Comparative Evaluation of Three-Phase High Power Factor AC-DC Converter Concepts for Application in Future More Electric Aircrafts

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Abstract. – A passive 12-pulse rectifier system, a two-level, and a three-level active three-phase PWM rectifier system are analyzed for supplying the DC voltage link of a 5kW variable speed hydraulic pump drive of an electro-hydrostatic actuator to be employed in future More Electric Aircrafts. Weight, volume and efficiency of the concepts are compared for an input phase voltage range of 98V…132V and an input frequency range of 400….800Hz. The 12-pulse system shows advantages concerning volume, efficiency and complexity but is characterized by a high system weight. Accordingly, the three-level PWM rectifier is identified as most advantageous solution. Finally, a novel extension of the 12-pulse rectifier system by turn-off power semiconductors is proposed which allows a control of the output voltage and therefore eliminates the dependency on the mains and load condition which constitutes a main drawback of the passive concept.

I. INTRODUCTION

Large transport category airplanes are currently equipped with three independent hydraulic systems and two independent electrical systems [1]. On future More Electric Aircrafts the conventional fly-by-wire hydraulic flight control surface (rudder, aileron, spoiler etc.) actuation which is supplied from the centralized hydraulic generation and distribution systems will be associated with electrically powered electro-hydrostatic actuators (EHA). This will allow to eliminate one hydraulic system without impairing safety objectives since one hydraulic supply is replaced by two electrical systems and/or the power source redundancy is increased. Further advantages are higher flexibility in routing and weight and cost savings due to the reduction of the total number of hydraulic generation and distribution components as well as higher efficiency.

The hydraulic power of the EHAs is generated locally by dedicated hydraulic pumps which are driven by variable speed electric motors being fed by an inverter from a voltage DC link (cf. Fig.2 in [2]). In order to prevent a distortion of the supply voltage and/or an interference with sensitive avionics equipment, rectifier concepts with low effects on the mains have to be employed. There, (currently) only unidirectional power conversion is required; energy which is fed form the loaded surface back into the DC link is dissipated in a resistive dump circuit (cf. Fig.3 in [3]).

Rectifier systems with high input current quality basically can be realized as passive multi-pulse rectifiers or as high switching frequency PWM rectifier systems which allow an active control of the shape and phase displacement of the input current and of the output voltage [4]. In order to provide a basis for a decision, both concepts have to be compared concerning volume, weight, efficiency, complexity/reliability and realization costs.

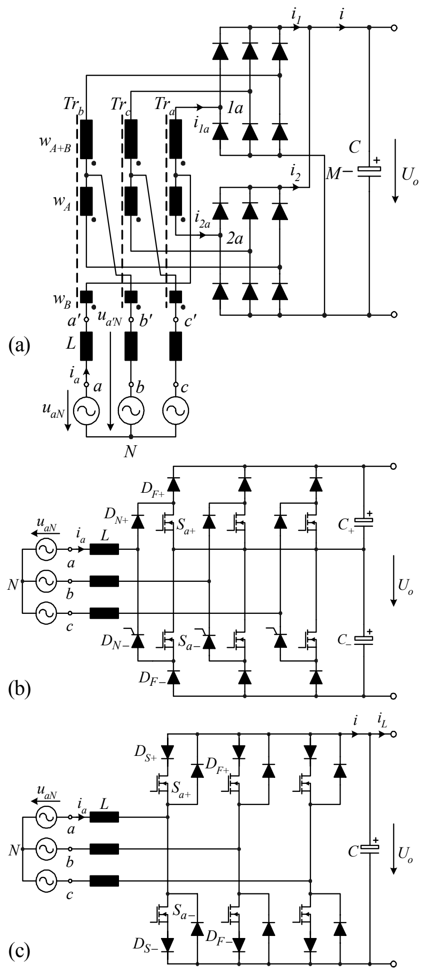

Fig.1: Three-phase AC/DC converter concepts with low effects on the mains. **(a)** twelve-pulse passive rectifier according to [5], **(b)** unidirectional three-level PWM rectifier according to [6], **(c)** conventional two-level PWM rectifier.

The topic of this paper is a comparative evaluation of a passive autotransformer-based 12-pulse rectifier with capacitive smoothing (cf. **Fig.1**(a), [5]) and of an active three-level boost-type PWM rectifier system (cf. Fig.1(b), [6]). Furthermore, a conventional two-level PWM rectifier concept as known from industrial automation systems is considered for reference purposes (cf. Fig.1(c)). There, the system operating parameters are defined as

$U_{N,rms}$ = 115V ± 15% ≈ 98…132V
f_N = 400Hz….800Hz
P_O = 5kW

($U_{N,rms}$ denotes the RMS value of the mains phase voltage, f_N is the mains frequency, P_O is the output power) like typically given for aileron actuation systems. An actuation system is characterized by a high peak-to-continuous power rating, i.e. short periods (typ. 1…30s) of high power demand when the control surface is accelerated and moved at high speed against load, and longer periods where the surface is held in a defined position (cf. Fig. 3 in [2] or Fig.6 in [3]). In the case at hand P_O represents an average power level which is considered as continuous output power demand in a rough first approximation.

In a similar way the specified input voltage range $U_{N,rms}$=98V…132V (115V±15%) considers long duration transients in normal operation down to a fraction of an actuation interval (cf. Fig.2 in [3]).

Currently for the generation of constant frequency (400Hz) AC voltages from variable frequency engine shafts complex hydraulic systems are employed which will be replaced in future by less complex lighter variable speed generators in combination with power electronics (cf. p. 249 in [7]). Accordingly, a wide input frequency range of f_N=360…800Hz (cf. p. 5 in [8]) will be required for future aircrafts. As 360Hz is close to today's nominal frequency f_N=400Hz, for limiting to the essentials only the input frequency range f_N=400…800Hz is considered in the following.

In **Section II** and **Section III** the basic function of the rectifier systems is discussed briefly and the distribution of the losses to the main power components is shown. In **Section IV** a comparative evaluation of the concepts concerning volume, weight and efficiency is given considering worst case operating points as resulting from the specified input voltage and input frequency range. In **Section V** a novel concept for controlling the output voltage of the 12-pulse rectifier is proposed and experimentally verified. Finally, in **Section VI** topics of further research are identified which will be oriented to active buck-type PWM rectifier system which allows a direct system start-up without output voltage pre-charging and/or the limitation of the input current in case of an output short circuit.

II. TWELVE-PULSE AUTOTRANSFORMER RECTIFIER SYSTEM

In order to eliminate low-frequency input current harmonics 6-pulse diode bridge rectifier systems could be combined to multi-pulse systems using phase-shifting isolation or autotransformers. In the case at hand the 12-puls rectifier concept shown in Fig.1(a) with impressed output voltage, mains side interphase transformer [5] and input inductors is considered which directly provides the DC link voltage of a PWM inverter connected in series. Alternatively, a rectifier system with DC side interphase transformer [6] could be employed. Both concepts are characterized by an about equal total rated power of the magnetic components, i.e. of the autotransformer and the AC or DC side inductors of ≈20% of the DC output power. However, for the system according to [6] higher amplitudes of the input current harmonics would occur (cf. Fig.4 in [5] und Fig.4(b) in [9]) and the blocking voltage stress on the rectifier diodes would not be defined directly by the output voltage.

The time behavior of characteristic quantities of the 12-pulse rectifier system is depicted in **Fig.2**.

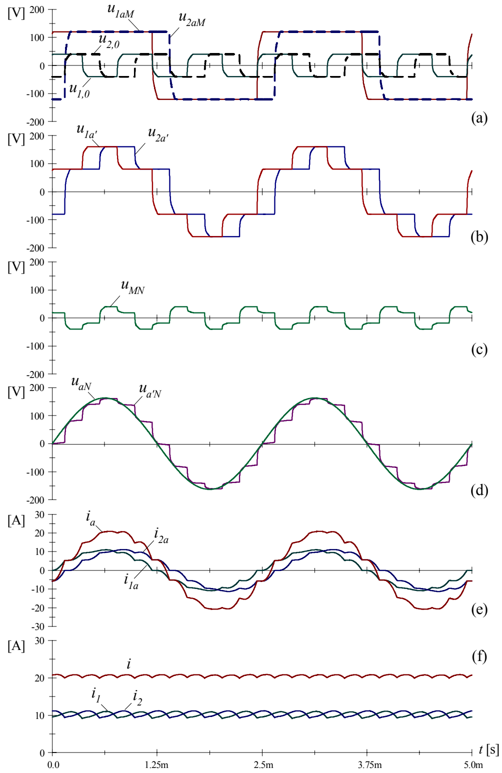

Fig.2: Operating behavior of the passive 12-pulse rectifier system shown in Fig.1(a); **(a)** input phase voltage of corresponding diode rectifier bridge legs with reference to the DC output voltage center point M and zero-sequence components $u_{1,0}$, $u_{2,0}$ contained in the output phase voltages of the diode rectifier bridges; **(b)** zero-sequence free components u_{1a}', u_{2a}' of the phase voltages shown in (a); **(c)** common-mode component u_{MN} of the output voltage; **(d)** mains phase voltage u_{aN} and corresponding input phase voltage $u_{a'N}$ of the line interphase transformer; **(e)** mains phase current i_a and input currents i_{1a}, i_{2a} of the corresponding interphase transformer windings; **(f)** output currents i_1, i_2 of the rectifier bridges and total DC current i. Simulation parameters: $\hat{U}_N$=115V, f_N=400Hz , P_O=5kW, for further parameters see Tab.1; for each phase ideal magnetic coupling of the windings of the interphase transformer has been assumed.

The autotransformer (interphase transformer) provides a symmetric splitting of the largely sinusoidal mains current into two partial current systems i_{1a}, i_{1b}, i_{1c} and i_{2a}, i_{2b}, i_{2c} (cf. i_a and i_{1a}, i_{2a} in Fig.2) with a phase displacement ±15°. Due to the continuous shape of the phase currents i_{1i} and i_{1i} , i=a,b,c, the rectifier bridge input voltages will show a square wave shape with reference to the center point M of the output voltage U_O where the voltages of corresponding phases, e.g. u_{1aM} and u_{2aM}, will exhibit a phase difference of 30° in correspondence to the phase displacement of the related phase currents i_{1a} und i_{2a}. After subtraction of the zero sequence components $u_{1,0}=\frac{1}{3}\Sigma u_{1iM}$ and $u_{2,0}=\frac{1}{3}\Sigma u_{2iM}$ (i=a,b,c) contained in the voltage systems u_{1aM}, u_{1bM}, u_{1cM} and u_{2aM}, u_{2bM}, u_{2cM} the 6-pulse voltage systems $u_{1i}'= u_{1iM}-u_{1,0}$ and $u_{2i}'= u_{2iM}-u_{2,0}$ are remaining at the autotransformer outputs. There, corresponding voltages, i.e. u_{1i}' and u_{2i}' again show a phase displacement of 30° and are combined by the autotransformer into a 12-pulse voltage system occurring at the input terminals a', b', c' (cf. e.g. $u_{a'N}$ in Fig.2).

The difference of the autotransformer input voltage $u_{i'N}$ and the mains voltage u_{iN} occurs across the input inductors L and defines a mains current i_i of largely sinusoidal shape as only harmonics with ordinal numbers n= 11,13, 23,25 are present in $u_{i'N}$.

Due to the low distortion of the mains current the instantaneous system input power is largely constant what results a largely constant

DC output current i. Accordingly, a low current stress on the output capacitor will occur and the output voltage will show a low ripple also for low smoothing capacitance C_O.

According to [5] w_A/w_B= 0.366 has to be selected for the autotransformer windings in order to guarantee a balancing of the ampere-turns of the corresponding phase windings and/or to achieve the above-mentioned symmetric current splitting of the mains current and/or 12-pulse combination of the diode bridge input voltages.

The fundamentals of the corresponding phase voltages and currents at the rectifier bridge inputs do not show a phase displacement (cf., e.g., u_{1a}' and i_{1a} in Fig.2). Accordingly, for limiting to the fundamentals a symmetric three-phase ohmic loads can be assumed at the autotransformer outputs which translates into an ohmic autotransformer fundamental input behavior. Considering the fundamental voltage drop across input inductors we therefore have for the phase displacement of the mains voltage and mains current fundamental

$$\cos\varphi = \sqrt{1-\left(\frac{\omega L \hat{I}_{N,(1)}}{\hat{U}_N}\right)^2} \qquad (1)$$

(cf. Eq.(3.1.1/13) in [5], $\hat{U}_N$ denotes the amplitude of the purely sinusoidal mains voltage, $\hat{I}_{N,(1)}$ is the amplitude of the mains current fundamental). For neglecting system losses the system output power then is given by

$$P_O = \frac{3}{2}\hat{U}_N \hat{I}_{N,(1)} \cos\varphi = \frac{3}{2}\hat{U}_N \hat{I}_{N,(1)} \sqrt{1-\left(\frac{\omega L \hat{I}_{N,(1)}}{\hat{U}_N}\right)^2}\ . \qquad (2)$$

In [5] the fundamental of the 12-pulse phase voltages $u_{i'N}$ at the autotransformer input has been shown to be

$$\hat{U}_{U,(1)} = \frac{2}{3} U_O \frac{\sin\frac{\pi}{12}}{\frac{\pi}{12}} \approx 0.66\, U_O\ . \qquad (3)$$

For light load (Index 0) only a low fundamental voltage drop across the input inductors does occur and/or $\hat{U}_{U,(1),0} \approx \hat{U}_N$ or $U_{O,0} \approx 1{,}52\hat{U}_N$ (considering (3)) is valid. Accordingly, we have for the output voltage at larger load (continuous shape of the bridge rectifier input

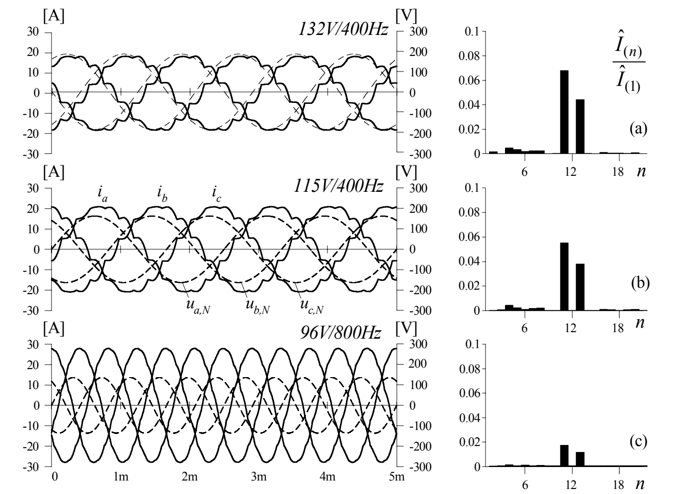

Fig.4: Time behavior of the mains current and related low-frequency harmonics (normalized to the amplitude $\hat{I}_{N,(1)}$ of the fundamental) for **(a)** maximum input voltage, $U_{N,rms}$=132V, and minimum input frequency, f_N=400Hz (worst case concerning the amplitudes of the current harmonics); **(b)** rated mains voltage and rated frequency, $U_{N,rms}$=115V, f_N=400Hz; **(c)** minimum input voltage $U_{N,rms}$=98V and maximum frequency f_N=800Hz.

phase currents) under neglection of the conduction voltage drops of the diodes and the parasitic windings resistances and under assumption of an ideal coupling of the autotransformer phase windings

$$U_O \approx U_{O,0}\cos\varphi \approx 1.52\,\hat{U}_N \cos\varphi \qquad (4)$$

(cf. Eq.(3.1.1/12) in [5]), and Fig. 6). The actual dependency of the output voltage determined by a digital simulation shown in **Fig.3** reveals that (4) only constitutes a very rough approximation. This is due to the fact that the assumed continuous shape of the bridge rectifier input currents and/or pure staircase shape of the voltages $u_{i'N}$ (cf. (3)) is only given for high output power and/or close to rated load (cf. $u_{a'N}$ in Fig.2(c) and Bild 3.1.2/5 in [5]). For low output power the current conduction of the diode bridges is discontinuous and sinusoidal segments of the autotransformer input voltage do occur resulting in $U_O > U_{O,0}$. For no load operation the DC output voltage reaches the amplitude of the mains line-to-line voltage, $U_{O,max} = \sqrt{3}\hat{U}_N$ (cf. Fig.6). Equation (4) therefore can only be employed for the calculation of the output voltage resulting for rated load.

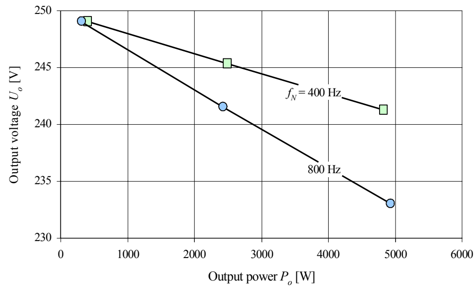

Fig.3: Dependency of the rectifier output voltage on the output power for $U_{N,rms}$=115V and mains frequencies f_N= 400Hz, 800Hz as determined by digital simulation. Simulation parameters: cf. Tab.I, coupling of the windings of a phase leg of the inpterphase transformer assumed ideal.

The input current harmonics $\hat{I}_{N,(n)}$ are determined by the harmonics of the autotransformer input voltage

$$\hat{U}_{U,(n)} = \frac{1}{n}\hat{U}_{U,(1)} \qquad (5)$$

with ordinal numbers n=12k±1 (k=1,2,3..) and the inductance L connected in series

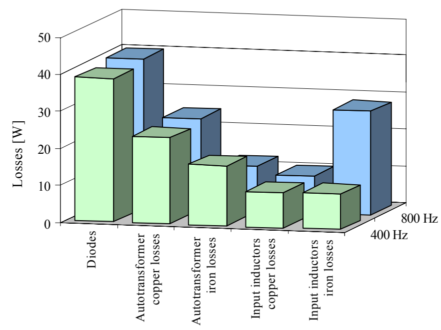

Fig.5: Distribution of the losses for $U_{N,rms}$=115V and different mains frequencies f_N= 400Hz and 800Hz. Frequency dependent losses of the windings of the inductive components due to skin and proximity effect and eddy currents are not considered. Iron losses of the interphase transformer decrease with increasing frequency because of the decreasing amplitude of the magnetic flux density. On the other hand, the iron losses of the input side inductances increase with increasing frequency as the current is defined for a certain output power and mains voltage amplitude and the hysteresis losses are increasing with increasing frequency.

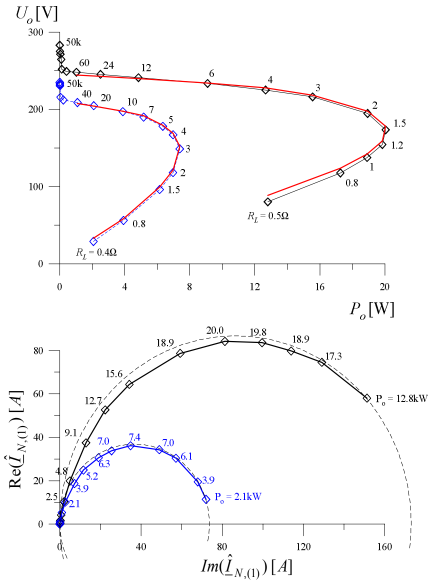

Fig.6: Simulation **(a)** of the dependency of the output voltage U_O on the output power P_O (parameter: load resistor) and **(b)** of the trajectory of the phasor of the mains current (the mains voltage is assumed to be lying in parallel to the vertical real axis; parameter: P_O). Simulation parameters: $U_{N,rms}$= 115V@f_N=400Hz, and $U_{N,rms}$=98V@f_N=800Hz, for further parameters see Tab.I. The analytically calculated dependencies of the quantities (cf. (1)-(4)) are drawn in dashed lines. For small output power the current of the diodes is discontinuous and the output voltage increases to the peak value of the line-to-line mains voltage $U_{O,max}$= $\sqrt{6}U_{N,rms}$ (281,7V and/or 240V).

$$\hat{I}_{N,(n)} = \frac{1}{n^2 \omega L}\hat{U}_{U,(1)} \tag{6}$$

Accordingly, for limiting the low-frequency current harmonics within the entire input voltage and frequency range to given values, e.g.,

$$\begin{aligned} \hat{I}_{N,(11),r} &= 0.1\hat{I}_{N,(1),r} \\ \hat{I}_{N,(13),r} &= 0.08\hat{I}_{N,(1),r} \end{aligned} \tag{7}$$

(for rated power, cf. [3])

$$L \geq \frac{1}{n^2 \omega_{\min}} \frac{\hat{U}_{U,(1),\max}}{\hat{I}_{N,(n)}} \tag{8}$$

has to be ensured. There, $\hat{U}_{U,(1),max}$ corresponds to the maximum mains voltage and ω_{min} is the minimum mains angular frequency.

In the case at hand we select the minimum value L= 376µH (cf. Tab.I) as the actually occurring current harmonics (cf. **Fig.4**) show lower amplitudes than defined by (7). This is again (as for (4)) caused by the discontinuity of the diode bridge input currents in the vicinity of the zero crossings which results in an autotransformer input voltage of lower low-frequency harmonic content.

As at the outputs (*1a,1b,1c*) and (*2a,2b,2c*) of the autotransformer different zero sequence voltages $u_{1,0}$ and $u_{2,0}$ (cf. Fig.2(a)) are present, a zero sequence voltage does occur across each set of phase windings. Therefore, the autotransformer has to be realized using individual magnetic cores for the phases or a common five-limb core which allows to accommodate a zero sequence magnetic flux. The magnetic flux occurring in an individual core is determined by the difference of the phase voltages at the corresponding outputs, e.g. u_{1aM} and u_{2aM}. Concerning further details of the dimensioning of the system we would like to refer to [5] (cf. Fig.2 in [5]) for the sake of brevity.

In summary, the 12-pulse rectifier is characterized by a total rated power of all magnetic components (autotransformer and input inductors L) of ≈20% of the DC power. The main components to be employed in the power circuit are compiled in **Tab.I**, the losses in the power components are shown in **Fig. 5** (concerning the resulting efficiency see Fig.11).

Compared to active solutions passive rectifier systems show a relatively high overload capacity of the power components which makes such systems especially interesting for applications requiring high peak-to-continuous power demand. If a largely sinusoidal shape of the mains current should be given also for overload conditions, the magnetic design of the input inductors has to ensure that up to maximum output power no magnetic saturation occurs.

One has to note that according to (1) mains current and voltage show an increasing phase displacement with increasing output power (with decreasing output resistance and/or increasing amplitude $\hat{I}_{N,(1)}$ of the mains current fundamental). As can be derived form (2) and/or from a phasor diagram, the output power of the system reaches a maximum for $\hat{I}_{N,(1)} = \hat{U}_N/(\sqrt{2}\omega L)$ and/or a phase displacement of φ=45° (load/source matching)

$$P_{O,\max} = \frac{3}{2}\frac{\hat{U}_N^2}{2\omega L} \tag{9}$$

(cf. **Fig.6**). A further decrease of the load resistance results in an increase of the mains current amplitude, but due to the increasing phase displacement the power delivered to the output decreases further and, therefore, also the output voltage. One has to consider the power limit according to (9) when designing a system with a desired maximum output power. Here, the operation at minimum mains voltage and maximum mains frequency is critical (cf. Fig.6). If low amplitudes of the 11th and 13th mains current harmonics are required the input inductors have to show a high inductance L which means that the desired output power could not be available for certain operating conditions.

III. THREE-PHASE PWM RECTIFIER SYSTEMS

A. Unidirectional Three-Level PWM Rectifier

Three-phase PWM rectifier systems allow the formation of pulse width modulated voltages of sinusoidal local average value at the bridge leg inputs. In combination with the mains voltages this results

TABLE I
MAIN POWER COMPONENTS EMPLOYED IN THE 12-PULSE AUTOTRANSFORMER RECTIFIER.

Part	Quantity	Type	Manufacturer	Typical Data	Package
L	1	Core: S3U 39b / TRAFOPERM N2 0.1mm 31 turns / 0.72 mm air gap	VAC	376 µH / 27.3 A	
$Tr_{a,b,c}$	3	Core: SM55 / TRAFOPERM N2 0.1mm	VAC	w_A/w_B = 34/12 turns	`
C	2	56 µF / 450 V / 105°C / AXF	Rubycon	0.45 A @ 120 Hz	∅25 x 20 mm
D	12	10ETF04	IR	400 V / 10 A	TO220

TABLE II

MAIN POWER COMPONENTS EMPLOYED IN THE 5KW UNIDIRECTIONAL THREE-LEVEL PWM RECTIFIER SYSTEM.

Part	Quantity	Type	Manufacturer	Typical Data	Package
L	3	107µH / 17.8 A_{DC} (33667)	Schott	Helical winding	125 Series
C	10	100µF / 250V / 105°C / YXF	Rubycon	1.2 A / 0.18 Ω	∅18 x 35.5 mm
D_{N-}	3	TYN640 (Thyristor)	ST	600 V / 40 A	TO220
D_{N+}	3	15ETH03	IR	300 V / 15 A / 40 ns	TO220
D_F	6	15ETH03	IR	300 V / 15 A / 40 ns	TO220
S	6	STU26NM50	ST	500 V / 26 A / 0.10 mΩ	TO220

in a sinusoidal mains current and/or no low frequency harmonics do occur. The switching frequency current harmonics can be suppressed by a mains-side filter of relatively small volume.

If no energy feed-back into the mains is required the three-level PWM rectifier system given in Fig.1(b) shows significant advantages compared to a conventional two-level system realization (cf. Fig.1(c)). Besides the potential of the positive and negative DC voltage bus also the neutral point potential is available for voltage formation. The mains current therefore shows a lower switching frequency current ripple, and/or input inductors of smaller value and smaller size can be employed. Furthermore, all power semiconductors face only half of the output voltage resulting in a reduction of the switching losses approximately by a factor of two. Accordingly, this results in an increase of the efficiency of the system, a reduction of the cooling effort and a further increase of the power density.

The output voltage of the system has to be set higher than the amplitude of the line-to-line mains voltage ($\hat{U}_{N,l-l}$ ≈323V for $U_{N,rms}$=132V) and is defined in the following as U_O=350V. The pre-charging of the output capacitors is performed by pre-charging resistors which are short-circuited by thyristors D_{N-} which perform rectifier function for regular operation (cf. Fig.1(b)). A pre-charging relay can be omitted, therefore.

Figure 7 shows the block diagram of a system control proposed in [10] that guarantees operation of the rectifier also in case of a loss of one phase. The total output voltage $u_{C+}-u_{C-}$ (u_{C+} and u_{C-} are defined with reference to the output voltage neutral point M) is compared to the reference voltage U_O*. Dependent on the control error a reference value of the charging current i_C* of the output capacitors if formed. Adding the load current (in case of load current pre-control) results in a reference DC current i* to be fed into the output circuit. Considering the mains voltage condition and a power balance the reference DC current i* can be transformed into a reference amplitude of the mains current which is proportional to the amplitude $\hat{I}_D$ of the switching frequency carrier signal i_D of the current control (cf. of Fig.7). The equal distribution of the total output voltage u_O between the capacitors C_+ and C_- and/or the potential of the output voltage center point M is controlled by an offset i_0 of the three-phase current values resulting in a center point current. Concerning the detailed function of the control scheme we would like to refer to [10] for the sake of brevity.

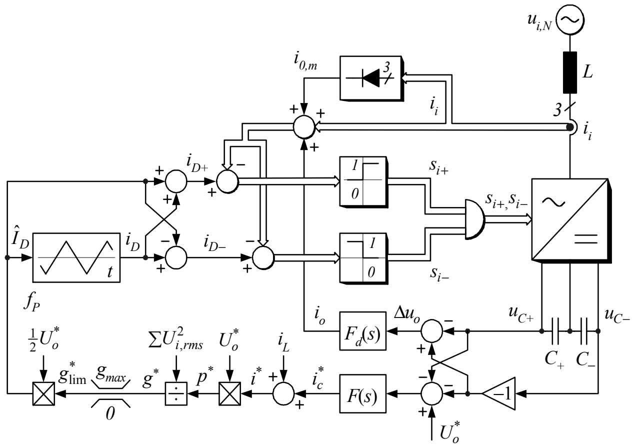

Fig.7: Cascaded control of a unidirectional three-level PWM rectifier system (cf. Fig.1(a)) according to [10]. Signal paths being equal for all phases are combined in double lines.

A practical realization of the control circuit is shown in **Fig.8**. The time-behavior of input phase currents and of the mains currents resulting after single-stage LC-filtering is given in **Fig.9** for a switching frequency of f_P=50kHz (T_P=20µs). This is a compromise between switching losses, sufficient pulse number within the mains period at maximum mains frequency $f_{N,\max}$=800Hz (T_N= 1250µs) and volume of the input side inductors L.

The selection of the power components given in **Tab.II** can be derived from the analytical expressions of the current stress of the components given in [11]. The resulting distribution of the system losses at rated power are compared to a conventional PWM converter system (cf. Fig.1(c)) in Fig.10.

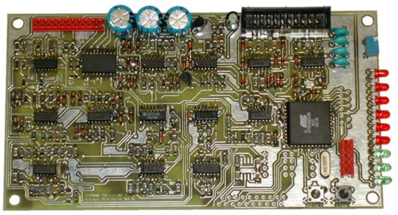

Fig.8: Control board hardware, overall dimensions: 75mm x 140mm (3in x 5.5in) as developed at the Power Electronic Systems Laboratory of the ETH Zurich.

B. Two-Level PWM Rectifier

PWM rectifier systems for industrial drives are generally realized as two-level topologies (cf. Fig.1(c)) where the pulse width modulated input voltage is formed by switching between positive and negative output voltage bus. This results in a relatively high blocking voltage stress of the power semiconductors defined by the total output voltage that is again chosen as U_O=350V. Because of the relatively high reverse recovery time of the internal diodes of the power MOSFETs one has to provide explict free-wheeling diodes for high switching frequencies (cf. Fig.1(c)). Furthermore, a relay short-circuiting the pre-charging resistors in regular operation has to be employed for two-level system.

Concerning the output voltage and mains current control and the resulting time-behavior of the input and mains currents there are no basic differences of the two-level and the three-level PWM rectifier system discussed in section III.A except the higher switching frequency harmonics of the two-level system. We therefore would like to omit a detailed discussion for the sake of brevity.

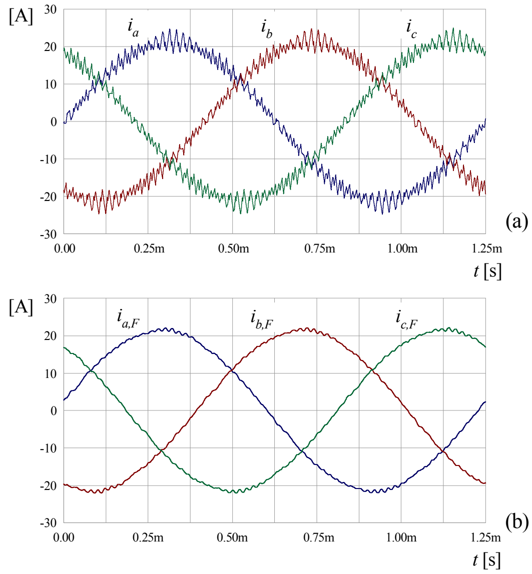

Fig.9: Simulated time behavior of the rectifier input phase currents i_i, $i=R,S,T$ and **(b)** of the mains currents $i_{N,i}$ resulting after single-stage LC-filtering (damping resistor connected in parallel with the filter inductor in order to provide passive filter damping). To fulfil the requirements concerning EMI for a practical realization generally a multi-stage input filter has to be employed. Furthermore, we want to point out that the output voltage of active three-phase PWM rectifier systems shows a common-mode voltage component with switching frequency. Simulation parameters: $U_{N,rms}$=115V, f_N=800Hz, U_O=350V, switching frequency f_P=50kHz.

For the dimensioning of the power components analytical expressions for the current stresses on the components are given in [12]. The selected components are listed in **Tab. III**.

The comparison of the loss distributions between the two-level and the three-level system given in **Fig.10** clearly shows the

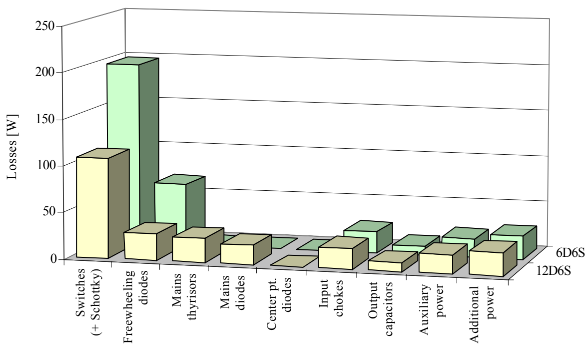

Fig.10: Distribution of the system losses for the unidirectional three-level PWM rectifier system (comprising 12 diodes and 6 switches, 12D6S, cf. Fig.1(a)) and for the bidirectional two-level PWM rectifier system (comprising 6 switches and 6 antiparallel diodes, 6D6S, cf. Fig.1(c)). Assumed operating parameters: input voltage $U_{N,rms}$=115V, U_O=350V, P_O=5kW, f_P=50kHz.

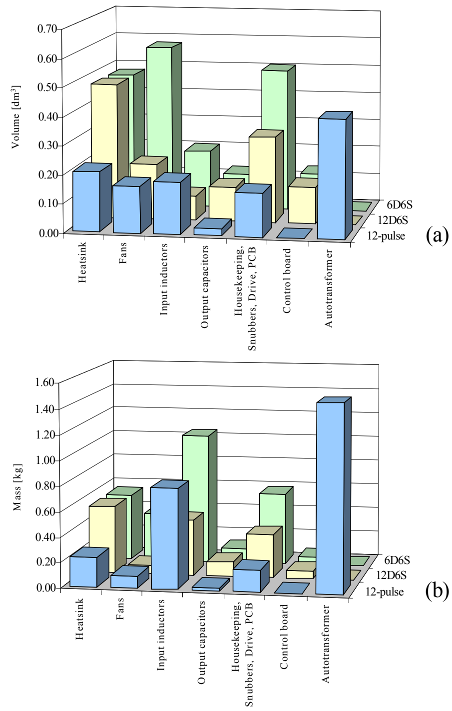

Fig.11: Comparison of volume (cf. **(a)**) and weight (cf. **(b)**) of the main components of the rectifier concepts shown in Fig.1 (12-pulse rectifier, cf. Fig.1(a); 12D6S, cf. Fig.1(b), and 6D6S, cf. Fig.1(c)); rated power: P_O=5kW.

disadvantages of the conventional realization employing two-level technology. Higher turn-off voltages of the power transistors and a higher reverse recovery time of the diodes with higher blocking capability result in higher switching losses and, therefore, in a significantly reduced overall system efficiency (cf. Fig.12).

IV. COMPARISON OF CONVERTER CONCEPTS

A comparison of the rectifier systems concerning volume (cf. **Fig.11, Tab. IV**) identifies a minor advantage of the 12-pulse system even in comparison to the three-level PWM rectifier (denoted as 12D6S in the following). On the other hand the weight of the 12D6S converter is considerably lower than for the passive 12-pulse system as the heatsink constitutes a considerable share of the total converter volume but shows a low specific weight (cf. **Tab. V**) compared to the magnetic components dominating the 12-pulse system.

The two-level rectifier system (6D6S) exhibits disadvantages concerning the power density due to the high cooling effort resulting from the high switching losses. The overall system weight is about equal to the weight of the 12-pulse rectifier.

Due to the missing switching losses and the low number of power semiconductor forward voltage drops in the main current paths the efficiency of the 12-pulse system is significantly higher than for the

TABLE III
MAIN POWER COMPONENTS EMPLOYED IN THE 5KW BIDIRECTIONAL TWO-LEVEL PWM RECTIFIER SYSTEM.

Part	Quantity	Type	Manufacturer	Typical Data	Package
L	3	196µH / 21.7Adc (33488)	Schott	Helical winding	168 Series
C	10	100µF / 250V / 105°C / YXF	Rubycon	1.2 A / 0.18 Ω	∅18 x 35.5 mm
D_F	6	15ETX06	IR	600 V / 15 A / 18 ns	TO220
D_S	6	20L15T	IR	15 V / 20 A	TO220
S	6	SPP20N60C3	Infineon	600 V/ 190mΩ	TO220

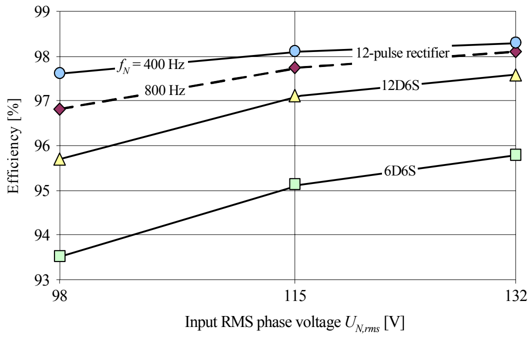

Fig.12: Comparison of the power conversion efficiencies of the rectifier concepts shown in Fig.1 (denomination as in Fig.11) for different levels of the input voltage $U_{N,rms}$=98, 115, 132V; assumed operating frequency: f_N=400Hz, rated power: P_O=5kW.

active rectifier systems (cf. **Fig.12**). The relatively low efficiency of the 6D6S system is mainly due to the high switching losses (cf. Fig.10).

The input current of the active rectifier systems contains switching frequency components which must be attenuated by a multi-stage EMI filter in order to ensure compliance to EMI standards. Besides differential mode filtering there special attention has to be paid to the common mode filtering which prevents common mode currents occurring due to the switching frequency common mode component of the rectifier output voltage and the parasitic capacitive coupling of the power semiconductors and the DC side components to the heatsink and/or the safety ground. In contrast, for the 12-pulse system only a low-frequency common mode output voltage component with three-times the mains frequency is present which does not require a special filtering.

For the 12-pulse system the dependency of the output voltage on the load and mains condition has to be seen as a main drawback besides the high system weight. For no-load operation the output voltage is determined by the amplitude of the line-to-line mains voltage, in the case at hand $U_{O,max}$=323V (for $\hat{U}_N$=132V), and decreases to $U_{O,min}$=233V (cf. Fig.3) for rated load and maximum mains frequency f_N=800Hz. This voltage variation has to be accommodated in the design of the supplied PWM inverter and/or electric machine. For the active rectifier systems an output voltage of U_O=350V is ensured independent of the load condition.

A further disadvantage could be the emission of acoustic noise dependent on the employed magnetic material and the mechanical construction of the input inductors and the autotransformer.

In summary, the three-level rectifier system (12D6S) represents the most advantageous solution concerning volume, weight and functionality as it allows to supply $1/\sqrt{3} \approx 50\%$ of the rated power at sinusoidal current also in case of a failure of a mains phase, i.e. for two-phase operation.

The weight of the 12-pulse system could be reduced for replacing the three individual magnetic cores of the autotransformer by a five-limb core. In connection with advantages of the system concerning complexity/reliability and robustness/overload capacity this motivates a further consideration of the system. There, the main aim is an extension of the circuit topology allowing a control of the output in order to eliminate the main drawback compared to active systems.

TABLE IV
TOTAL VOLUME, TOTAL WEIGHT AND OVERALL SPECIFIC WEIGHT OF THE RECTIFIER CONCEPTS SHOWN IN FIG.1. FOR ALL SYSTEMS A RATED POWER OF P_O=5KW HAS BEEN ASSUMED.

System	Mass [kg]	Volume [dm^3]	Specific weight [kg/dm^3]
12-pulse rectifier	2.8	1.14	2.5
12D6S	1.6	1.3	1.2
6D6S	2.7	2.0	1.4

TABLE V
TYPICAL SPECIFIC WEIGHTS OF POWER COMPONENTS OF PWM RECTIFIER SYSTEMS.

Component	Specific weight [kg/dm^3]
Inductor (Schott 168 Series)	5.0
Electrolytic capacitor (450V)	3.0
Control board	1.2
Heatsink	1.2
Fan	0.7

For the sake of completeness a novel 12-pulse rectifier topology with controllable output voltage will be introduced and briefly discussed in the following in combination with results of a first experimental analysis. Concerning a more detailed analysis of the system we would like to refer to a future publication being currently under preparation.

V. 12-PULSE RECTIFIER WITH CONTROLLED OUTPUT VOLATGE

A controllability of the output voltage of a 12-pulse rectifier system (Fig.1(a)) can be achieved by extending the basic topology as given in **Fig.13**. There, the system shows the functionality of a boost-converter. According to the duty cycle of the transistors T_1 und T_2 (which are preferably switched in an interleaved manner) the output voltage of the diode bridges and, therefore, the input voltage of the autotransformers is reduced. Accordingly, a higher DC voltage U_O is required for balancing the mains voltage. For guaranteeing a symmetric distribution of the mains current to the partial systems a zero-sequence current control has to be employed which is not discussed here for the sake of brevity.

First experimental results of the rectifier are shown in Figs.13(b) and (c). With the exception of a ripple with switching frequency the shape of the mains current is identical to a passive 12-pulse rectifier. Also in the shape of the input voltage of the autotransformer the typical 12-pulse waveform is obvious.

Furthermore, we want to point out that no sinusoidal PWM has to be performed and/or the transistors are controlled with essentially time-constant duty-cycle what facilitates a very simple realization of the control and the transistor gate driver circuits. It is important to note that the output voltage control ensures a constant output voltage value but does not increase the power limit $P_{O,max}$ of the system (cf. (9)).

VI. CONCLUSIONS

Passive rectifier systems for operation in the public mains, i.e. for a mains frequency of 50/60Hz show a considerably higher volume than active rectifier systems of equal power. However, as shown in this paper, an increase of the mains frequency by a factor of only 10 already results in an approximately equal power density (W/dm^3) of both concepts. Therefore, the selection of the rectifier concept for supplying the PWM inverter stage of a EHA to be employed in future

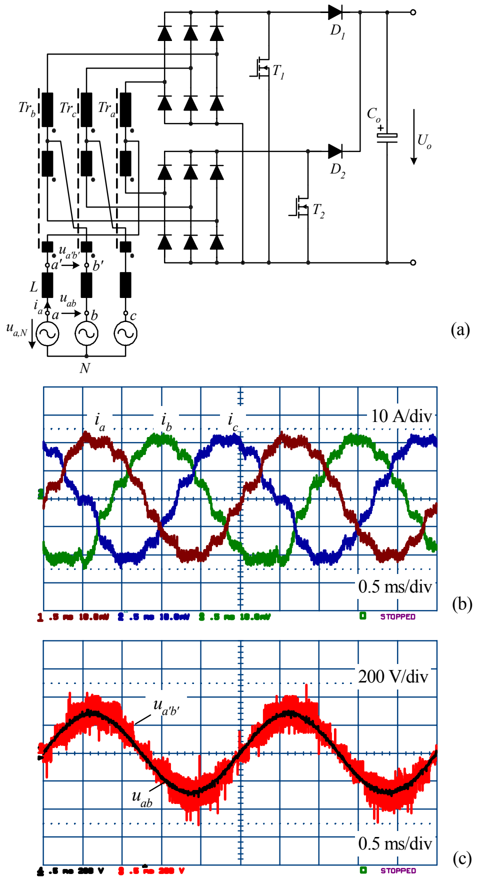

Fig.13: Experimental analysis of an extension of the passive 12-pulse rectifier system to controlled output voltage (cf. **(a)**). The power transistors are operated in interleaved manner, the boost inductor is formed by the input inductors L and partly by the stray inductance of the interphase transformer. Furthermore shown: **(b)** time behavior of the mains phase currents; **(c)** autotransformer input line-to-line voltage $u_{a'b'}$ and related mains line-to-line voltage u_{ab}. Operating parameters: $U_{N,rms}$=115V, f_N=400Hz, U_O=350V, P_O=4.85kW, switching frequency f_P= 33kHz, efficiency η=95%.

more electric aircrafts can be based mainly on functionality and complexity/reliability.

Active rectifier systems are characterized by a controlled output voltage and a controlled sinusoidal shape of the input current and do allow a two-phase operation. However, one has to accept a relatively complex structure of the power and control circuit. Main aspects of a practical realization are therefore the integration of the power semiconductors into a multi-chip power module and a fully digital system control.

Considering the continuous progress in power semiconductor technology active systems show the potential of a further increase of the power density with increasing switching frequency and/or decreasing volume of the passive components.

Passive 12-pulse rectifier systems are characterized by a low complexity / high reliability even for extension to controlled output voltage. As a first experimental analysis shows the efficiency of a controlled 12-pulse system is comparable the three-level active rectifier. However, asymmetries and harmonics of the mains voltage still take influence on the mains current quality. Also the disadvantage of a high system weight will not be reduced significantly in future due to missing progress in the development of new magnetic materials.

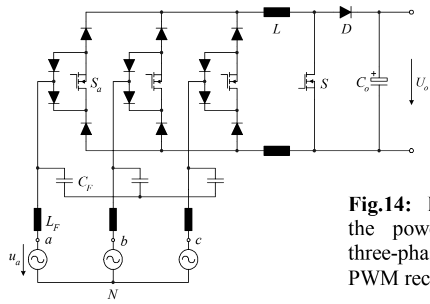

Fig.14: Basic structure of the power circuit of a three-phase buck+boost PWM rectifier [13].

In summary the three-level PWM rectifier constitutes the most advantageous solution and should be considered as a reference for a comparison of further rectifier concepts. There, e.g., the performance of the three-phase buck+boost PWM rectifier topology [13] shown in **Fig.14** is of special interest as it allows a direct start-up without output capacitor pre-charging and a direct limitation of the load current in case of an output short circuit and therefore ensures a high system reliability.

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