A Novel Concept for Regenerative Braking of PWM-VSI Drives Employing a Loss-Free Braking Resistor

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Abstract. In this paper a simple circuit for feeding back the braking energy from the DC voltage link of a three-phase PWM AC drive system into the mains is proposed. The operating behavior of the system which consists of a series connection of a line-commutated three-phase bridge and a turn-off power semiconductor device and in principle does not require passive components is discussed. The stationary operating behavior is analyzed by analytical calculations and by digital simulation and represented in graphical form. Furthermore, the effects on the mains caused by the system for braking operation are analyzed and compared to the conditions for motoring operation and/or to the conditions for feeding of the DC link by an uncontrolled rectifier bridge. Finally, measures for guaranteeing high system operational safety are discussed and the advantages and disadvantages of the circuit are compiled in form of an overview.

1 Introduction

For feeding the DC voltage link of pulse width modulated drive systems in many cases three-phase diode bridge circuits are applied. According to the required DC link voltage this circuits are connected to the mains either directly or via transformers. A substantial limitation of this concept, which is advantageous especially regarding economy, robustness, reliability and power density (W/cm^3) is given by the unidirectional structure of the feeding circuit. This makes feeding back of energy from the DC link into the mains impossible. (Another drawback are the relatively high effects on the mains.) For four-quadrant drive systems one has to provide, therefore, an extension of the circuit for handling the energy being fed back into the DC link by the machine-side converter, e.g., for lowering the load in crane or elevator drives [1] or for braking of large inertias (e.g., for centrifugal drives).

If the braking operation and/or the feedback of energy is limited to short dynamic processes and/or if the drive system has a relatively low rated power, the braking energy can be converted into heat in a pulsed braking resistor which is connected in parallel to the DC link capacitor (cf. **Fig.1**(a) and/or p. 421 in [2]).). If the DC link voltage exceeds (due to the load-side energy feedback) a preset maximum value, the resistor R_B is switched on resulting in decreasing DC link voltage. If the control of the pulse operation of R_B is realized as on-off-control, the braking resistor is turned off again when the DC link voltage crosses the lower switching threshold. Therefore, for continued energy flowback R_B is switched on and off in a repetitive manner. The economic limit of this technique is reached for drives of high power and/or for frequent braking operation of a drive system. Essential points to be considered here are (i) the additional cooling effort due to the heat caused by the losses, (ii) the reduction of the power density due to the physical size of the braking resistor (dimensions of a braking resistor for 10kW rated power: 74x48x30cm^3 (29x19x12in^3), weight 32kg for safety class IP00) and, in general, (iii) the higher energy cost caused by the loss of the braking energy.

If the DC link feeding circuit is realized in bidirectional form, the braking energy can be fed back into the mains. Basic concepts of bidirectional mains converter systems are shown in Figs.1(b) and (d).

If the diode bridge circuit of Fig.1(a) is replaced by a fully controlled three-phase bridge circuit, one can obtain a reversal of the power flow (i.e., a feedback of the braking energy into the mains) by shifting the phase control angle. However, due the fact that the thyristor bridge reverses the output voltage polarity for inverter operation as compared to rectifier operation (and/or because the current direction is given by the valve direction) we have to apply the DC link voltage in reversed polarity to the output of the thyristor bridge by a DC-to-DC converter connected in series in this case (cf. Fig.1(b) and/or [3], [4]). Furthermore, this DC-to-DC converter stage provides a matching of the DC link voltage to the output voltage level of the thyristor bridge (which is lower for inverter operation as compared to the rectifier operation). Alternative possibilities of realization of the circuit according to Fig.1(b) are described in [5], [6] and [7]. Properties characterizing this class of systems in general are (besides the possibility of a reversal of the power flow) a rectangular shape of the mains current and controlled DC link voltage.

If the feeding of the DC link via a diode bridge is maintained, the feedback of braking energy into the mains can be performed via an anti-parallel thyristor bridge circuit operating as inverter (cf. Fig.1(c) and/or Fig.3 in [1]). There, besides a smoothing inductor L, for connecting the feedback converter to the mains a matching transformer is required in order to avoid commutation failures. A mains voltage-independent control of the DC link voltage is not possible in this case. The mains current shows a rectangular shape being dependent on the size of L. The matching transformer can be omitted if the diode bridge is replaced by a thyristor bridge and if the system is operated with reduced DC link voltage (cf. p. 495 in [2], or Fig.3 in [8]).

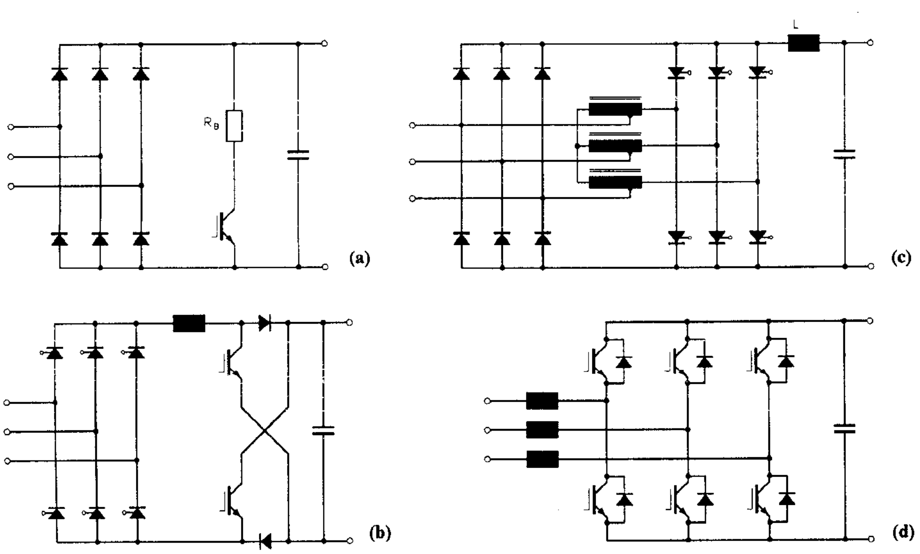

Fig.1: Conversion of the braking energy for conventional feeding of the voltage DC link of PWM AC drive systems. **(a):** unidirectional, unipolar feeding circuit (three-phase diode bridge), conversion of the braking energy into heat loss (braking chopper). **(b):** feedback of braking energy into the mains by a unidirectional, bipolar feeding circuit; inverter operation and polarity reversal of the output voltage of the thyristor bridge for braking operation; rectangular mains current shape, controlled output voltage. **(c):** bidirectional, unipolar anti-parallel connection of mains-controlled rectifier systems with matching transformer of the inverter stage feeding back the braking energy into the mains. **(d):** bidirectional PWM rectifier system with fixed polarity of the output voltage, feedback of the braking energy with sinusoidal mains current shape; controlled output voltage.

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A bidirectional feeding circuit which is characterized by sinusoidal mains current consumption, adjustable mains current phase angle and controlled output voltage is shown in Fig.1(d) (cf., e.g., [9] or p. 499 in [2]). The structure of the power circuit is largely identical to the structure of the machine-side power converter. Therefore, the technical advantages being mentioned here are paid for by a relatively high realization effort.

In general one has to point out (as is clearly shown by Figs.1(b)–(d)) that the requirement of feeding back of braking energy into the mains leads to a substantial increase of the complexity of the converter system (as compared to simple diode rectification and dissipative braking) if known circuit concepts are applied.

A substantial property of the circuit according to Fig.1(a) (besides the simple structure and the simple control) is especially the lack of inductive components (smoothing inductances and/or a matching transformer). This advantage is paid for by relatively high effects on the mains (due to utilizing the inner mains impedance for current limitation) which are tolerated, however, in many cases due to the economical advantages, the robustness and the presently not existing regulations. Basically, now one has to ask the question whether (in analogy to a simple diode *rectification*) also an energy *feed-back* into the mains under basic avoidance of additional inductive components can be performed. The circuit resulting from these considerations is shown in **Fig.2** and will be analyzed in detail in this paper. The functional principle and the operational behavior of the new converter concept will be discussed in **section 2**. Within a $\frac{\pi}{3}$-wide interval of the mains period the resulting conduction states of the system are being obtained via the analysis of the trajectory of the mains current space vector. Furthermore, results of a digital simulation of the stationary converter operation are shown. The topic of **section 3** will be the analytical calculation and graphical representation of the dependency on the operational parameters of the power fed back into the mains. Furthermore, the current stresses on the power semiconductors are given. In **section 4** a simple control-oriented model as well as the transfer function of the system will be derived. In **section 5**, the mains behavior of the system is analyzed; the time dependency and the spectrum of the mains current as well as its fundamental phase displacement and the total harmonic distortion are compared to the properties of simple diode rectification. Finally, in **section 6** the advantages and disadvantages of the system are compiled in an overview. Furthermore, modifications of the basic structure regarding an improvement of the operational safety of the system are proposed and an outlook concerning the future treatment of the topic is given.

Remark: The basic thought of feeding back braking energy from the DC link into the three-phase mains with basic avoidance of passive components has been described already in [10]. The circuit given there shows three turn-off power semiconducters besides a thyristor bridge, however. Furthermore, we want to point out a circuit proposed in [11] which can be thought to be formed by a combination of the circuits according to Fig.1(a) and Fig.2 and an inductor on the DC side for limitation of the feeding-back current. This circuit also allows a braking operation of the drive in cases of mains failure.

2 Theory of Operation

2.1 Control and Conduction-States of the Converter System

The supply of the DC link from the three-phase mains is realized via an uncontrolled three-phase diode bridge in the circuit shown in Fig.2. There, always only the two phases forming the maximum line-to-line voltage participate in the current flow (with the exception of the commutation intervals). The current flow within the conduction intervals is defined by the small difference between line-to-line mains voltage and DC link voltage appearing across the mains source impedances of these phases.

Due to the lack of passive components in the inverter part of the circuit, now also for the feed-back of braking energy into the mains (for DC link voltage being higher than the mains peak value) the current limiting property of the mains source impedance can be used and/or a conduction of the thyristors similar to the diode rectification concerning the on-intervals of the valves has to be performed. **Figure 3** shows the firing diagram of the inverter part which follows from this considerations and which summarizes the control commands of the thyristors and of the power transistor. By firing with a displacement of $\frac{\pi}{3}$ of two valves of the thyristor bridge in each case (e.g., T_5 and T_6 in $\varphi_{N,1}$) always the maximum line-to-line voltage (in the case at hand $u_{N,TS} = u_{N,T} - u_{N,S}$) will be switched to the mains and/or the thyristor bridge is being operated in the inverter mode at full control angle. If now in $\varphi_{N,1}$ also the power transistor S_1 is turned on, the DC link voltage U_O being higher than the mains voltage (due to the feed-back of braking energy into the DC link) a current flow against the mains voltage is formed, i.e., energy is fed back into the mains (cf. **Fig.4**). Such as in rectifier operation the current flow into the mains is limited by the mains source impedances L_i. The then given conduction state of the feed-back unit is shown in **Fig.5**(a).

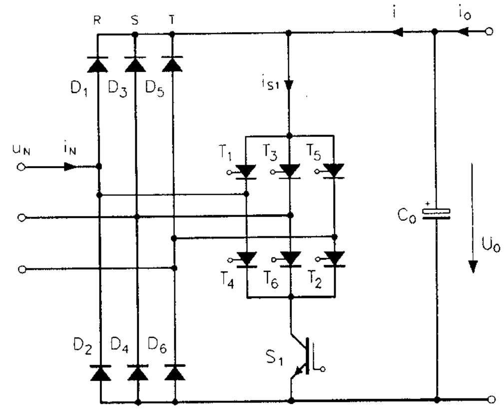

Fig.2: Basic structure of the power circuit of the proposed bidirectional system for feeding the voltage DC link of a PWM AC drive system. The circuit can be thought to be formed via replacing of the braking resistor in Fig.1(a) by a three-phase thyristor bridge. The braking energy is not converted into heat loss but fed back into the mains. The structural similarity of the circuits according to Figs.1(a) and 2 shall be pointed out by calling the thyristor bridge a Loss-Free Braking Resistor. Remark: The term Loss-Free Resistor has been introduced by S. Singer in [12] as denomination of power electronic systems with resistive input characteristic and largely loss-free transfer of the input power to the output. In the case at hand the application of this term shall point out (besides the freedom of loss) especially a structural similarity of converter systems.

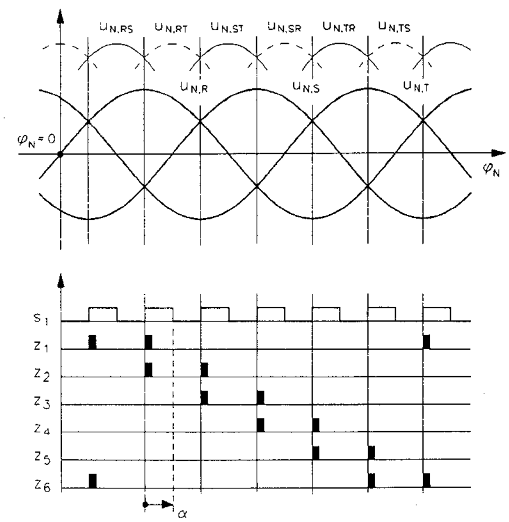

Fig.3: Firing pulses z_j of the thyristors T_j and control signal s_1 of the power transistor S_1 of the circuit shown in Fig.2 with respect to the mains phase source voltages $u_{N,i}$, $i = R, S, T$ and sections of the line-to-line mains phase voltages taking part in the current conductance in each case.

By turning off S_1 in $\varphi_{N,2}$ the current flow i from the DC link is interrupted. The current on the mains side being impressed by the mains source impedance L_i now is carried by a leg of the thyristor bridge (thyristor T_5) and by a diode of the positive bridge half of the rectifier bridge (diode D_5) lying antiparallel to the thyristor bridge. The current is being decreased against the mains voltage.

However, as a closer analysis shows, now also the phase R (which does not carry current in the on-state interval of S_1) starts to carry current if $\alpha > \frac{\pi}{6}$ is valid ($\alpha = \varphi_{N,2} - \varphi_{N,1}$). For a clear representation of the then relatively involved conditions an inspection of the trajectories of the space vector $\underline{i}_N$ of the mains phase currents and of the space vector $\underline{u}_N$ of the mains phase voltages proves to be advantageous (cf. **Fig.6**). As shown in Fig.6(a), the mains voltage space vector covers the angle region $(\varphi_{N,1}, \varphi_{N,2})$ within α. There, the mains current $\underline{i}_N$ is being built up against the mains voltage in direction of the axis $i_{N,R} = 0$ ($\varphi_N = 0$) due to the dominance of the space vector $\underline{u}_{U,5}$ formed by the feed-back unit.

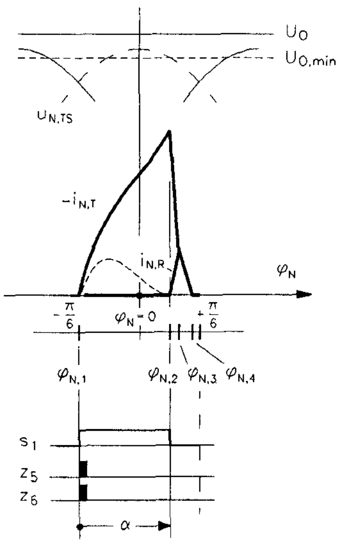

Fig.4: $\frac{\pi}{3}$ wide section of the shape of the mains phase currents $-i_{N,T}$ and $i_{N,R}$, of the line-to-line voltage $u_{N,TS} = u_{N,T} - u_{N,S}$ and of the DC link voltage U_0; in order to allow a clear representation, the demagnetization of the mains source inductances L_i (interval $(\varphi_{N,2}, \varphi_{N,4})$) are shown in an extended time scale. Furthermore shown are: transistor control signal s_1 and firing signals z_5 and z_6 of the thyristors T_5 and T_6; $U_{O,\min} < \hat{U}_N$ defines that DC link voltage for which $-i_{N,T}$ obtains zero in $\varphi_{N,2}$ (limit of validity of the mathematical analysis of the system behavior as given in section 3);

Remark: The complex space vector $\underline{u}_N$ associated to a triple of phase quantities $u_{N,R}, u_{N,S}$ and $u_{N,T}$ is obtained by a transformation $\underline{u}_N = \frac{2}{3}(u_{N,R} + \underline{a}\, u_{N,S} + \underline{a}^2 u_{N,T})$ with $\underline{a} = \exp j\frac{2\pi}{3}$. Therefore, the back-transformation of a space vector into phase quantities can be performed simply by a projection onto the phase axes R, S and T (cf. Fig.6(c)). If sinusoidal symmetrical mains phase source voltages $u_{N,R}, u_{N,S}$ and $u_{N,T}$ are assumed there follows for the space vector of the mains voltage $\underline{u}_N = \hat{U}_N \exp j\varphi_N$. There, $\varphi_N = \omega_N t$ is the rotational angle, ω_N the constant angular velocity and $\hat{U}_N$ the amplitude of the phase voltages. The voltage space vector $\underline{u}_U$ formend at the output of the energy feed-back unit is determined by the switching state of the thyristor bridge (for power semiconducters being in on-state condition). $\underline{u}_{U,1}$ denotes the voltage space vector (not shown in Fig.6(c)) given for S_1, T_5 and T_6 being turned-on. For generating the mains current within α the projection of the voltage difference $\underline{u}_N - \underline{u}_{U,1}$ onto the axis $i_{N,R} = 0$ ($\varphi_N = 0$) is in effect for the angle interval considered.

For $\varphi_N > \varphi_{N,2}$ we have $\underline{u}_U = 0$ due to the free-wheeling of $\underline{i}_N$ via D_5 and T_5 as described before. Therefore, the current decrease results in the direction of $\underline{u}_N \approx \underline{u}_N|_{\varphi_N = \varphi_{N,2}} = \underline{u}_{N,\varphi_{N2}}$ as shown in Fig.6(c). (For the demagnetization of the inductances L_i the full amount $\hat{U}_N$ of the mains voltage space vector becomes effective; due to the therefore relatively short duration of the demagnetization the rotation of $\underline{u}_N$ can be neglected in a first approximation.) Because $\underline{u}_{N,\varphi_{N2}}$ has a component $u_{N,R} > 0$ and because the diode bridge (diode D_1) permits a current flow $i_{N,R} > 0$, now also the phase R participates in the current conduction, as already mentioned (cf. Fig.5(b)). If now $\underline{i}_N$ reaches the axis $i_{N,S} = 0$ in $\varphi_{N,3}$, the current in phase S becomes 0; because D_5 does not permit a sign reversal of $i_{N,S}$, only the phases R and T remain conducting for $\varphi_N > \varphi_{N,3}$ (cf. Fig.5(c)). For the decrease of the current $i_{N,R} = -i_{N,T}$ we only have the component $u_{N,\varphi_{N3}} \approx u_{N,\varphi_{N2}}$ (which is small in comparison to $\hat{U}_N$ being effective within $(\varphi_{N,1}, \varphi_{N,2})$) in the direction of the axis $i_{N,S} = 0$. With this a reduction of the rate of current change is connected and the characteristic shape of $-i_{N,T}$ within $(\varphi_{N,3}, \varphi_{N,4})$ results. After complete demagnetization of the inductances L_i in $\varphi_{N,4}$ the valves do not carry current until S_1 is turned-on again in $\varphi_N = +\frac{\pi}{6}$. The recovery time of the thyristor T_5 (typically 100…200$\,\mu$s), being stressed by a (forward) blocking voltage when T_1 and T_6 are turned on in $\varphi_N = +\frac{\pi}{6}$ is determined by the angle interval $\beta = (\frac{\pi}{6} - \varphi_{N,4})$.

Based on the previous considerations we can now explain immediately also the shape of the trajectories of $\underline{i}_N$ resulting for $\alpha < \frac{\pi}{6}$ and $\alpha = \frac{\pi}{6}$ in $\varphi_N = (+\frac{\pi}{6}, -\frac{\pi}{6})$. If $\alpha < \frac{\pi}{6}$ is valid, $\underline{u}_N$ has a negative component $u_{N,R} < 0$ in $\varphi_{N,2}$; then no current increase in phase R is possible. Therefore, phase R remains without current within the demagnetization interval, the trajectory being characteristically of triangular shape for $\alpha > \frac{\pi}{6}$ does not exist and is replaced by a straight line. $\alpha = \frac{\pi}{6}$ is a limiting case; $u_{N,R}$ becomes 0 in $\varphi_{N,2}$ here; the shape of the trajectories coincides with that of $\alpha < \frac{\pi}{6}$, therefore.

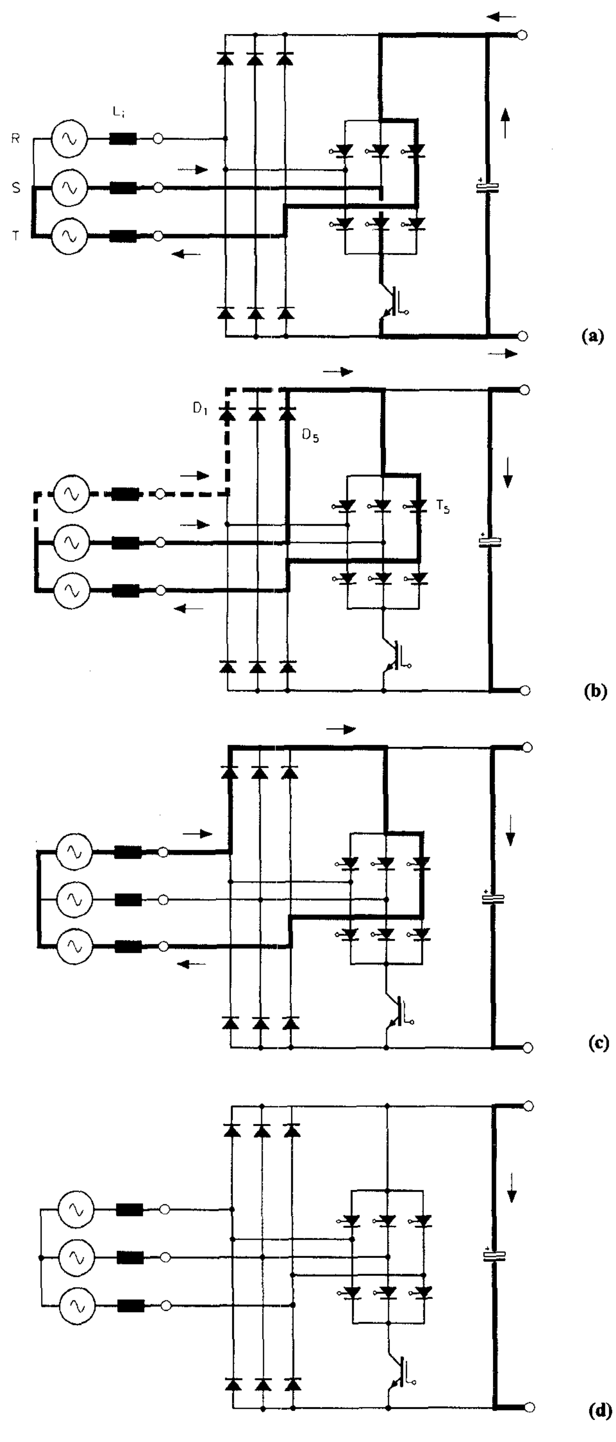

Fig.5: Conduction states of the converter system associated to the section $\varphi_N \in (-\frac{\pi}{6}, +\frac{\pi}{6})$ of the mains period shown in Fig.4; the feeding mains is replaced by ideal phase voltage sources $u_{N,i}$, $i = R, S, T$ (mains source voltage) and by the source inductances L_i. Within the turn-on interval $\alpha = \varphi_{N,2} - \varphi_{N,1}$ of S_1 (cf. **(a)**) the power being taken from the DC link by the feed-back unit is larger than the average braking power $P_{O,\mathrm{avg}}$ as being delivered from the load to the DC link (the DC link capacitor C_O is discharged); in the stationary case C_O is recharged again to the voltage u_O given in $\varphi_{N,1}$ during the off-interval $(\varphi_{N,2}, \frac{\pi}{6})$ of S_1 (cf. **(b)**-**(d)**).

So far the analysis of the operating principle of the circuit has been limited to the angle interval $\varphi_N \in (-\frac{\pi}{6}, +\frac{\pi}{6})$. As Fig.6 shows, we would have to consider a $\frac{2\pi}{3}$-wide angle interval in any case for a complete analysis of the system behavior within a mains fundamental period, due to the asymmetrical system structure. Because with this no basically new findings would follow, we want to omit this for the sake of clearness and brevity.

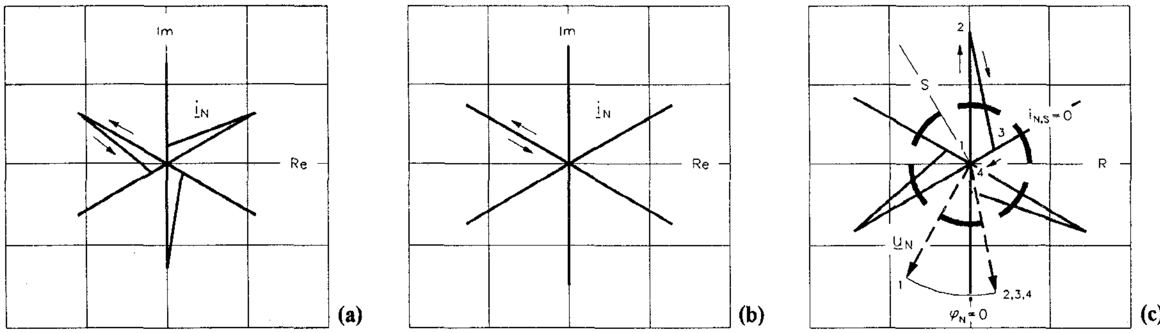

Fig.6: Characteristic shapes of the trajectories of the mains current space vector $\underline{i}_N$ within a mains period for $\alpha < \frac{\pi}{6}$ (cf. **(a)**), $\alpha = \frac{\pi}{6}$ (cf. **(b)**) and $\alpha > \frac{\pi}{6}$ (cf. **(c)**); in (c) that branch of the trajectory is pointed out which is passed through within the section of the mains period shown in Fig.4. Also, the positions of the space vector $\underline{u}_N$ of the mains source voltage for the turn-on and turn-off instants ($\varphi_{N,1}$ und $\varphi_{N,2}$) of S_1 are shown. For numbering of consecutive points of the trajectory $\underline{i}_N$ only the indices i of the angle values $\varphi_{N,i}$ are given. The conducting intervals of S_1 within a mains period are illustrated by circle segments.

2.2 Digital Simulation

For illustrating the stationary operating behavior of the system **Fig.7** shows the mains phase current $i_{N,R}$ (cf. (a)) determined by digital simulation, furthermore the transistor current i_{S_1} (cf. (b)) as well as the time relationship of these quantities to the phase voltage $u_{N,R}$ and to the line-to-line voltages participating in the mains current flow in each case. Furthermore, the current shape and the shape of the blocking voltage of a thyristor of the positive and of the negative bridge halves of the energy feed-back unit (T_1 and T_5, cf. Fig.7(c) and (d)) given.

The simulation is based on the following operating and device parameters:

$$\begin{aligned} U_{N,\mathrm{rms}} &= 230\,\mathrm{V} \\ U_O &= 575\,\mathrm{V}(M = 1.025) \\ P_O &= 50\,\mathrm{kW}(\text{rated power}) \\ L &= 250\,\mu\mathrm{H}(x_L = 0.025) \\ \alpha &= 40^\circ \end{aligned}$$

($U_{N,\mathrm{rms}}$ denotes the rms value of the mains phase voltage, P_O the power being fed back from the DC link into the mains). The inductance L has to be considered as a series connection of the mains source inductance $L_i = 50\ldots100\mu$H (typical values for the European low-voltage mains) and of an explicite series inductance L_N. L_N also has to be provided for a pure rectifier operation for limiting the peak current stress on the rectifier diodes, the current stress (and/or the life time) of the DC link capacitor and the effects on the mains. (According to Fig.3 in [13] a short circuit impedance of $x_L = 0.02\ldots0.03$ is advantageous to select).

As shown in Fig.7(a) the shape of $i_{N,R}$ has a pronounced similarity to the current shape given for rectifier operation (but with opposite sign) according to the control of the feed-back unit being similar to the diode rectifier bridge concerning the on-intervals. Due to the fact that the diodes D_1, D_3 and D_5 are lying in antiparallel to the valves T_1, T_3 and T_5, no reverse blocking voltage stress exists for the thyristors of the positive bridge half. (For circuit design we have to consider, therefore, the increased recovery time in comparison to the case with reverse blocking voltage.) We have to point out that the shape of the forward blocking voltage appearing after the reverse blocking interval β (cf. Fig.4) has a relatively flat rate of rise; the stress on the thyristors with a steep voltage rate of rise in forward direction takes place only after the end of the conduction interval α of S_1 following β; therefore, actually the angle interval $\beta + \alpha$ serves the thyristors for regaining the forward blocking capability.

3 Mathematical Analysis

After the analysis of the basic system function we now want to describe the sta-

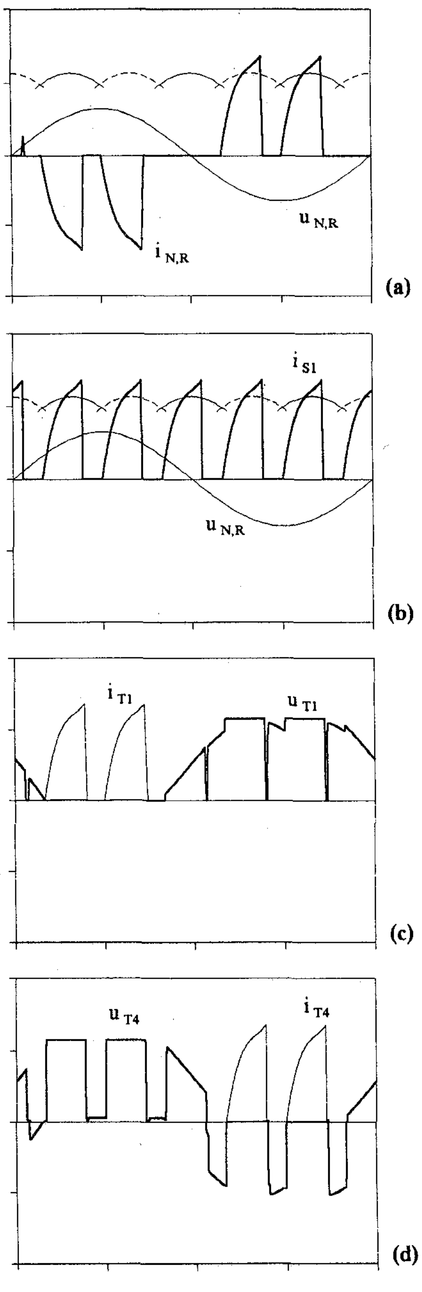

Fig.7: Simulation of the stationary operation of the feed-back unit. Representation of the time behavior within a mains period of the phase voltage $u_{N,R}$, the associated phase current $i_{N,R}$ (cf. **(a)**) and the transistor current i_{S_1} (cf. **(b)**) as well as of sections of the line-to-line voltages participating in the current conduction in each case. Furthermore shown are current and blocking voltage of the thyristors T_1 (cf. **(c)**) and T_4 (cf. **(d)**); scales: 100A/div, 500V/div.

tionary operating behavior of the energy feed-back unit also mathematically. The aim is to determine the dependency of the power P_O being fed back into the mains on the turn-on interval α of S_1 and on the DC link voltage level U_O.

In order to limit the considerations to the essential aspects the following assumptions are made for the further considerations

- sinusoidal, symmetrical mains phase voltages
- constant output voltage (ripple of DC link voltage neglected)
- source impedance of the mains purely inductive and linear (resistive component of the mains impedance neglected)
- idealized power semiconductors (neglection of on-state voltage drops, switching losses, etc.) and/or lossless system behavior, respectively.

Furthermore, in general the demagnetization phase $(\varphi_{N,2}, \varphi_{N,4})$ is excluded from the calculations, i.e., only the on-interval α of S_1 is considered in order to obtain simple and clear mathematical relations. Due to the very short duration of the demagnetization (cf. Fig.7(a) or (b)) as compared to α this causes only a minor error.

Remark: The assumption of a purely sinusoidal mains voltage is a critical point regarding the exactness of the calculations. In practice one has to expect a deviation of the mains voltage shape from a pure sine wave in any case around the peak of the line-to-line voltage depending on the load of the mains by uncontrolled rectifiers. Therefore, the shape of the mains current of the feed-back unit is in certain cases substantially influenced. Despite this fact, the dependency of the operating behavior on operating parameters of the system can only be calculated with a sensible effort if sinusoidal mains voltages are assumed. This assumption therefore corresponds to a compromise between clearness and exactness and shall be maintained in the following. If an increased exactness of modelling is required, an analytical calculation seems to be less sensible. In this case a digital simulation would be more advantageous.

3.1 Normalizations

In order to obtain results with clear and as much as possible general validity, the characteristic system quantities are given in normalized form. I.e., all quantities are related to characteristic system parameters. In place of the DC link voltage we set from now on

$$M = \frac{U_O}{\sqrt{3}\hat{U}_N} , \tag{1}$$

where $M = 1$ denotes a voltage U_O of the ammount of the amplitude of the line-to-line mains voltage. The mains current is related according to

$$i_{N,r} = \frac{1}{I_r} i_N \qquad I_r = \frac{U_{N,rms}}{\omega_N L_N} \tag{2}$$

to the short-circuit current I_r, the power P_O being fed back into the mains is related to the short-circuit power P_r being available at the terminals of the thyristor bridge

$$p_{O,r} = \frac{1}{P_r} p_O \qquad P_r = 3\frac{U_{N,rms}^2}{\omega_N L_N} . \tag{3}$$

3.2 Calculation Results

For the calculation the stationary operation of the system is considered, i.e., the average power being fed into the DC link by a load side current i_O for a constant value of U_O is expressed by

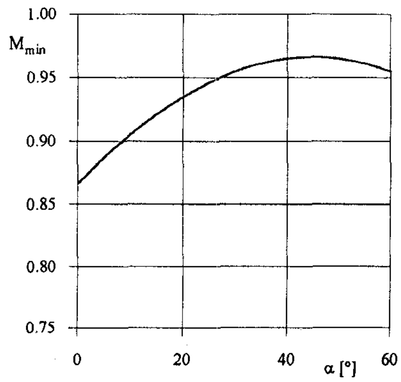

Fig.8: Limit of the validity of the analysis of the stationary operating behavior of the feed-back unit.

$$P_{O,avg} = U_O I_{avg} = U_O I_{S_1,avg} . \tag{4}$$

There the index 'avg' denotes the averaging of i_{S_1} performed over $\frac{\pi}{3}$.

The calculation of the average value I_{avg} of the current i being taken from the DC link by the feed-back unit (and/or of the average value $I_{S_1,avg} = I_{avg}$ of the transistor current i_{S_1}) shall be performed related to the angle interval $\varphi_N \in (-\frac{\pi}{6}, \frac{\pi}{6})$ shown in Fig.4. Within α the phase current $-i_{N,T}$ shown there and i_{S_1} have identical shapes.

For keeping the calculation of the power $P_{O,avg}$ as simple and clear as possible, we now consider that region $M > M_{min}$ of the normalized DC link voltage for which i has an always positive sign within α (the shape of $-i_{N,T}$ given in the limiting case is shown by a dashed line in Fig.4). Therefore, we can exclude from the considerations a partial recharging of C_O (occurring for $M < M_{min}$, but being of no importance for practical system applications) via the rectifier bridge lying antiparallel to the feed-back unit.

For the shape of the transistor current and the mains phase current within the considered angle region there follows with Fig.4 after a short calculation

$$-i_{N,T,r} = i_{N,R,r} = i_{S_1,r} = \frac{\sqrt{3}}{\sqrt{2}}[M(\varphi_N+\frac{\pi}{6})-\sin\varphi_N-\frac{1}{2}] \quad \varphi_N \in (\varphi_{N,1}, \varphi_{N,2}) . \tag{5}$$

The limit of validity $U_{O,min}$ or M_{min} for the equations derived in the following is given, therefore, according to $i_{N,T,r}|_{\varphi_{N2}} = 0$ by

$$M_{min} = \frac{1}{\alpha}[\frac{1}{2} + \sin(\alpha - \frac{\pi}{6})] \tag{6}$$

(cf. **Fig.8**). The power being fed back into the mains can now be calculated simply from Eq.(5) by averaging over $(-\frac{\pi}{6}, +\frac{\pi}{6})$. Considering Eqs.(4,3) yields

$$p_{O,avg,r} = \frac{3}{\pi} M[\sin(\alpha + \frac{\pi}{3}) + \frac{1}{2}\alpha(M\alpha - 1) - \frac{\sqrt{3}}{2}] . \tag{7}$$

A graphical representation of Eq.(7), i.e., of the dependency of $P_{O,avg}$ on the turn-on angle α and on the normalized DC link voltage M is shown in **Fig.9**. As becomes immediately clear from the considerations in section 2, $P_{O,avg}$ is increased by an extension of the on-interval α of S_1. For constant α an increase of M is

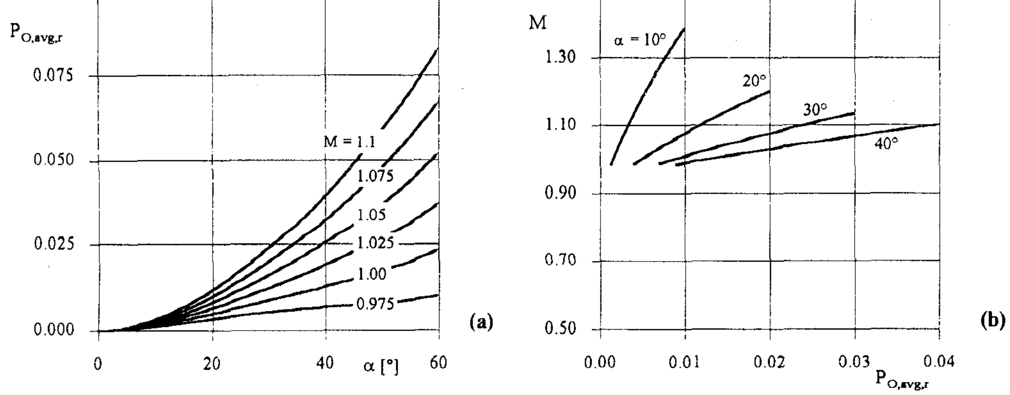

Fig.9: Dependency of the (normalized) power $P_{O,avg,r}$ being fed back into the mains on the transistor on-interval α (cf. **(a)**, parameter of the family of curves: M) and normalized DC link voltage M (cf. **(b)**, parameter of the family of curves: α).

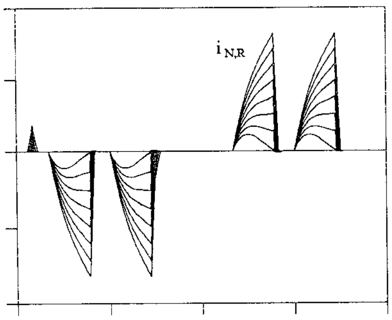

Fig.10: Dependency of the shape of the mains phase current $i_{N,R}$ of the feed-back unit on $P_{O,r} = 0.005 \ldots 0.04$ for $\alpha = 40°$; remaining operating parameters of the system according to section 2.2; scale: 200A/div.

required for an increase of the power delivery to the mains. For a practical system application especially the region $P_{O,\text{avg,r}} < 0.03$ is of interest according to the considerations regarding x_L (and/or regarding the short-circuit power P_r of the mains) in section 2.2. The on-angle will be typically in the vicinity of $\alpha \approx 40°$ for guaranteeing a sufficient long time with no forward voltage across the thyristors and an as small as possible blocking voltage stress on the power semiconductors (an as small as possible value of M).

Besides the determination of the power being fed back into the mains, Eq.(5) also yields a simple way to calculate the current stress on the power transistor which is of interest for a dimensioning of the system. For the average value of the transistor current i_{S_1} there follows

$$I_{S_1,\text{avg,r}} = I_{O,\text{avg,r}} = \frac{1}{\sqrt{6}M} P_{O,\text{avg,r}} \ . \tag{8}$$

The rms value of i_{S_1} results as

$$I^2_{S_1,\text{rms,r}} = \frac{9}{2\pi}(\frac{\sqrt{3}}{4}\cos^2\alpha + [M(\sqrt{3}\alpha + 1) - \frac{1}{4}\sin\alpha]\cos\alpha + (\alpha - \sqrt{3})M\sin\alpha - \sin(\alpha + \frac{\pi}{3}) + \alpha^2 M(\frac{M\alpha}{3} - \frac{1}{2}) + \frac{3}{4}\alpha - M + \frac{\sqrt{3}}{4}) \ . \tag{9}$$

For the calculation of the peak value $\hat{I}_{S_1}$ of i_{S_1} one has to destinguish for given on-time α between different cases concerning the value of the DC link voltage U_O (the value of M) and/or the power $P_{O,\text{avg}}$. As **Fig.10** shows, only for sufficient high $P_{O,\text{avg}}$ and/or sufficient high U_O the maximum of i_{S_1} results at the end $\varphi_{N,1}$ of the on-interval α. In this case we have

$$\hat{I}_{S,\text{r}} = \frac{\sqrt{3}}{\sqrt{2}}(M\alpha - \sin(\alpha - \frac{\pi}{6}) - \frac{1}{2}) \qquad \begin{matrix} \alpha \leq \frac{\pi}{6} & M > \cos(\alpha - \frac{\pi}{6}) \\ \alpha > \frac{\pi}{6} & M > 1 \ . \end{matrix} \tag{10}$$

If $\hat{I}_{S_1,\text{r}}$ lies within α, then Eq.(10) has to be replaced by

$$\hat{I}_{S_1,\text{r}} = \frac{\sqrt{3}}{\sqrt{2}}[M\arcsin(M) + \sqrt{1-M^2} - \frac{\pi}{3}M - \frac{1}{2}] \qquad \begin{matrix} \alpha \leq \frac{\pi}{6} & M_{\min} \leq M \leq \cos(\alpha - \frac{\pi}{6}) \\ \alpha > \frac{\pi}{6} & M_{\min} \leq M \leq 1 \end{matrix} \tag{11}$$

($M_{\min}$ is defined by Eq.(6)). The dependency calculated here of $\hat{I}_{S_1}$ and $I_{S_1,\text{rms}}$ on the power $P_{O,\text{r}}$ being fed into the mains and on the on-time α of S_1 is shown graphically in **Fig.11**. For a current stress on S_1 as low as possible one has to select a value α as high as possible for given $P_{O,\text{r}}$.

Based on Eqs.(8,9,10) and Eq.(11) one can now also easily determine the current stress on the thyristors. According to the on-intervals of the thyristors being limited to $\frac{1}{3}$ of the mains period there follows

$$\begin{aligned} I_{T,\text{avg,r}} &= \frac{1}{3} I_{S_1,\text{avg,r}} \\ I_{T,\text{rms,r}} &= \frac{1}{\sqrt{3}} I_{S_1,\text{rms,r}} \\ \hat{I}_{T,\text{r}} &= \hat{I}_{S_1,\text{r}} \ . \end{aligned} \tag{12}$$

An estimate of the current stress on the DC link capacitor C_O caused by the feed-back unit can be performed by considering Eqs.(8,9) via

$$I^2_{C_O,\text{rms}} = I^2_{S_1,\text{rms}} - I^2_{S_1,\text{avg}} \tag{13}$$

(the load current i_O is assumed to be constant there: $i_O = I_{O,\text{avg}} = I_{S_1,\text{avg}}$).

4 Control-Oriented Modelling

As one can see from the considerations of the previous section, one has to select a value of α as high as possible for a voltage and current stress on the valves as low as possible. For the simplest case then a system control could be realized in such a way that for energy feed-back operation the maximum allowable value of $\alpha_{\max}$ would be adjusted (under consideration of the recovery time of the thyristors); then, the control of the power flow would follow automatically by a (minor) increase of the DC link voltage.

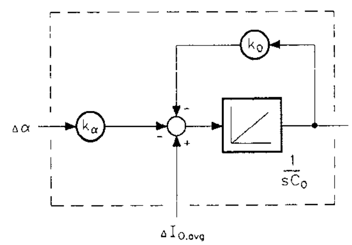

Fig.12: Control-oriented small-signal equivalent circuit of the energy feed-back unit under the assumption of an averaging of the input and output quantities over a $\frac{\pi}{3}$-wide interval of the mains period.

The automatic control characteristic of the energy feed-back unit can be described for given α by a differential source conductance

$$k_O = \frac{\mathrm{d}I_{\text{avg}}}{\mathrm{d}U_{O,\text{avg}}} = \frac{3}{4\pi}\frac{\alpha_0^2}{\omega_N L_N} \ . \tag{14}$$

If, in a similar manner, also the effect of a change of α on the average current consuption $I_{\text{avg,r}} = I_{O,\text{avg}}$ of the feed-back unit is characterized by a differential gain

$$k_\alpha = \frac{\mathrm{d}I_{\text{avg}}}{\mathrm{d}\alpha} = \frac{3}{2\pi}\frac{\hat{U}_N}{\omega_N L_N}[\cos(\alpha_0 + \frac{\pi}{3}) + \alpha_0 M_0 - \frac{1}{2}] \tag{15}$$

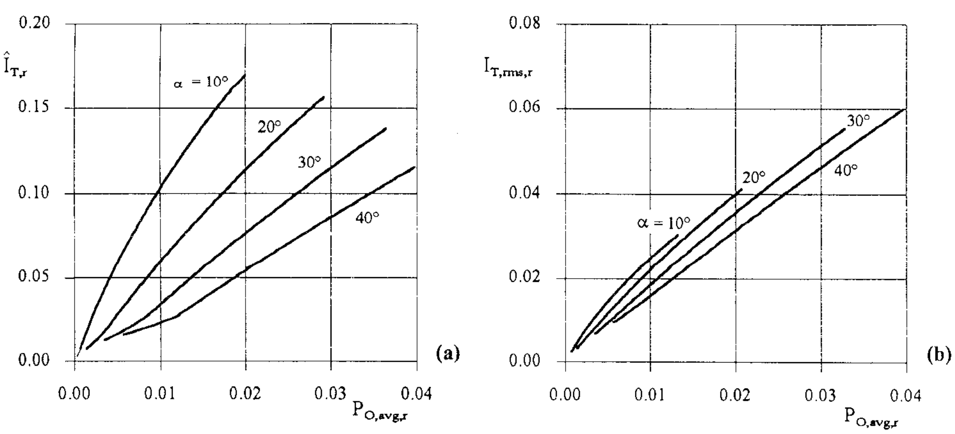

Fig.11: Rated peak value $\hat{I}_{S_1,\text{r}}$ (cf. (a)) and rated rms value $I_{S_1,\text{rms,r}}$ (cf. (b)) of the transistor current i_{S_1} in dependency on the (rated) power $P_{O,\text{avg,r}}$ being fed back into the mains; parameter of the family of curves: transistor on-interval α.

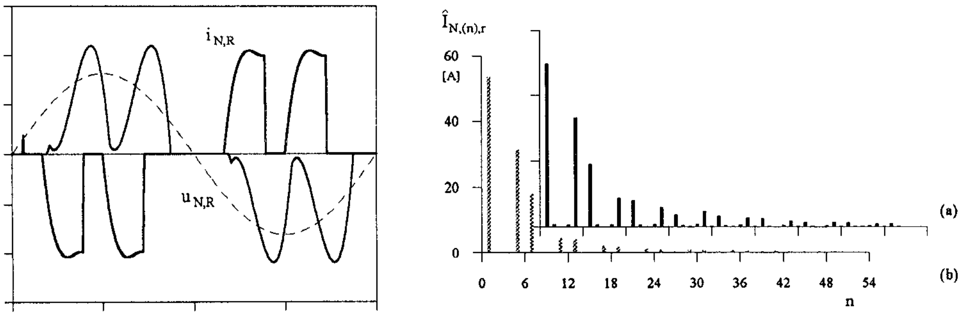

Fig.13: Comparison of the time dependencies and spectra of the phase currents of the energy feed-back unit (cf. **(a)**) and of a diode rectifier (cf. **(b)**) for equal power $P_{O,\text{avg},r}$. Operating parameters: $P_{O,r} = 0.0125$, $\alpha = 40°$, remaining operating parameters as in section 2.2; for buffering of the DC-link voltage a DC-link capacitor $C_O = 9\,\text{mF}$ is applied (the DC-link voltage is *not* impressed by an ideal volatge source); $P_{O,\text{avg},r}$ is fed into the DC-link by a source with constant power and/or taken from the DC-link for rectifier operation; scales: 40A/div, 200V/div.

(stationary operating point of the system defined by α_0 and M_0), there follows the control-oriented small-signal equivalent circuit of the system as shown in **Fig.12**. An estimate of the dynamic behavior of the feed-back unit could therefore be based in a first approximation on the transfer functions

$$G(s) = \frac{\Delta U_O(s)}{\Delta\alpha(s)} \approx -\frac{k_\alpha}{k_O}\frac{1}{1+s\frac{C_O}{k_O}} \qquad (16)$$

and

$$H(s) = \frac{\Delta U_O(s)}{\Delta I_{O,\text{avg}}(s)} \approx \frac{1}{k_O}\frac{1}{1+s\frac{C_O}{k_O}}\ . \qquad (17)$$

5 Effects on the Mains

For judging the mains behavior of the system the comparison to a diode rectification is of primary interest (of more interest than a comparison to an ideal system with purely sinusoidal mains phase currents). This is true especially because, e.g., for supplying the DC link of a drive system a simple three-phase diode bridge is provided in most cases; the thereby caused effects on the mains are accepted due to the economy and the robustness of this solution. Therefore, it is not very sensible to strive for a sinusoidal shape of the mains current (and/or ideal mains behavior) for feeding back braking energy into the mains.

The time behavior of the mains current of the feed-back unit and of a diode rectifier with equal power $P_{O,\text{avg},r}$ are shown in **Fig.13**(a). Both systems have a current flow interval being constrained to the vicinity of the maxima of the line-to-line mains voltage; both have approximately the same peak value of the mains current; therefore, there result (as one can see from Fig.13(b)) approximately the same amplitudes of the low-order harmonics (ordinal numbers n= $5, 7, 11, 13$) which are especially important for a judgement the effects on the mains. Harmonics of higher order are more pronounced for the energy feed-back unit than for the diode rectifier due to the steeper edges of the mains current.

The dependency on the fundamental phase displacement factor $\cos\varphi$ (as calculated by the digital simulation) and of the total harmonic distortion (THD) of the mains current of the feed-back unit on the power $P_{O,\text{avg},r}$ being fed into the mains are shown in **Fig.14**. Furthermore, in Fig.14 the characteristic of the rectifier unit with equal power is given again by a dashed curve.

With increasing power $P_{O,\text{avg},r}$ a phase shift between mains current fundamental and mains phase voltage results for the diode rectifier due to the then longer commutation intervals. This phase lag of the current results in a decrease of the $\cos\varphi$ with increasing power. However, for higher power a long conduction interval and, therefore, a better ratio between peak value and amplitude of the mains current fundamental is obtained due to the then continuous shape of the DC-side current. This can be seen by a reduction of the THD of the mains current for increased power. For the feed-back unit there remains also for high power by all means a discontinuous shape of the DC-side current and, thereby, also of the mains phase current. The increase of the power therefore can be obtained only by increasing the peak value of the mains current. This results in a higher current distortion as compared to the diode rectification. Based on a similar condition also the increase of the THD for a decrease of the on-time α of the power transistor becomes understandable. Because the on-intervals of the feed-back unit have a fixed time relationship to the line-to-line mains voltage and, therefore, also to the mains phase voltage, the displacement factor of the mains phase current (except for small power values $P_{O,\text{avg},r}$) is influenced only to a very minor extent and, therefore, it has higher values than a diode rectifier with equal power rating).

By summarizing the previous explanations one has to limit the braking power to values of $P_{O,\text{avg},r} < 0.02$ if the effects on the mains for energy feed-back operation shall not substantially exceed the effects on the mains for rectifier operation. Therefore, the system can be applied advantageously especially in cases where not the full rated power of the drive is required as braking power (according to section 2.2 the rated power for motoring (rectifier) operation is selected typically in the region $P_{O,\text{avg},r} = 0.02 \ldots 0.03 P_r$).

Finally, we want to point out briefly the possibility of minimizing the effects on the mains of several VSI drive systems operated in parallel based on a modification of the control of the feed-back unit. As becomes immediately clear from Fig.13, the feed-back current can be brought approximately into opposite phase to the input current of a diode bridge by delayed power transistor turn-on and/or delayed firing of the thyristors. (Then the on-time has to be shortened accordingly in order to guarantee a sufficient long time of negative voltage across the thyristors.) E.g., one could select the on-interval α symmetrically around the maxima of the line-to-line mains voltage participating in the current conduction in each case (e.g., $u_{N,TS}$ in Fig.4) . Thereby one would obtain a partial cancellation of the low-frequency harmonics of the systems operating in the motoring and in the braking mode. By this the current consumption of the feed-back unit with a high harmonic content would be used advantageously for a reduction of the effects on the mains of the overall system.

6 Conclusions

The system for supplying the DC-link of a drive system and/or for feed-back of

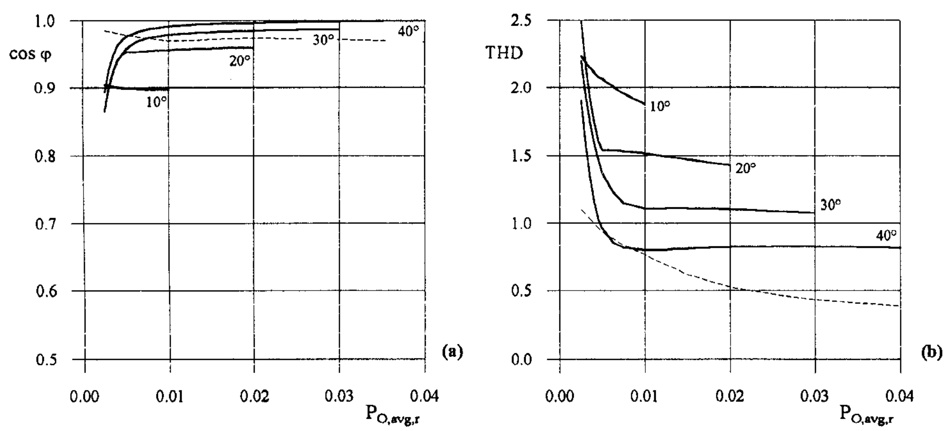

Fig.14: Displacement factor $\cos\varphi$ and total harmonic distortion (THD) of the mains current in dependency on the rated feed-back power $P_{O,\text{avg},r}$; parameter of the family of curves: on-interval α of the power transistor, remaining operating paramters according to section 2.2; the characterisitc of a three-phase diode bridge with output power $P_{O,\text{avg},r}$ (being operated at a mains with equal source inductance L) is shown by a dashed curve.

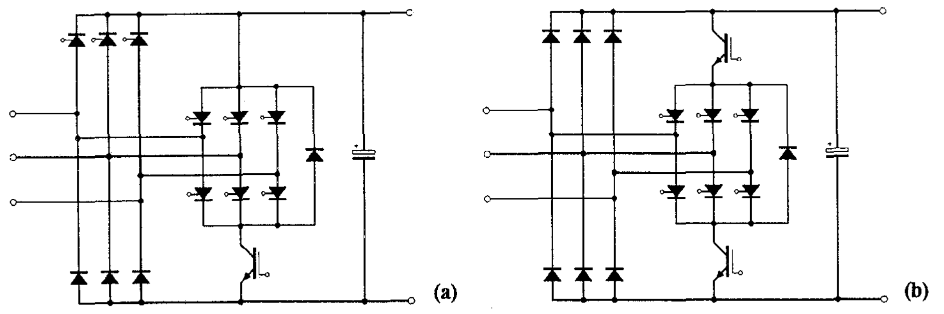

Fig.15: Improvement of the reliability of the supply/energy feed-back circuit according to Fig.2; **(a)**: the diodes D_1, D_3 and D_5 of the input diode bridge are replaced by thyristors; furthermore, an explicite free-wheeling diode D_F is provided; **(b)**: a further power transistor S_2 and an explicite free-wheeling diode D_F are provided, the input diode bridge remains unchanged; the system has a structure similar to the circuit proposed in [10].

the braking energy of the drive from the DC-link into the mains, as proposed in this paper, has th following advantages and disadvantages:

+ simple structure of the power and control circuit (power circuit contains only a single turn-off power semiconductor)
+ in principle no passive components required (especially, no inductors connected in series on the mains side or auto-transformers for voltage matching are required)
+ in principle no current measurement required for controlling the system
+ the system shows a natural voltage control property; if an increase of the DC link voltage by about 10% above the peak value of the mains line-to-line voltage is admissible a natural adaption of the mains current flow to the braking energy supplied into the DC link is achieved (a small increase of the DC link voltage – if lying above the peak value of the mains line-to-line voltage – rapidly increases the mains current flow); there, the conduction interval α of the transistor S_1 can be held constant at its maximum $\alpha \approx 40° \ldots 45°$ resulting in a very simple control circuit
+ no high-frequency effects on the mains (simplifies EMI-filtering as compared to PWM rectifier systems)
+ $\cos\varphi \approx -1$ for braking operation
+ for proper control of the feed-back unit the low-frequency harmonics of the mains current can be shifted approximately into opposite phase to the corresponding harmonics of the input current of diode recitifers supplying other VSI drive systems connected in parallel (partial cancellation of harmonics)
+ low losses and/or high efficiency (especially switching losses can be neglected as compared to conventional high-frequency PWM rectifier systems); the high efficiency of the power converter allows a compact design and a mounting on the heat sink of the machine-side inverter; therefore, no external mounting or additional fans are required (compare dimensions of a braking resistor for a continuous rated braking power of 10kW (safety class: IP00): 74x48x30cm3 (!), weight: 32kg(!)
+ the whole system could be integrated into a single power module; this would simplify the extension of existing unidirectional systems to regenerative braking (no additional passive components of high weight and size are required).

– relatively high peak current stresses on the power semiconductors
– for given saturation current of the inductors on the mains side (if there are any, e.g., for improving the THD of the diode rectifier) the maximum braking power typically is about 25% less than the maximum rectifier power due to the pronounced mains current peaks of the regenerating unit
– DC link voltage increases for energy feedback into the mains as compared to rectifier operation (about 5...10% above the peak value of the mains line-to-line voltage)
– if high reliability (especially in case of asymmetric mains phase voltages) is required a transducer for monitoring of the peak value of the transistor current and the recovery time of the thyristors has to be used
– the concept can be applied only up to a braking power of ≈ 100kW under consideration of the high current stress on the power devices and the high effects on the mains.

Besides the advantages and disadvantages mentioned here so far, also a major point is concerned with the question of reliability if one considers the practical application of the converter system. If, e.g., for the circuit acording to Fig.2 for $\varphi_N \in (-\frac{\pi}{6}, +\frac{\pi}{6})$ (cf. Fig.4) the demagnetization of the mains-side inductances is not completed in time (e.g., due to a dip in the mains voltage), a short-circuit via D_1 and T_5 results due to the sign reversal of the line-to-line voltage $u_{N,RT}$ in $\varphi_N = +\frac{\pi}{6}$. This current cannot be turned off by control means (the conditions shown in Fig.4(c) are given, however, with opposite sign of $u_{N,RT}$).

For industrial application one has to modify, therefore, the feed-back circuit proposed here with respect to a high reliability as shown in **Fig.15**(a) or (b). For Fig.15(a) the diodes of the positive half of the diode bridge are replaced by thyristors which are in the off-state during the feed-back of the braking energy into the mains; also, an explicite free-wheeling diode D_F is provided. Therefore, also during the turn-off phases of S_1 always only two phases participate in the current conduction; the mains current is deceased rapidly; no extension (which would decrease the interval of negative voltage across the thyristors) of the demagnetization process results due to a participation of a third phase in the current conduction (cf. Fig.4). If the demagnetization process is not finished in time , then the overcurrent resulting after turning the power transistor on again can be limited by turning off S_1 (and not by triggering a fuse, which would lead to a system interrupt) and the demagnetization can be continued.

A further advantage of the circuit according to Fig.15(a) is given by the fact that the controllability of the rectifier bridge can be applied during the start-up for a controlled charging of the DC-link capacitor. Therefore, no explicit precharging unit has to be provided. A more detailed discussion of the circuit shown in Fig.15(b) shall be omitted here for the sake of brevity (and because the realization effort would be higher than for (a)).

By the circuit modifications described so far only the opeating behavior of the system within the turn-off intrevals of S_1 is changed. Therefore, the mathematical description of the system behavior (for which the demagnetization phase has been neglected) given in the paper remains invariably valid.

Within the scope of further research a laboratory model of the circuit shown in Fig.15(a) is realized at present. A summary of the results of the experimental investigations of the system are planned to be presented at the IEEE Industry Applications Society Annual Meeting 1997.

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