic of these systems [68]. Therefore, under further consideration of the efficiency reduction (2...3%) as compared to simple diode rectification) resulting typically for connecting a pulse rectifier system in front of each single power electronic system one has to pose the question in general how a technically and economically optimal solution of the problem area of effects on the mains can be achieved (e.g., by a combined application of mains-friendly pulse rectifier systems and centrally located active or hybrid filters).

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