ANALYSIS OF DIFFERENT CURRENT CONTROL CONCEPTS FOR FORCED COMMUTATED RECTIFIER (FCR)

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A B S T R A C T

The application of pulse width modulated (PWM) inverter drives, which seems to expand very much in the future, implies the demand for a new look at the analysis of the energy supply for the DC voltage link of such drives. Today the DC link usually gets its energy from the three phase mains via line commutated - controlled or uncontrolled - three phase SCR bridges. These are connected with a series of drawbacks whose avoidance seems to be possible by application of a PWM converter working in the rectifier mode (Forced Commutated Rectifier - FCR) under retention of economy and reliability.

In this paper the following points are treated: the general operating principle of an FCR system, its stationary behavior, the description by space vectors, and - starting from a cascade control structure - the application of a bang-bang current controller or, current control by a "Ramp Comparison Controller", respectively. Based on the application of the space vector concept finally a control concept is proposed which offers (compared to the bang-bang (two-step) control with a superimposed DC link voltage controller) advantages with respect to digital realization - besides advantages with respect to systems technology.

1. INTRODUCTION AND SYSTEM DESCRIPTION

One of the basic problems of power electronics is the connection of a generally polyphase AC system (mains) with a DC system (see [1], [2]). A polyphase (e.g. three-phase) system with symmetrical load and with sinusoidal currents supplies a constant total power. Without energy storage elements the "balance of power" between DC and AC side is given at all times and for all phase angles. (Energy storage devices would be necessary for coupling a single-phase AC system with a DC system due to the difference of the instantaneous power values in these systems - there is a power pulsation in the AC system with twice the fundamental frequency and no such pulsation in a three-phase system under the idealizations made above.)

In the case of unsymmetrical loads in polyphase systems only the power sum of the positive sequence system is steady. The negative sequence system caused by unsymmetry, however, results in power pulsations with twice the mains frequency.

These considerations show the basic possibility to develop a system which can serve as a coupling system between different mains without additional capacitances or inductances. The requirements for this so called "ideal transformer" between, e.g., three-phase AC systems (or between a three-phase AC system and a DC system) can be listed as follows (the points have to be weighted according to the specific application):

1. Possibility of bidirectional power exchange
2. Sinusoidal mains input current
3. Avoidance of electrical or magnetic power storage elements
4. No distortion of the mains voltage
5. Defineable reactive power consumption
6. Decoupling of mains and output voltage amplitudes

7. Possibility of buffering of load steps

8. Simple converter structure, minimization of the number of active components (controlled and uncontrolled semiconductors), reliability

9. Realizability with semiconductors according to the status of technology

10. Good dynamic performance

11. In the case of AC-AC conversion: free choice of the output frequency.

Requirements 2, 3 and 4 follow one from the other by the physical dependencies (of the ideal transformer). Because of the nonlinear semiconductor switches without storage capability and because of the time-discretization of the switching instants (note the maximum switching frequency, as always given in real systems), the requirement of an energy conversion with sinusoidal currents on the AC side can only be fulfilled approximately. Especially the periodic switching operations will cause harmonics. These can only be suppressed by additional magnetic or capacitive storage elements which absorb the high frequency power oscillations. Their required size is dependent on the order of the harmonics. In general a switching frequency as high as possible has to be strived for. This frequency mainly is limited by the switching properties of the semiconductors.

The structure of the "energy transformer" described here in a general case corresponds to a "matrix" of bidirectional switches. This makes possible the free choice of connections between arbitrary phases of the input and output system. According to today's status of semiconductor technology bidirectional controllable switches are only realizable by a combination of unidirectional switches. (For small power levels this combination can consist of, e.g., a series-opposing connection of FETs with integrated free wheeling diodes, for higher power levels an antiparallel application of two GTOs.) If, (due to economic reasons) for the basic element of the converter structure a power semiconductor with inherent turn-off capability and with antiparallel freewheeling diode (integrated into the same package) is chosen the "matrix converter" mentioned, in the case of AC-DC power conversion,

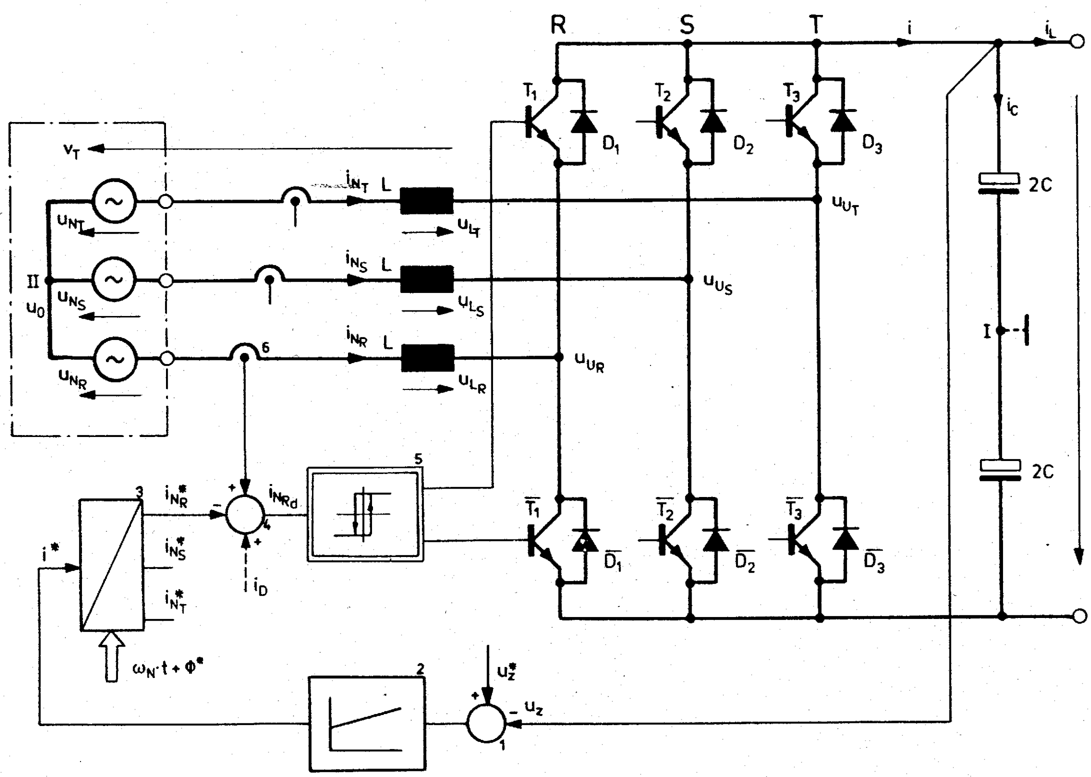

Fig.1: Structure of the power circuit and of the control of the converter (shown for bang-bang current control)

degenerates to the (PWM = pulse width modulated) inverter well known from the electrical drives area. The limitations connected with this concerning controllability do not essentially limit the applicability.

A simplification of this nature is not possible for the AC-AC converter as e.g., described in [3]. This, however, would be no real limitation if the required semiconductor devices would be available; this seems to be possible in principle with today`s technology of integrating complex semiconductor structures into power modules. However, the application of a system which is proven in its whole concept and working behavior, offers great economic advantages: such a system is the PWM inverter. Here the necessary modifications are limited to the control of the system and do not concern the power circuit.

The AC-AC energy conversion as required in today`s drive technology in general will be achieved advantageously with two PWM inverters connected via a DC voltage link. This DC link represents, so to say, the axis of electrical symmetry of the system. It also makes possible, as opposed to the AC-AC-converter, the simple decoupling of the two parts of the system by application of an energy storage device (DC link capacitor) between them. Only this can equalize the "balance of power" in each point of time for the aforementioned high frequency power pulsations of a "real transformer". Furthermore this offers buffering of power steps originated by the machine side inverter (converter), which is not possible for direct AC-AC conversion.

With respect to affecting the mains, converter circuits generally can be interpreted as "harmonic generators" whose currents are shorted by the impressed mains voltage fundamental. Without the inductive storage elements these harmonic voltages would lie completely across the mains impedance where they would cause a corresponding current distortion. A further critical point of a real system is given by the resonant frequencies of the supplying mains. They generally are determined by a frequency behavior which depends on the mains configuration and its load conditions. If expensive filter circuits have to be avoided, the current harmonics have to be shifted (by appropriate converter switching) into the region where the mains impedance decreases monotoneously with increasing frequency.

Within the scope of this paper in the following only the connecting system to the supplying mains shall be treated and - in the sense of energy conversion (AC-DC) - called FCR (Forced Commutated Rectifier). According to the descriptions given before its realization is possible by the structure shown in Fig.1. There the power circuit is formed by the energy storage elements on the mains side and on the DC side (inductances and capacitances) and by the power semiconductors which are connected according to the structure of a PWM inverter.

The simplification made relative to an ideal transformer (replacement of the bidirectionally controllable switches by unidirectionally controllable ones) leads to only partial decoupling of mains and output voltage amplitudes. Therefore the total controllability of the system is guaranteed only for DC link voltages which are (dependent on the load condition) in any case larger than the mains peak-to-peak voltage; this shall be an assumption for the following. For this case a converter leg can be viewed as switch between $+u_z/2$ and $-u_z/2$. (As reference point for the electric potential we choose the (virtual) center point of the DC link; in general there exists a voltage from the AC neutral to this center point which depends on the converter switching state.) By appropriate control (e.g., PWM) of the switches a converter (input) voltage can be produced which, if averaged over the switching period, is sinusoidal; this voltage forms a balance to the applied mains voltage across the inductances L. Therefore ultimately the vector difference of the phasors of the converter and mains voltages determines amplitude and phase of the input current. Because the converter does not contain storage elements the steady instantaneous real power (which is supplied by the mains) immediatly is transferred into the DC link. This means a steady converter current for constant DC link voltage.

The mains current resistive component is determined by the power requirement of the output circuit. The reactive power of the system can be selected freely via the mains current reactive component. This can be explained in an instructive manner by the fact that the instantaneous value of the sum of the reactive powers in the phases is zero at each point of time. The power pulsations with twice the mains frequency in each phase are transferred by the converter from phase to phase and therefore do not appear in the DC link. It would be possible to achieve a reactive power exchange with the supplying mains by an ideal converter (working with infinitely high switching frequency, where then the inductances on the AC side and the DC link capacitor could be omitted) without a physically present energy storage element. The possibility of omitting these storage elements can be understood only under consideration of the converter nonlinearity (switching system). Via the bidirectional power switches (bidirectional only with respect to the power flow) energy reversal is possible for given DC link voltage polarity. (Energy reversal means power flowing back from the DC link into the supplying mains.) Thereby the input current phase has to be adjusted via angle and magnitude of the converter voltage.

The description of the operational behavior of a real system so far shows no fundamental deviation as compared to the ideal system - besides parasitic effects which are basically linked to the nonlinear behavior of power electronic circuits. The real system therefore appears to be ideal for energy exchange between a three phase AC system and a DC voltage link (e.g., for a PWM inverter drive) - even under consideration of limitations caused by the actual realization. Besides that there exists a multitude of other applications; one of them, especially to be mentioned, is the possibility of building a "Static Reactive Power Compensator".

Today the power supply of a DC link of an inverter in the simplest way is achieved by an uncontrolled rectifier. The major drawback of this system is the unidirectional power flow (caused by the diodes), besides the voltage distortions in the mains caused by harmonics. The major advantage, on the other hand, is the extreme simplicity and the economical realizability. Due to the fixed DC link voltage polarity the possibility of feeding power back into the mains is only given by a line commutated converter in antiparallel to the uncontrolled rectifier. This causes additional harmonics in the mains and adds reactive contributions due to the control and commutation reactive power of the converter. A further drawback of this arrangement is given by the (according to the mains frequency) relatively low frequency of the harmonic current in the DC link capacitor. This would require large capacitances for sufficient smoothing of the DC link voltage. The expenditure concerning the required commutation inductances on the mains side basically can be said to be about equal to the inductive storage elements of the FCR system treated here. As a most important reason for the application of line commutated circuits as energy converters we can mention their economy, reliability and simplicity - besides their prooven technology. Their broad application is also given by the fact that up to the introduction of power semiconductors with turn-off capability no real economical alternative existed. With the further developement and price reduction of power electronic devices a substitution of line commutated circuits by "forced commutated" circuits (i.e., circuits with switches with turn-off capability) in the lower and medium power range seems to be thinkable. This substitution undoubtedly has technological advantages, it might be economically justifiable and could take place in the not too distant future. This possibility is - besides other viewpoints - especially given by the development of modern control and signal electronics which is connected closely to modern power electronics.

Finally the essential features of the FCR system described in this paper shall be summarized once more:

* low influence on the mains - sinusoidal mains current except harmonics which are basically linked to the switching frequency
* possibility of feeding power back into the mains
* arbitrarily selectable cos Φ (especially cosΦ=1!)
* controlled DC link voltage - good voltage utilization of the semiconductors
* DC link voltage higher than the mains voltage amplitude - no upper voltage limit linked to the basic working principle (advantage for drive systems)
* small DC link capacitor
* for drive systems: equal structure of mains side and machine side inverter
* possibility of integration of the whole power circuit into one module.

2. STATIONARY BEHAVIOR OF THE FRC SYSTEM - CONTROL CHARACTERISTICS

In general stationary working conditions are defined by steady amplitudes on the AC side and by constant values of DC output current and voltage. To get a simple and clear description of the system we assume purely sinusoidal variables on the mains side here. (The voltage between AC neutral and DC link center point in this case is zero.) Because the converter basically represents a discontinuous system, this point of view is justifiable only for very high (infinite) switching frequency or for exclusive consideration of the fundamentals. With these assumptions the behavior of the input portion can be described by phasor diagrams known from AC electrical engi-

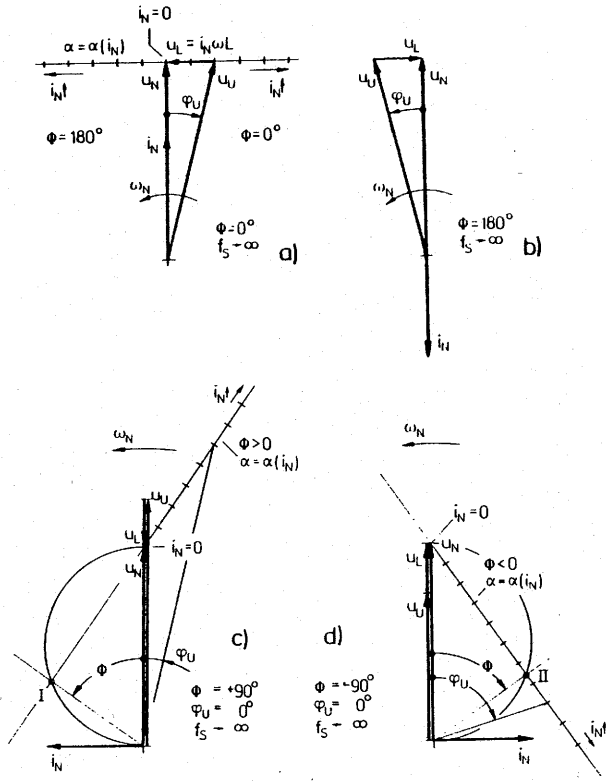

Fig.2: Phasor diagrams of possible load conditions as seen from the mains

a) resistive load
b) energy flow from the DC link back into the mains
c) inductive load
d) capacitive load

neering. As shown in Fig.2a, for purely resistive load of the mains the converter voltage u_U lags the mains voltage u_N by the phase angle φ_U. Also, u_U shows a larger amplitude than u_N. The (voltage) difference between the two phasors is required for impressing the current i_N into the inductance L. As shown by Figs.2b,c and d, giving u_U by amplitude and phase angle makes possible to force any load condition (given by the phase angle ϕ) onto the mains.

For the idealizations assumed here the converter voltage u_U is given by multiplication of the DC link voltage u_Z by the complex control factor (modulation index) $\underline{\alpha}$

$$\underline{u}_U = \underline{\alpha}.u_Z \qquad (1)$$

whose phase plots are given in Fig.2 for the different load conditions a)-d). The mathematical formulation of the relation-

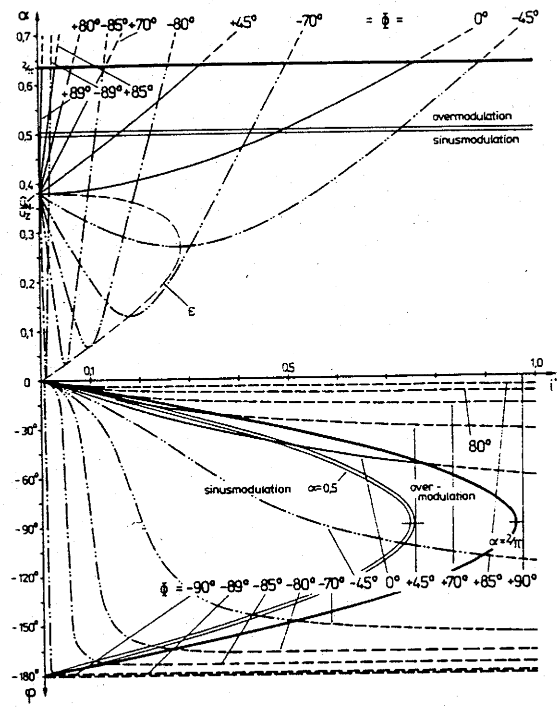

Fig.3: Control characteristics of the FCR system for u'_Z=2.0 (u_Z=620 V) $\hat{u}_N$=310 V

$\varphi_U = \varphi_U(\Phi, i, u_Z)$
$\alpha = \alpha(\Phi, i, u_Z)$

$$\alpha = \sqrt{\left(\frac{1}{u'_Z} + \frac{2}{3}.i'.\tan\Phi\right)^2 + \left(\frac{2}{3}.i'\right)^2} \qquad (2)$$

$$\tan\psi_U = \frac{-1}{\tan\Phi + \frac{3}{2.u'_Z.i'}} \qquad (3)$$

With the following normalizations:

$u'_U = u_U/\hat{u}_N$, $u'_Z = u_Z/\hat{u}_N$, $i' = i.\omega_N.L/\hat{u}_N$

ships can be derived from Figs.2 and leads to the control characteristics $\varphi_U=\varphi_U(i,\Phi,u_Z)$ and $\alpha=\alpha(i,\Phi,u_Z)$ (u_N and L fixed) which are given by Eqs.(2) and (3). They completely describe the system behavior in the stationary case. Thereby the equivalence of power (which determines the transformation of the load current i into the mains current i_N) on the converter input and output sides has to be considered. This balance of power is easily derived due to the fact that there are no power storage elements in the converter.

Equations (2) and (3) are shown in graphical form in Fig.3. The value α of the modulation index is limited by α_{max} = +1/2 if the region of overmodulation is excluded. Concerning the current, the region shown in Fig.3 for full controllability ultimately is given by the ratio u_z/u_N for given Φ and L. (The maximum is given by the highest allowable blocking voltage of the switching devices.) This becomes especially clear for the resistive-capacitive load case which in any case requires a converter output voltage larger than twice the mains voltage amplitude. The minimum of the "modulation depth" α appearing for the inductive load case can be explained by the phasor diagram (Fig.2d) in a simple way. Φ shall be given fixed. For linear increase of the mains current amplitude the reactive power "supplied" by the mains also increases linearly. On the other hand the reactive power in L depends on i_N^2. Therefore the converter has to show inductive behavior for small currents and capacitive behavior for high currents - for guaranteeing the power conditions given by the defined quantities in the mains. For the points II lying on a semi-circle with u_N as base, converter current and voltage are in phase. This means purely resistive converter behavior. The reactive power "supplied" by the mains in this case (for a defined mains current value, which is dependent on Φ) is only determined by the inductance L. The semicircle is mapped into a quadratically distorted ellipsoid ε in Fig.3. This represents the locus diagram of the minima of the relationship $\alpha = \alpha(i,\Phi,u_z)$. As follows from simple considerations of the geometry the turning points of $\varphi_U = \varphi_U(i,\Phi,u_z)$ coincide concerning i with the extrema mentioned.

The control characteristics for generatory system operation can be derived based on simple symmetry considertions from the locus diagrams given in Fig.3 (generally for resistive-inductive and resistive-capacitive load cases). Therefore no further detail shall be given here.

3. DESCRIPTION OF THE FCR SYSTEM VIA SPACE VECTORS

If the FCR system shown in Fig.1 is described by differential equations, the resulting set of equations is relatively involved. This is due to the three phases and their coupling via the voltage u_0. (The voltage between DC neutral (I in Fig.1) and AC neutral (II) is dependent on the converter switching state; eight different states are possible). For the system analysis it is advantageous therefore to find a form of description where the instantaneous system state is described by a set of variables which consider the variables of the three phases simultaneously. This problem (existing with the description of all polyphase systems) has been handled for AC machines by introduction of space vectors in the complex plane (e.g., [2], [11]). The space vectors in this case can be interpreted physically, e.g., as rotating flux vectors. Going further, space vectors can be used in general for description of three- and polyphase systems even if no physical interpretation is possible. Then these vectors serve as pure mathematical symbols which give a clearer description of a system and therefore simplify analytical treatments.

For description of the FCR system the space vectors have to be defined as follows:

mains current space vector:

$$\underline{i}_N = \frac{2}{3}\,(i_{N_R} + \underline{a}\cdot i_{N_S} + \underline{a}^2\cdot i_{N_T}) \,, \qquad (4)$$

where i_{N_R}, i_{N_S}, i_{N_T} are the instantaneous phase current values. For $\underline{a}$ and $\underline{a}^2$ (the complex factors of a 120° and 240° rotation) we have:

$$\underline{a} = e^{j120^\circ} = -\frac{1}{2} + j\frac{\sqrt{3}}{2}$$
$$\underline{a}^2 = e^{-j120^\circ} = -\frac{1}{2} - j\frac{\sqrt{3}}{2} \qquad (5)$$

The constant factor (here 2/3) basically is arbitrarily chosen because it only determines the vector length. Sometimes in literature other factors are used. With the definition used here the instantaneous phase values can be formed by projection of the space vector on the corresponding phase axis, e.g.:

$$i_{N_R} = \mathrm{Re}(\underline{i}_N)$$
$$i_{N_S} = \mathrm{Re}(\underline{i}_N\cdot\underline{a}^2) \qquad (6)$$
$$i_{N_T} = \mathrm{Re}(\underline{i}_N\cdot\underline{a})$$

It should be mentioned that according to the definition (Eq.(4)) so called "zero-values" do not contribute to the space vector. Such a "zero vector" is given, e.g., by a current value which is the same in the three phases in all instants ("zero-current"). The description by space vectors therefore is not uniquely reversible. According to this the values we gain by reverse transformation (Eq.(6)) are only the "zero-current"-free contributions. To these we generally

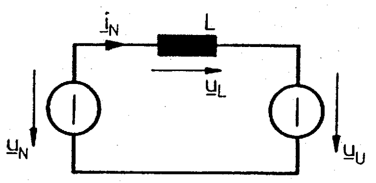

Fig.4: Equivalent circuit of the mains side; description via space vectors

would have to add the "zero-currents". Because in the FCR system described here no connection exists between the AC neutral and the (virtual) DC link center point, due to $i_{N_R}+i_{N_S}+i_{N_T}=0$ no "zero-current" can exist. Equations (6) therefore give the real phase values. On the other hand there exists a "zero-voltage" u_0 between the AC neutral and the DC link center point (which is the voltage reference point). Figure 4 shows the equivalent circuit of the mains side of the FCR system.

For the three phases we can write:

$$
\begin{aligned}
u_{N_R} + u_O &= L.(di_{N_R}/dt) + u_{U_R} \\
u_{N_S} + u_O &= L.(di_{N_S}/dt) + u_{U_S} \qquad (7) \\
u_{N_T} + u_O &= L.(di_{N_T}/dt) + u_{U_T}
\end{aligned}
$$

From the addition of these equations there follows under consideration of $u_{N_R}+u_{N_S}+u_{N_T}=0$ (zero voltage free mains) and $i_{N_R}+i_{N_S}+i_{N_T}=0$ (unconnected AC neutral):

$$u_O = \frac{1}{3}(u_{U_R} + u_{U_S} + u_{U_T}) \qquad (8)$$

The zero voltage here only exists due to the switching of the converter which is performed at discrete points of time. The converter basically can give a purely sinusoidal - and therefore zero voltage free - voltage system only for infinite converter switching frequency ($f_s \rightarrow \infty$). The converter phase voltage in general can be $\pm u_z/2$. From that it follows that the zero voltage u_0 can assume the values $\pm u_z/6$ and $\pm u_z/2$ if the converter is in one of the two freewheeling states.

If one defines the space vectors for mains and converter voltages in a similar way as has been done for the current, one receives the space vector differential equation for the FCR system from Eq.(4)

$$\underline{u}_N = L.(d\underline{i}_N/dt) + \underline{u}_U \qquad (9)$$

where the zero voltage u_0 does not appear any longer. Usually the space vectors are given in the orthogonal a,b-coordinate system ($\underline{i} = i_a + j\, i_b$). If the space vectors are written in a polar coordinate system

$$
\begin{aligned}
\underline{u}_N &= u_N \cdot e^{j\psi_N} \\
\underline{i}_N &= i_N \cdot e^{j\psi'_{\underline{i}N}} \\
\underline{u}_U &= u_U \cdot e^{j\psi'_U}
\end{aligned}
$$

we receive from Eq.(6):

$$
\begin{aligned}
u_N \cdot e^{j\psi_N} = {} & L.(di_N/dt).e^{j\psi'_{\underline{i}_N}} + \\
& j\omega_{\underline{i}_N}.L.i_N.e^{j\psi'_{\underline{i}_N}} + u_U.e^{j\psi'_{\underline{u}_U}} \qquad (10)
\end{aligned}
$$

If Eq.(7) is related to a new coordinate system (d,q-reference frame, Fig.5), which is rotated by the angle φ' and rotates with the angular speed ω, then Eq.(7) has to be multiplied by $e^{-j\varphi}$:

$$
\begin{aligned}
u_N \cdot e^{j\psi_N - \psi'} = {} & L.(di_N/dt).e^{j\psi'_{\underline{i}_N} - \psi'} + \\
& + j\omega_{\underline{i}_N}.L.i_N.e^{j\psi'_{\underline{i}_N} - \psi'} + \qquad (11) \\
& + u_U.e^{j\psi'_{\underline{u}_U} - \psi'}
\end{aligned}
$$

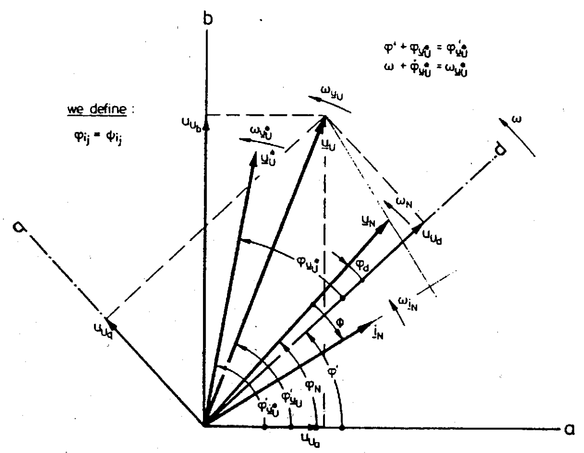

Fig.5: Definition of the angle relationships and of the angular speed relationships between system variables which are mapped to space vectors.

If we define now

$$\underline{u}_N = u_{N_d} + j.u_{N_q}$$

$$\underline{i}_N = i_{N_d} + j.i_{N_q}$$

$$\underline{u}_U = u_{U_d} + j.u_{U_q}$$

and if we split Eq.(11) into its real and imaginary part, we receive:

$$\begin{aligned} u_{N_d} &= L.(di_{N_d}/dt) - \omega_{\underline{i}_N}.L.i_{N_q} + u_{U_d} \\ u_{N_q} &= L.(di_{N_q}/dt) + \omega_{\underline{i}_N}.L.i_{N_d} + u_{U_q} \end{aligned} \qquad (12)$$

For calculation of the output current of the FCR system the balance of power has to be set up. Because the (ideal) converter does not contain any storage elements, its input and output power (at the converter terminals I and

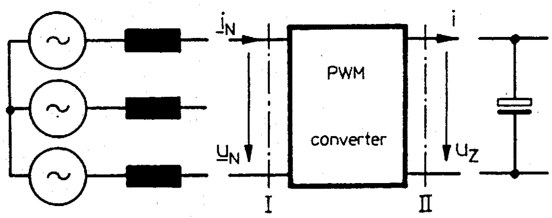

Fig.6: Overall system configuration

left: mains
center: PWM converter
right: DC voltage link, e.g., of an AC drive

II - see Fig.6) has to be identical at each instant of time. For the instantaneous power we have (independent of the chosen reference frame, see [2], [11]):

$$p = \frac{3}{2}\,\mathrm{Re}(\underline{u}_U^+.\underline{i}_N) \qquad (13)$$

$\underline{u}_U^+$ ... conj.complex value of $\underline{u}_U$

If this is set equal to the power $i.u_z$ entering the DC link and if the space vectors are described by the corresponding d,q-components, one receives for the converter output current:

$$i = \frac{3}{2}.\frac{1}{u_Z}.(u_{U_d}.i_{N_d} + u_{U_q}.i_{N_q}) \qquad (14)$$

For the differential equation of the DC link we then get:

$$i - i_L = C.(du_Z/dt) \qquad (15)$$

Considering the relationship for the angular velocity of the mains current space vector

$$\omega_{\underline{i}_N} = \omega + \frac{d}{dt}\,(\arctan i_{N_q}/i_{N_d}) \qquad (16)$$

we can determine from Eqs.(11,14,16) the block diagram of the FCR power circuit in the d,q-coordinate system (see Fig.7). If the input variables $\underline{u}_N$ and $\underline{u}_U$ are given in the R,S,T coordinate system, they are to be converted into the d,q system according to the transformation equations

$$\begin{aligned} u_{N,U_a} &= u_{N,U_R} = \frac{1}{3}\,(2\,u_{N,U_{RS}} + u_{N,U_{ST}}) \\ u_{N,U_b} &= \frac{1}{\sqrt{3}}\,(u_{N,U_S} - u_{N,U_T}) = \frac{1}{\sqrt{3}}\,u_{N,U_{ST}} \end{aligned} \qquad (17a)$$

$$\begin{aligned} u_{N,U_d} &= u_{N,U_a}\,.\cos\psi + u_{N,U_b}\,.\sin\psi \\ u_{N,U_q} &= -\,u_{N,U_a}\,.\sin\psi + u_{N,U_b}\,.\cos\psi\,. \end{aligned} \qquad (17b)$$

For practical realization application of line to line voltages, as given in Eq.(17), is preferable. These voltages are easier to measure and they are free of "zero-components".

The block diagram shown in Fig.7a is generally valid, i.e. the transformation angle φ can be chosen arbitrarily. It is advantageous to choose it equal to the mains voltage phase angle, whereby automatically $\omega = \omega_N$ results. This means that the d-axis of the d,q-reference frame is aligned along the mains voltage vector. By this orientation it follows that $u_{N_q}=0$; this hardly simplifies the block diagram structure, however. The great advantage of this transformation is, on the other hand, that there are only steady variables in the d,q system for stationary conditions. This is important for the control principle and its realization by a microprocessor system (sampled-data system). (It should be mentioned here that a PI-controller can only handle constant quantities without control errors because it has no "infinite gain" for alternating quantities.)

Originally the rotating coordinate systems have been used for analysis and control of AC machines ("field oriented control"). There orientation has been made in most cases in the direction of the rotor flux. By the orientation in the direction of the mains voltage as done here, the d- and q-axes physically have the meaning of true and reactive power axes. The current i_{N_d} corresponds with

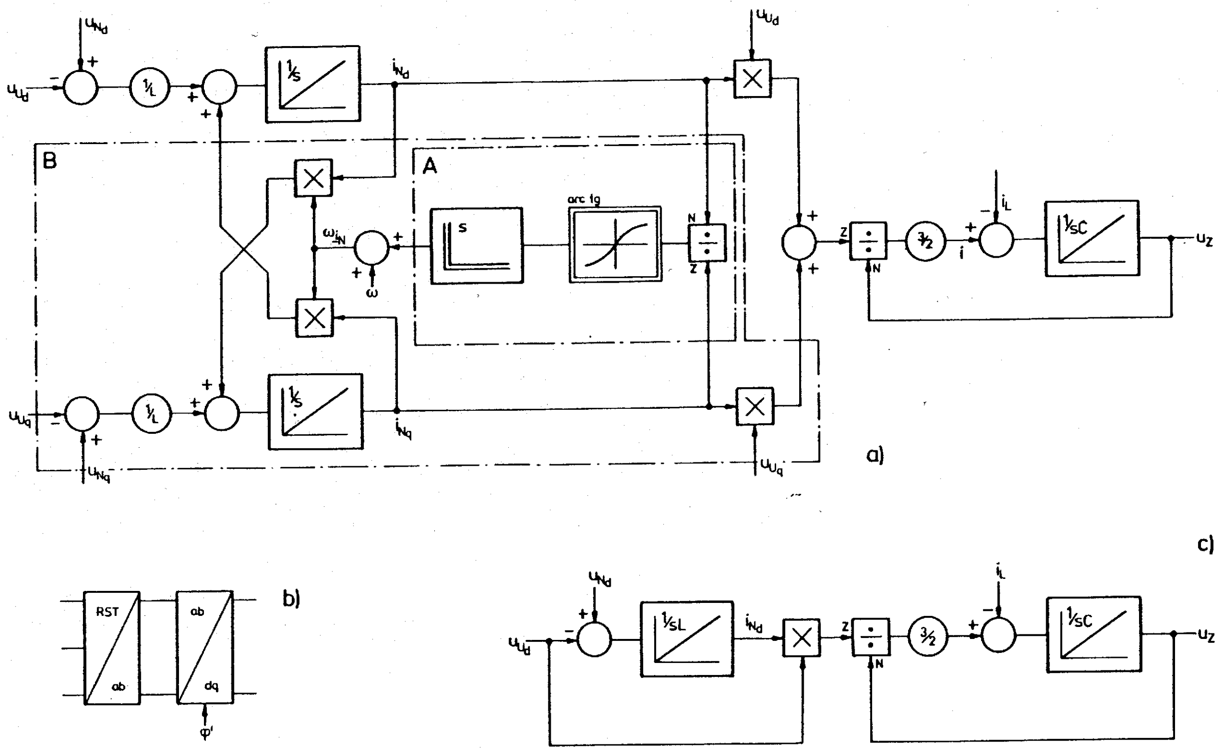

Fig.7: a) Block diagram of the control loop (power circuit) for representation of the describing space vectors in the d,q-coordinate system
b) Coordinate transformer
c) Simplified block diagram for the stationary case and $i_{Nq} = 0$

the resistive current component, i_{Nq} with the reactive current component. Therefore, ϕ = arctan i_{Nd}/i_{Nq} corresponds to the power factor (cos ϕ). In the stationary case of sinusoidal variables (for convertes this would mean $f_S \rightarrow \infty$ or averaging over the switching periods, i.e., neglection of harmonics) also here the duality between the space vectors and the phasors of complex calculus for AC systems is valid.

Simplifications of the block diagram (Fig.7) are possible under the following conditions. In the stationary case ($\omega_i = \omega_N = \omega$) block A can be omitted completely. Coupling of i_{Nd} and i_{Nq} remains, however. If the FCR shall be operated as purely resistive load (as seen by the mains) - which corresponds to the DC link load as transformed to the mains - block B can be omitted also. (This requires a control system that is able to maintain the operating condition $i_{Nq}=0$.) There remains a largely simplified control structure as shown in Fig.7c where there is no coupling between d- and q-quantities because the conditions in the DC link are only determined by the components of the d-axis.

It should be noted that the complete block diagram (Fig.7a) remains valid basically also for misalignment; this means that the relationships are represented correctly if the d-axis does not lie exactly in the direction of the mains voltage vector u_{Nd}. In this case, however, the possibility of linking the d- and q-quantities to resistive and reactive components is not given any more. This is apparent by the fact that also the q-quantities influence the DC link conditions (see also Eq.(14)). Misalignment can appear if the controller cannnot determine the mains voltage angle exactly. This can be caused by disturbances superimposed on variables necessary for determination of that angle.

4. CONTROL OF THE INPUT CURRENTS BY BANG-BANG CONTROL

The main task of the control of the system is the guidance of the DC link voltage under the condition of a desired mains phase angle. Prior to determination of the control structure the system behavior (which has been regarded so far based on various idealizations) has to be examined under consideration of the restrictions given by a realization in practice. For this purpose the behavior of the system which works in connection with a bang-bang controller shall be simulated by analog means. At first the DC link voltage shall be fixed. The investigation of a bang-bang controller offers itself because it represents a very simple solution for current control with respect to equipment realization. It can be designed without close knowledge of the controlled system. Furthermore the fast and phase-correct bang-bang control in connection with a DC link voltage (master) controller represents a possible control concept. In general the simulation provides guidelines for dimensioning of the system components. (Due to the high dynamics of the bang-bang controller this makes possible especially the determination of the minimum DC link capacitance.)

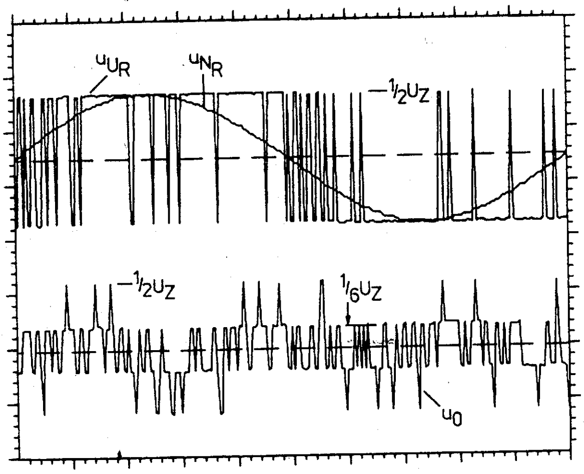

Fig. 8: Converter voltage u_{U_R}, mains voltage u_{N_R} and voltage u_0; (u: 250 V/D, t: 2 ms/D); parameters: u_z = 620 V, L = 10 mH, C = 1.1 mF, $\hat{i}^*_{N_R}$= 25 A, i_{TB} = 1.25 A

As shown in Fig.1 a hysteresis type bang-bang controller is provided for each of the three phases. Thereby the three controllers influence the system independently. If the current leaves the deadband, the corresponding converter phase is switched to such a potential which tends to decrease the control error. The switching process in the concerned converter phase exercises a substantial influence on the other two phases. This coupling of the control loops via the controlled system and as ultimately given by u_0 (see Fig.8) only shows effect via the actual system variables. However, it is not really considered for giving the controller outputs for the simple ("not intelligent") bang-bang controller. As shown by Fig.3.1, the time sequence of the control signals of the converter shows a very much irregular behavior. The switching frequency adjusts itself freely; only its mean value is given by the width of the hysteresis deadband. The two switching states for providing zero voltage (freewheeling state of the converter, u_0 = 1/2 u_z) appear more or less at random and not in every case at an appropriate instant (e.g., at the mains voltage maximum - see Fig.8).

In general working of the bang-bang controller requires a minimum value u_{zmin} of the DC link voltage. This minimum is given by the case where the control can only provide a three-phase voltage system whose fundamentals "balance" the mains voltages when the control makes the transition from overmodulation to full voltage blocks. At any rate, in the region of high overmodulation the current harmonics are shifted to lower frequencies and rise in amplitude. This means that the demand for low influence on the mains is not fulfilled any more. For a DC link voltage of 620 V (which is well manageable under consideration of today's semiconductor technology status) and for u_N=220 V_{eff} there only exists a minor (current dependent) deviation from sinusoidal modulation (Figs.8-12). The deviation of $\underline{i}_N$ from its reference value is only determined by the hysteresis (see also Fig.12).

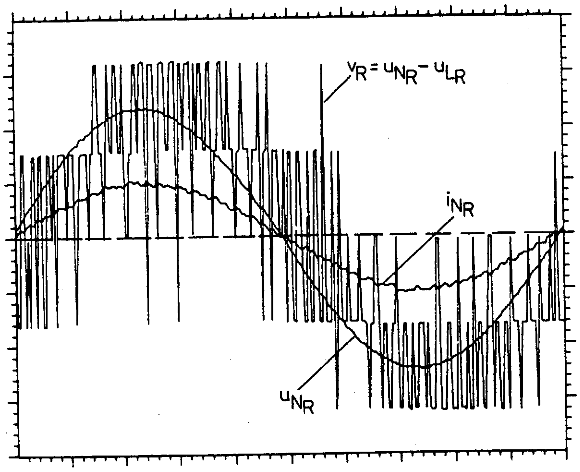

Fig. 9: Counter voltage v_R, mains voltage u_{N_R} and mains current i_{N_R}; (u: 125 V/D, i:25 A/D, t: 2 ms/D); parameters as in Fig.8

As also derived in chapter 2 the counter voltage (v_R) fundamental for resistive mains load lags the mains voltage by a phase angle which is dependent on L and i_N.

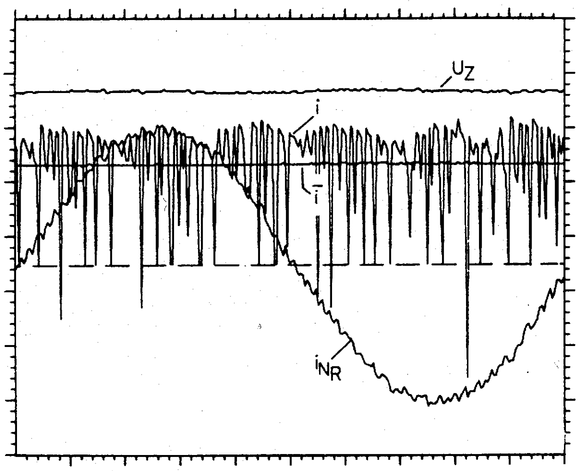

Fig.10: Converter output current i (mean value $\bar{i}$), mains current i_{N_R} and DC link voltage u_z; (u: 200 V/D, i:10 A/D, t: 2 ms/D); parameters as in Fig.8

The output current is formed based on the phase currents via the converter operating as "commutator". For given mains current u_z adjusts itself according to the load conditions such that the balance of power is fulfilled. For constant u_z and changing load an appropriate guidance of i_N is necessary.

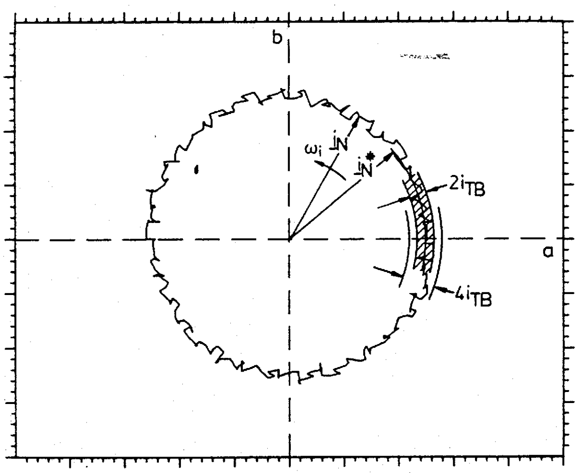

Fig.11: Mains current space vector $\underline{i}_N$ (reference value $\underline{i}_N^*$) shown for one fundamental period; i_{TB}: hysteresis of the bang-bang controller; (i: 10 A/D); parameters as in Fig.8

The current reference space vector $\underline{i}_N^*$ rotates with constant angular speed; only its mean value - taken over several periods of the switching frequency - is reproduced by the space vector of the actual current value.

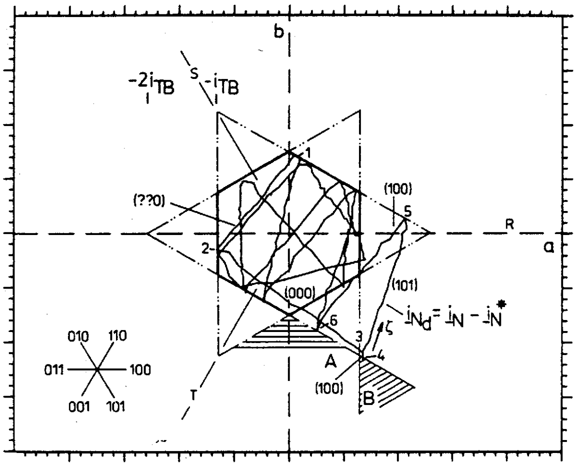

Fig.12: Space vector of the mains current control error $\underline{i}_{N_d}$; vector motion according to ζ ; shown for 1/10 of the fundamental period; (i: 1A/D); parameters as in Fig.8

Similar to the actual current values also the current control errors can be represented by space vectors. The deadbands of the phase-controllers are mapped on a hexagon (Fig.12). This is bounded by the switching lines of the phases which lie symmetrical to zero at $\pm i_{TB}$ and which are perpendicular to the phase directions. If the control error in each phase is not larger than the deadband, the space vector moves within the hexagon. However, under consideration of the mutual influence of the three phase current controllers and the counter voltage effect a further space for motion of $\underline{i}_{N_d}$ is given. The hexagon can be left at the moment of reaching the switching line of one phase despite the switching operation taking place in this same phase due to the switching state (which is dependent on system history) of the other two converter phases. Only on entering of one of the regions A further switching takes place in one more phase such that $\underline{i}_{N_d}$ is driven back to the center (an effective deadband width of $4.i_{TB}$ can be defined therefore). In the regions B the switching state of the whole system is determined. The guidance of the current control error within the two interlaced triangles is possible only for sufficiently high DC link voltage. If the counter-voltage is not sufficiently dominated by u_z, also these triangles are left (see ζ in Fig.12).
The time motion of the space vector takes place along ζ according to the sequence 1-6. The switching state of the converter can be determined based on the fact that the corresponding phase changes its

polarity when a switching line is reached. E.g., in 1 (Fig.12) i_{N_T} is too small - switching status 0 (lower switch closed), the other two switch positions are dependent on system history; because R is switched in 2 and S is not switched during motion from 2 to 3 when its lower deadband limit is crossed, the switching status valid between 2 and 3 is therefore determined: 000 (freewheeling). (On the bottom at the left hand side of Fig.12 the orientations of the converter voltages for the different switching states are shown.) From the transition between 2 and 3 via $\underline{u}_N = - L.(d\underline{i}_{N_d}/dt)$ the direction of the mains voltage can be determined. The switching status 101 between 4 and 5 determined by the controller as such would cause a motion toward the center. Due to the mains voltage (counter voltage for the current controller) the space vector motion between 4 and 5 is given according to the geometric difference of $\underline{u}_N$ and $\underline{u}_{U_{101}}$: the difference gives the direction of $\underline{u}_L$ showing that the triangle area is left.

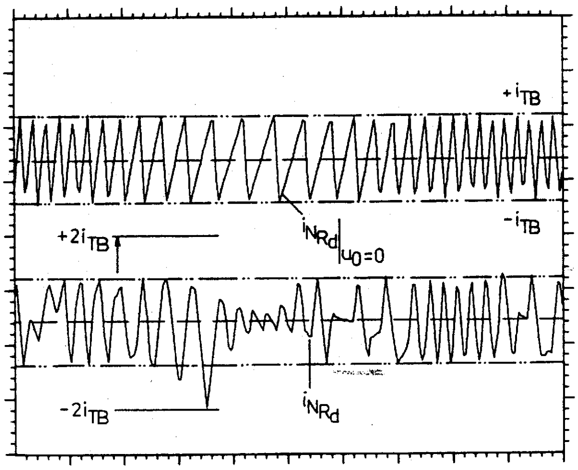

Fig.13: Current control error $i_{N_{R_d}}$ for phase R; hysteresis of the bang-bang current controller: i_{TB}; above: AC-neutral connected to the "virtual" center point of the DC link (u_0=0; below: AC neutral not connected to the center of the DC link; (i: 2.5 A/D, t: 1ms/D); parameters: u_z=820 V, L=10 mH, $\hat{i}_N^*$=25 A, i_{TB}=2 A

If u_0 = 0 is defined, meaning a direct connection between AC neutral and DC link center point, the three phases are decoupled.

The current controller of each phase in this case is in the position of guiding its current inside the deadband. Within the system then harmonics of the order 3.k (k=1,2,...) appear, however. The switching frequency here only is "modulated" by the counter voltage.

On the bottom of Fig.13 one can recognize the region which is characteristic for simple ("non-intelligent") three phase bang-bang controllers, where one phase is guided by the controllers of the two other phases. The phase itself (the phase with the highest counter-voltage) thereby is clamped to one side of the DC link (see [6], where, however, no clear statement is given for the effective deadband width).

In general for deadband controls which operate on alternating variables there remains a basic free range for the actual values of phase, frequency and amplitude relative to the reference value which is in the center of the deadband. The actual value will follow the reference value, dependent on the switching frequency, only with its time mean value. The averaging process for arriving at this mean value thereby can extend possibly over several periods of the fundamental. This basically implies the danger of causing low frequency harmonics (subharmonics). These may reach considerable values due to the reactances ω.L decreasing with decreasing frequencies. Furthermore, due to the random switching instances and processes, for bang-bang controllers there does not exist a possibility of reducing the influences on the mains by (resonant) filters. Besides the drawbacks mentioned there is the advantage of the excellent dynamic as documented by Figs.14, 15. For a reversal of the energy flow of the system the input current reference value has to be reversed

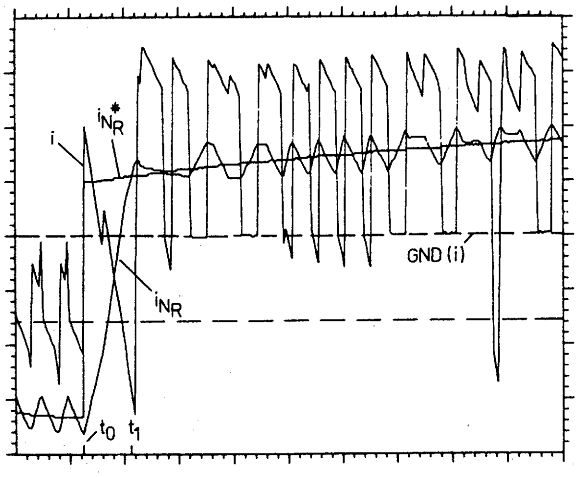

Fig.14: Reference value step of the mains currents amplitudes; at $t=t_0$ reversal of power flow direction (from DC link into mains to mains into DC link); mains current reference value $i_{N_R}^*$ and actual value i_{N_R} for phase R, converter output current i; (i: 10 A/D, t: 0.2 ms/D); parameters: u_z=820 V, L=2.2 mH, i_{TB}=3.4 A

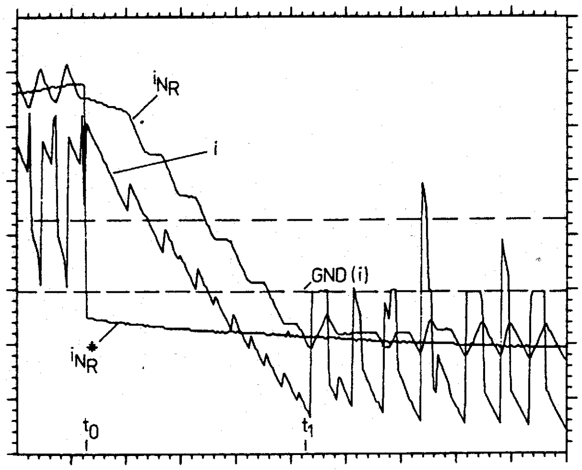

Fig.15: As Fig.14, but reverse order (from mains into DC link to DC link into mains)

in polarity; this is connected with a direction reversal of the converter output current. The current controller dynamic is determined by the DC link voltage, mains voltage and inductance L. According to the required direction of the energy flow reversal the mains voltage conditions existing at the time of controller action can accelerate or decelerate the current rate of change. The current $i_{N_R}^*$ required in Fig.14 from t_0 on is adjusted within t_1-t_0 by switching of the converter leg in position 0. The positive "instantaneous mains voltage value" existing at this point is added to the converter voltage u_{U_R} and accelerates the change of directions of i_{N_R} whose time behavior is further determined by L. According to the switching state of the converter phase the charging current will - for a short time - already assume the required polarity because i_{N_R} is partly fed into the negative DC link bus. If the mains voltage acts decelerating related to the current transfer duration, possibly a substantial increase of control time results (see Fig.15) - dependent on DC link voltage to mains voltage ratio.

The substitute time constant of the current control loop therefore is dependent on the counter voltage. The frequently mentioned high dynamic of the bang-bang controller therefore only can be reached by appropriate choice of the system parameters (DC link voltage, inductances).

The size of energy storage elements (inductances) on the input side essentially is determined by

* the maximum mean converter switching frequency
* the DC link voltage value and

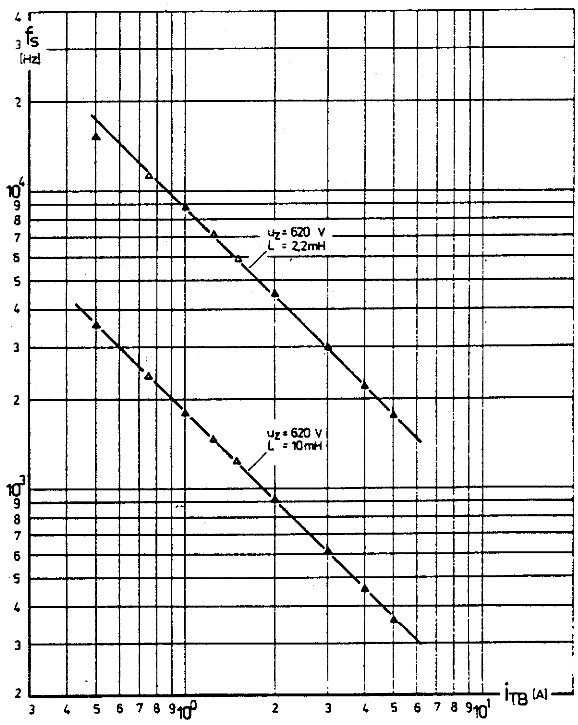

Fig.16: Switching frequency f_s as a function of i_{TB} (= half deadband width)

* the tolerable mains influence.

Thereby an economically optimal overall system has to be aimed at. The required minimum value for the inductance L regarding the effect on the mains according to the existing experiences lies at about 2 mH; this takes into consideration the impedance of energy distribution systems for the kilohertz region (as possible converter switching frequency). According to the (linear) current rise caused by L there exists proportionality between the current change and the time required for this. Consequently the mean switching frequency of the phase is inversely proportional to the deadband width (see [6]). From Fig.16 one can take the mains current ripple which follows for given inductance and converter switching frequency. (The peak-to-peak value of the current ripple is $4 \cdot i_{TB}$).

5. CURRENT CONTROL BY RAMP COMPARISON CONTROLLER

Due to the eight possible converter switching positions basically the possibility exists to generate six nonzero voltage space vectors at the converter output. The two remaining switching con-

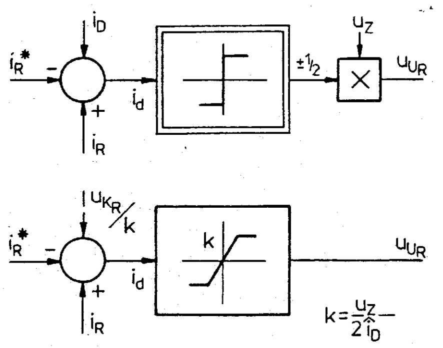

Fig.17: "Linearization" of the bang-bang controller by a triangle signal su_perimposed onto the control error - transfer into a ramp- comparison controller (pulse width modulator).

ditions (000, 111) represent the freewheeling state (zero voltage vector). For impressing of purely sinusoidal currents (in the stationary case) the converter voltage space vector should rotate with constant angular speed. However, this reference value vector only can be approximated in the average by a sequence of discrete vector positions. For the bang-bang controller for achieving this voltage vector "arbitrary" vector positions are used. The obviously optimal possibility for approximation of a rotating vector, however, is given by exclusive application of the neighbouring vector positions and the zero vector. This can be reached only by coupling of the phase current controllers because the converter voltage vector is determined by the switching state of the overall system. As can be seen from Fig.18 this can be reached as follows: the single control errors are compared with a triangle signal of sufficient amplitude (ramp comparison controller); from this the switching operations of the converter phases are derived. By addition of this "modulation signal" i_D to the control error and comparison with zero the bang-bang current controller shown in Fig.1 is transformed into a "quasi-linear" pulse width modulator (Fig.17). Therefore the current controller can be replaced in the sense of control theory by a proportional transfer element with gain $k = u_z/(2.\hat{i}_D)$ - if averaging over time is performed. If the modulation signal is chosen synchronous to the fundamental and - due to symmetry considerations - has 3k (k=1,2,3,...) times the fundamental frequency, subharmonics (caused by mixing of triangle frequencies and converter fundamental) can be avoided. Therefore only multiples of the fundamental will exist because the frequency of the triangular voltage is an integer multiple of the converter fundamental frequency. The spectra (Figs.20 and 21) are therefore defined, contrary to the bang-bang controller. They allow definite conclusions concerning the effects on the mains; these effects then are especially determined regarding their frequency spectrum.

By the switching frequency (which is fixed in this control concept) furthermore the requirements for the converter power circuit concerning the maximum pulse frequency are definitely determined. In general then also the acoustic noise caused by the system operation will be found less disturbing - acoustic damping will be made easier.

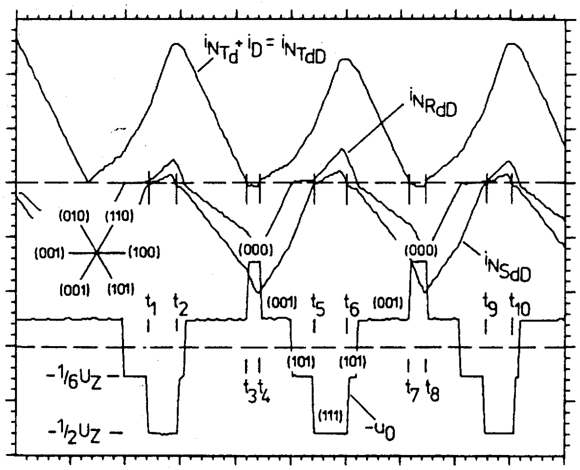

Fig.18: Sum of mains current control error and synchronization current (i_D) for phases R,S,T and voltage u_o ; (i: 5 A/D, u: 250 V/D, t: 0.2 ms/D); parameters: u_z=820 V, L=10 mH, i_{TB}=50 mA, $f_S=33.f_N$ (synchronized)

The two freewheeling conditions of the FCR system are mapped into a corresponding voltage u_o in a definite way (u_o(000):+1/2 u_z; (111):-1/2 u_z). If the control errors $\underline{i}_{Nd}$ existing in the different phases would be all zero, all converter phases would be switched simultaneously between the status 000 and 111; therefore for u_o there would result a square-wave voltage with the frequency of the triangle. By the control errors being nonzero the freewheeling time periods are shortened (freewheeling intervals: t_1-t_2, t_3-t_4, t_5-t_6, ...). If there appears a positive control error according to the definition $\underline{i}_{N_{Rd}}=\underline{i}_{N_R}-\underline{i}^*_{N_R}$ the time duration for output of a positive converter voltage in the corresponding phase is extended. By this the actual value is driven in the direction of the reference value. The switching states assumed by the FCR system within a period of the triangle signal are shown in Fig.18. As the consideration of the corresponding converter voltage vector positions shows, the voltage vector is approximated by neighbouring vector positions.

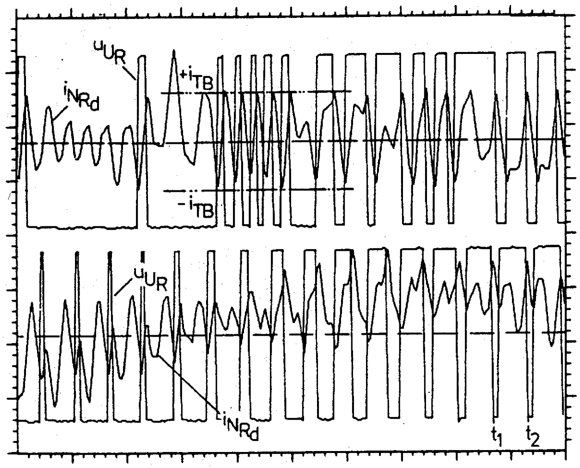

Fig.19: Converter voltage u_{U_R} and current control error $i_{N_{Rd}}$; below: ramp comparison controller with $f_D=33.f_N$ (i_{TB}=40 mA), without compensation of counter voltage v_R; above: bang-bang controller with hysteresis i_{TB}=2.3 A such that the resulting average switching frequency becomes f_D; (u: 250 V/D, i: 2.5 A/D, t: 1 ms/D); parameters: u_z=820 V, L=10 mH

As Figs.19 and 22 (bottom) show, the transformation of the current controller into a proportional (P-) element leads to the occurrence of a current control error with a fundamental frequency component. Amplitude and phase of this component can be determined by a phasor diagram. This occurrence is caused by the fact that only then the P-current-controller (in connection with the converter quasi acting as amplifier) is in the position to generate the required output voltage. The occurring (necessary) control error results in a phase error ϑ and amplitude error Δ of the actual value. The time difference t_2-t_1 (Fig.22) corresponds to the converter voltage phase lag which is necessary for resistive behavior as seen by the mains.

In the case of the bang-bang controller due to the "statistical nature" of its switching operation only the specification of a spectrum is possible which is gained from averaging over a number of Fourier transforms valid for arbitrary time intervals. A discrete spectrum for the bang-bang controller exists only in these cases where the mean switching frequency is an integer multiple of the fundamental frequency. This can be achieved by appropriate combination of the system parameters (e.g., the deadband width). Thereby a recurring "pulse pattern" is put out. The frequencies appearing in the vicinity of the fundamental frequency are determined by the reference value which deviates slightly from the ideal sinusoidal form. They are especially accentuated by the logarithmic

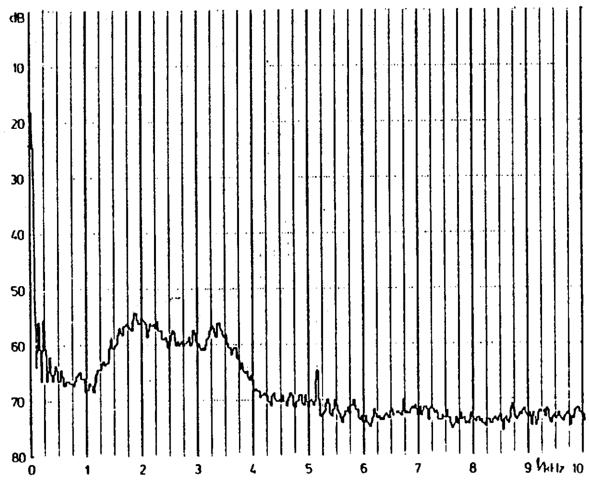

Fig.20: Spectrum of the mains current for phase R, according to Fig.19 top; (x: 1 kHz/D, y: 10 dB/D)

display. The side bands around the triangle frequency and its multiples accordingly consist of contributions of the fundamental frequency and the neighbouring frequencies. In general only harmonics of the order 2k+1 (k=1,2,3,...) are contained in the converter voltage. Due to the not connected AC neutral only currents of the order 6k+-1 can exist. The spectrum of the neutral voltage u_0 only contains harmonics with the order 3k relative to the fundamental frequency. This is caused by the fact that harmonics of the order 6k+-1 again form three-phase systems at the converter input terminals; for these the AC neutral is identical with the DC link center point concerning the potentials. In the spectrum of the converter output current i despite the possible mains current harmonic orders (6k+-1) only harmonics with frequencies 6k times the fundamental frequency are contained. The multiplication of the mains voltage and the harmonic currents of the order 6k+-1 leads to power pulsations of the orders 4j and 6j (j=1,2,3, order 4j are equalized via the converter they do not appear in the DC link. It shall be mentioned that the harmonic amplitudes only in the vicinity of the switching frequencies show up more; the spectrum (of the harmonics) can be said to be shifted by the switching frequency to higher frequencies.

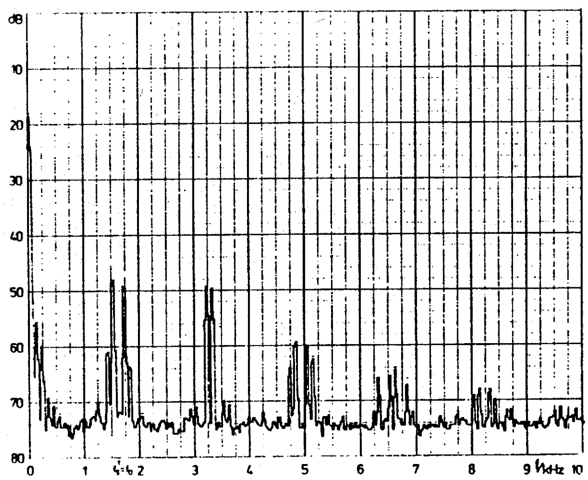

Fig.21: Spectrum of the mains current for phase R, according to Fig.19 bottom (x: 1 kHz/D, y: 10 dB/D)

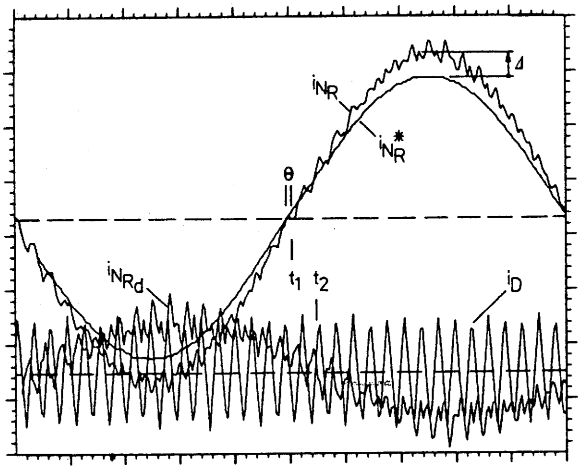

Fig.22: Mains current reference ($i_{N_R}^*$) and actual value (i_{N_R}), control error $i_{N_{Rd}}$ and synchronization current i_D for not compensated counter voltage v_R; Δ, θ amplitude and phase errors; $\omega_N.(t_2-t_1)$: phase shift between mains current and control error fundamental; (i_{N_R}, $i_{N_R}^*$: 10 A/D, $i_{N_{Rd}}$, i_D: 5 A/D, t: 2 ms/D); parameters: u_z=820 V, L=10 mH, i_{TB}=50 mA, f_D=33.f_N

The appearance of a (remaining) control error basically can only be avoided by the following: the converter voltage is formed by a signal which is fed into the controller and acts as a precontrol and which is not formed via a control error. Because the system structure is completely known in this case, the converter voltage simply can be for-

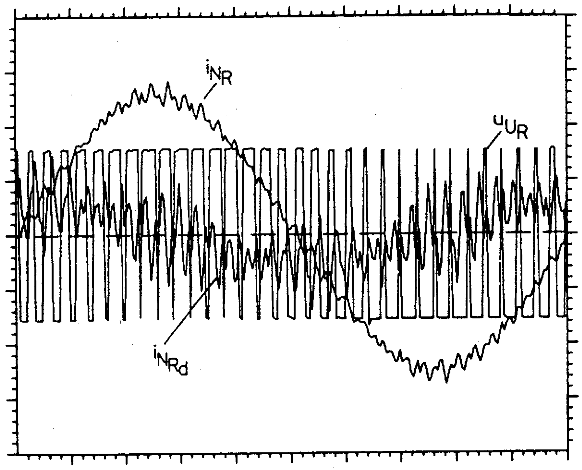

Fig.23: Mains current i_{N_R}, converter voltage u_{U_R} and current control error $i_{N_{Rd}}$ for compensation of the counter voltage contribution u_N ; (u: 250 V/D, i_{N_R}: 10 A/D, $i_{N_{Rd}}$: 2.5 A/D, t: 2 ms/D); parameter as in Fig.22

med based on the relationship $u_{U_R} = u_{N_R} - L.(di_{N_R}/dt) + u_o$. There for i_{N_R} the sinusoidal phase current reference values have to be substituted; this means that u_o is dropped. Therefore $u_{K_R} = u_{N_R} - L.(di_{N_R}^*/dt)$ is valid. The precontrol signal u_{K_R} has to be added under consideration of the gain of the current controller. An incorporation of the signal into the controller output signal is physically impossible due to the discontinuity given there. Because the voltage across the inductance in general is substantially lower than the mains voltage, possibly a precontrol with u_N/k is sufficient. The now remaining part of the control error $i_{N_{Rd}}$ (which is phase shifted by 90° relative to the current) is shown in Fig.23. (For this figure a relatively large inductance of 10 mH has been chosen in order to show this effect clearly.) From the voltage u_{U_R} shown one also can see the controller action as pulse width modulator.

Figures 24 to 28 in general shall illustrate the function of the FCR system. They have been determined for current control by the ramp-comparison controller and precontrol of the mains voltage. u_z is chosen for the different load conditions (Fig.24: resistive, Fig.25: resistive-inductive, Fig.27: resistive-capacitive) such that the controller reaches the limit of sinusoidal modulation. This means that the switching frequency ist just still equal to the triangle frequency. For capacitive behavior thereby the highest DC link voltage value is required (as also results from the considerations of section 2). For purely

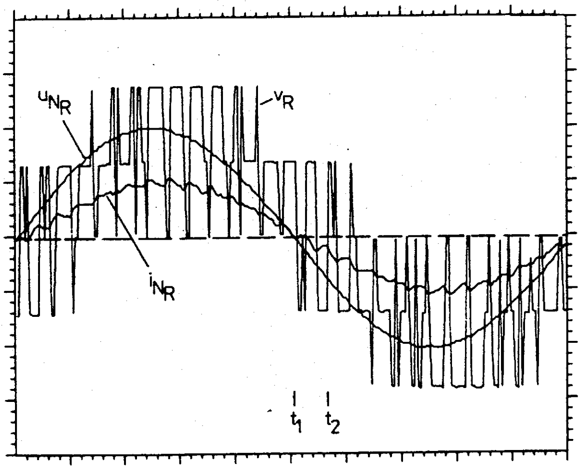

Fig.24: Mains voltage u_{N_R}, counter voltage v_R and mains current i_{N_R} for resistive behavior (as seen by the mains) and for u_z at the modulation limit; partly compensation of the counter voltage (contribution of mains); $\omega_N \cdot (t_2 - t_1)$: phase shift between counter voltage fundamental and mains voltage; (u: 150 V/D, i: 25 A/D, t: 2 ms/D); parameters: u_z=642 V, L=10 mH, $\hat{i}^*_{N_R}$=25 A, i_{TB}=50 mA, f_D=27.f_N.

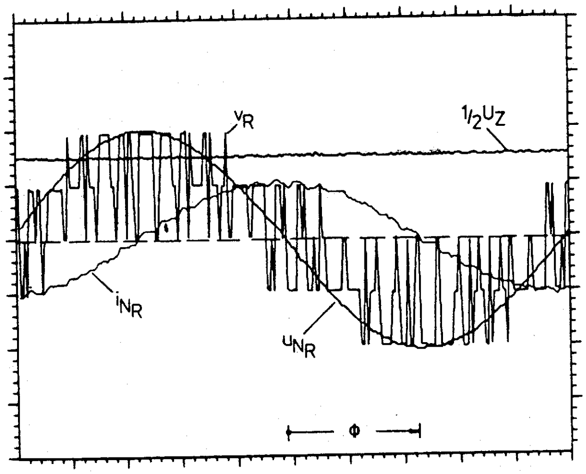

Fig.25: As Fig.24, but resistive-inductive behavior (as seen by the mains; Φ =83°); the modulation limit is reached for u_z=467 V as adjusted here.

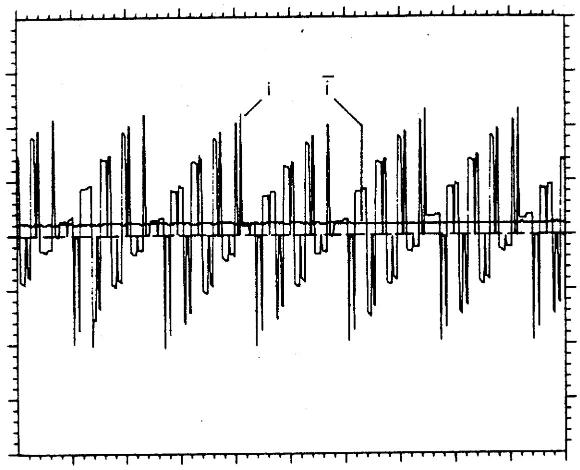

Fig.26: Converter output current i (mean value ī) for Fig.25 (i: 10 A/D, t: 1 ms/D)

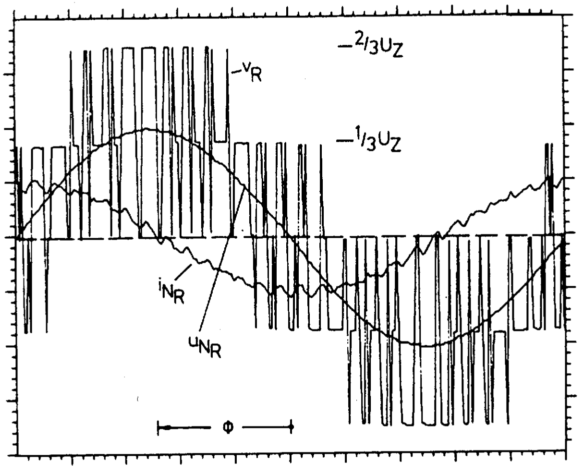

Fig.27: As Fig.24, but capacitive behavior (as seen from the mains, Φ =-83°); the modulation limit is reached for u_z=778 V as adjusted here.

inductive and purely capacitive system behavior the converter output current mean value becomes zero; according to this no real power supply into the DC link takes place. The amplitude modulation of the converter output current with six times the fundamental frequency may not lead to the assumption that this is connected with the occurrence of a harmonic of the same order (this also can be concluded from considerations based on signal theory). On the contrary the modulation leads to the occurrence of side bands of the carrier (being the switching frequency). The deadband given for the bang-bang controller leads to a switching frequency which adjusts itself freely. Because this frequency is given in the case of the ramp-comparison controller the degree of freedom exists in the current control error amplitude. There seems to be a visual difference between the curves shown in Fig.29 concerning the mean switching frequency; in reality this is not the case (as can be seen also from

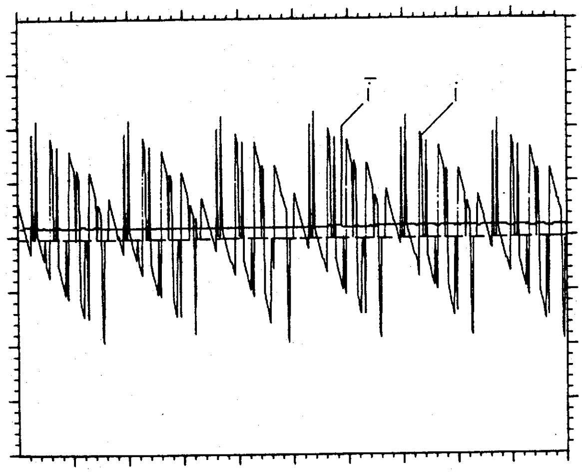

Fig.28: Converter output current i (mean value $\bar{i}$) for Fig.27; (i: 10 A/D, t: 1 ms/D).

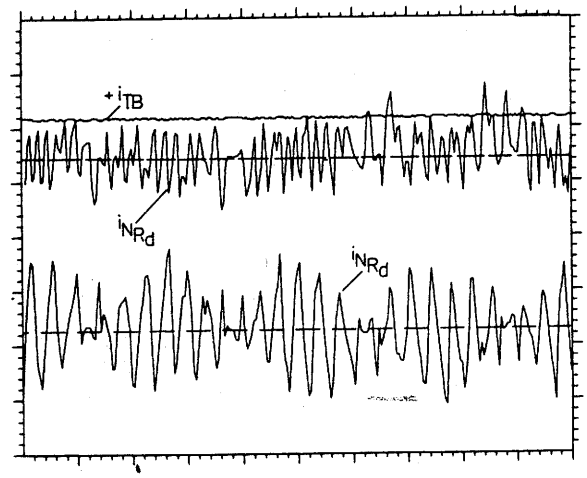

Fig.29: Mains current control error i_{NR_d} for parameters as for Fig.28; below: ramp comparison controller; bang-bang controller with hysteresis i_{TB}=1.82 A (for this value the average switching frequency is equal to f_D); (i: 2.5 A/D, t: 2 ms/D).

Fig.19). The reason is that especially for the bang-bang controller not each change of the current control error in one phase is connected with a switching operation in the corresponding converter branch.

The ramp-comparison controller seems to be substantially better suited as current controller for the FCR system than the bang-bang controller. This is especially valid concerning the stationary behavior.

6. SYSTEM CONTROL BASED ON VECTOR REPRESENTATION OF THE CONTROL VARIABLES

The control of the DC link voltage according to its reference value is achieved by effecting the converter output current. Generally this current consists of the load current and the recharging current of the DC link capacitor. Considering the converter operation a relationship can be derived (via equality of power) between output DC current and mains current (vector) component i_{Nd}. The mains current vector follows from the geometric addition of that given d-component and the arbitrarily selectable q-component (reactive component). The demand for a given converter output current leads to controlling the amplitude of the mains current under consideration of the desired phase relationship between mains current and mains voltage. This leads to a cascade control structure. The control loop for the DC link voltage (outer loop) generates the reference value space vector for the mains current controller (inner loop). The current controller output (space vector of the converter voltage) then determines the switching status of the converter. The inverter voltage vector could also be determined directly from the DC link voltage error. In this case, however, the current, which constitutes a fundamental system variable, would be controlled only indirectly. Because a defined mains current phase position really implies the direct control of the currents, the cascade structure described before is preferable.

As discussed in section 3, the introduction of space vectors simplifies the analysis of polyphase systems. It is obvious to use this vector concept also for controlling the FCR system. Here the rotating rectangular (d,q) reference frame is especially suitable because under stationary conditions all variables will appear as constants in this frame. This also means that remaining control errors are eliminated which would appear for processing alternating signals even if integrating controllers are used. As opposed to analog technique, for microprocessor realization of the controller no major additional effort is introduced by the coordinate transforms now necessary.

If the d-axis is aligned to the mains voltage vector $\underline{u}_N$, the d- and q-components of the system variables physically represent true and reactive components. Therefore the phase position φ_N of the mains voltage vector has to be determined for this control concept.

Economic reasons make it necessary to use a simple converter for realization of the FCR concept. For a power range

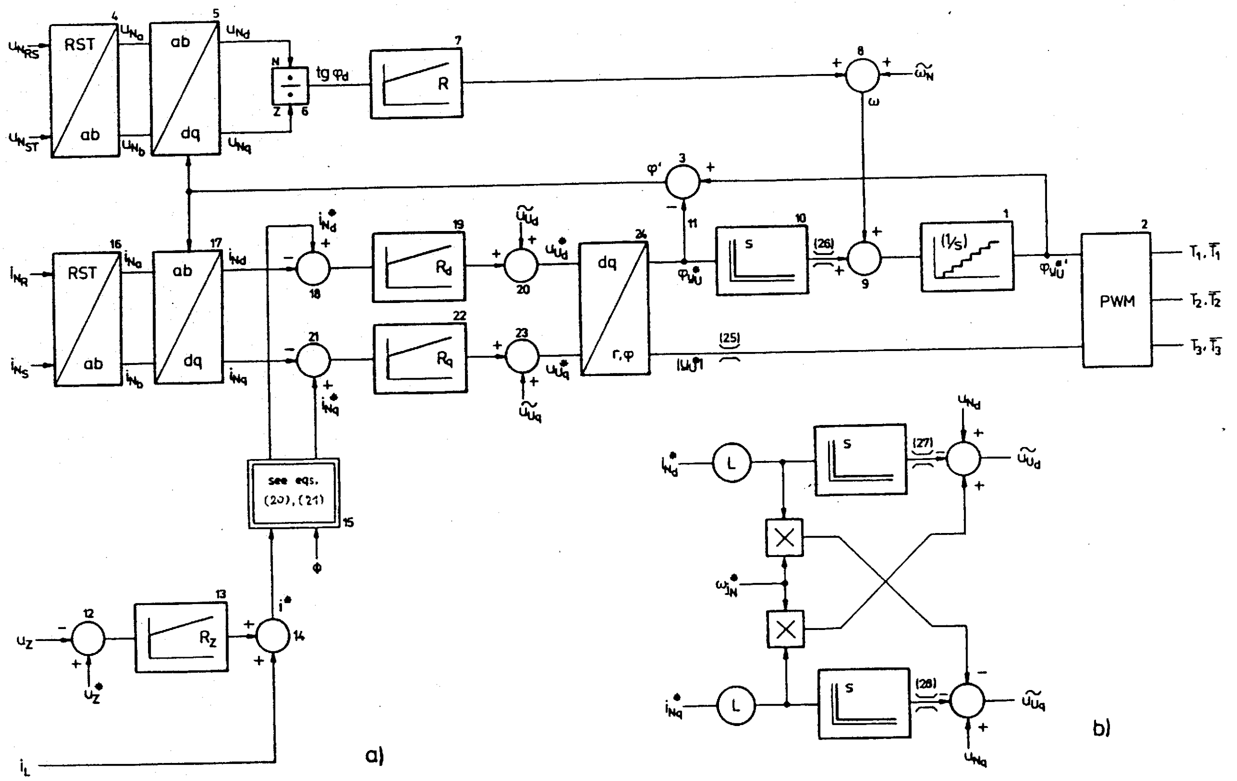

Fig.30: Structure of the vector controller of the FCR system

above 10 kVA then switching frequencies of 1 - 2 kHz are realistic if bipolar transistors without snubbers are used. The resulting ratio of switching frequency to mains frequency forces one to synchronize the converter pulses with the mains period. If this would not be done, critical intermodulation products (e.g., subharmonics of the mains frequency) could appear, as has been already described for the bang-bang current controller in section 4 . It might well be that this is a weak point of all current control concepts for which the converter switching instances are only derived from the system status and therefore the switching is performed in an asynchronous manner with respect to the mains voltage.

As mentioned above, microprocessor control shall be applied. Experience has shown that the cycle time achievable lies in the order of magnitude of the converter switching frequency (about 1 ms). Because the pulse pattern generator (which must be synchronous to the mains period) is part of the controller program, the cycle time of the processor has to be synchronous to the mains period. This also appears advantageous because calculation and execution of the output

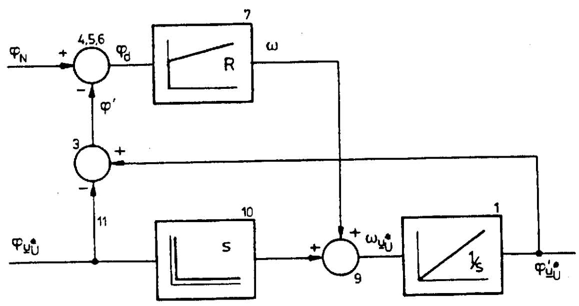

Fig.31: Linearized equivalent circuit of the control loop which generates the converter voltage space vector $\underline{u}_U$ relative to the "measured" mains voltage angle φ_N.

control commands are done in a synchronous manner. This is not possible in control of electrical drives because there the converter pulse period is variable according to the motor speed. Synchronization there then would mean a variable microprocessor cycle time with additional effort for the adjustment of the control parameters (e.g., the integration time of a PI-controller).

An essential advantage of the FCR concept - namely the low harmonic distortion of the mains currents - can be achieved especially when a pulse pattern generator is applied. This makes possible the application of pulse patterns which are optimized with respect to minimum mains interference.

The requirements given above can be realized by the control structure given in Fig.30:

a) Vector Orientation and Converter Control

The essential parts of this control structure are the digital integrator 1 and the pulse width modulator 2. According to the demand for synchronization to the mains frequency this integrator allows only discrete angle increments $2\pi/p$ (p converter pulses during one mains voltage period). According to the converter output frequency ω_U, then ΔT (the time between two angle increments) is calculated for a given value of p. ΔT of course has to be larger than the microprocessor cycle time and determines the sampling time of the whole FCR system. ΔT is only constant for stationary conditions.

The pulse width modulator (PWM) 2 determines the pulse pattern based on the angle information $\varphi_{\underline{u}_U}^{*\prime}$. $\varphi_{\underline{u}_U}'$ has to be given in the fixed (a,b) reference frame because the PWM implies the transformation from the (d,q)- to the (a,b) coordinate system. The absolute value $|\underline{u}_U^*|$ of the converter voltage reference value $\underline{u}_U^*$ determines the degree of the modulation. The angular speed $\omega_{\underline{u}_U}$ of the converter input voltage vector only in the stationary case corresponds to the mains frequency ω_N. The correct expression (even for the dynamic case) is

$$\omega_{\underline{u}_U} = \omega_N + \frac{d}{dt}\psi_{\underline{u}_U}$$

Thereby φ_N is generated (reproduced) by the control loop 1-3-(4 and 5)-6-7-8-9 (Fig.30) which is similar to a PLL (phase locked loop). The mains voltage vector is split up into its d-q-components by the coordinate transformers 4 and 5. Based on these components block 6 determines the tangent of the angle between d-axis (φ') and the direction φ_N of the mains voltage vector. The output of 7 controls the integrator 1 (which forms the controlled sytem of this control loop) such that φ' has to follow φ_N. The fully controlled condition therefore is given by $\varphi'=\varphi_N$ and $\omega=\omega_N$. The orientation $\varphi_{\underline{u}_U}'^{*}$ of the converter voltage $\underline{u}_U^*$ is formed dynamically via 10-9-1 based on the reference value $\varphi_{\underline{u}_U}^*$ (which is given in reference to the d-q-system). Without 11 this would constitute a disturbance for the control loop which reproduces the mains voltage vector position, whereby the converter voltage angle $\varphi_{\underline{u}_U}^*$ would be transferred back into the mains voltage vector angle. This can be eliminated by generation of the transformation angle via block 3 by subtraction of $\varphi_{\underline{u}_U}'^{*}$ (fixed coordinate system) and $\varphi_{\underline{u}_U}^*$ (rotating coordinate system) because then the parallel paths 11 and 10-9-1 compensate each other`s influence on ω in the dynamic as well as in the stationary case.

For small deviations from the fully controlled condition one can linearize Eqs.17 for blocks 4-5-6. This makes possible the description of the resulting system (Fig.31) by Laplace transform. The closed loop response for the two input variables φ_N and $\varphi_{\underline{u}_U}^*$ can be given by

$$\psi_{\underline{u}_U}^{*\prime}(s) = \psi_{\underline{u}_U}^{*}(s) + \psi_A \cdot \frac{1}{(s+R)} \qquad (18)$$

$$\psi'(s) = \psi_N(s) \cdot \frac{1}{(1+s/R)} + \psi_A \cdot \frac{1}{(s+R)} \qquad (19)$$

The initial value (φ_A) of the integrator decreases according to the factor $1/(s+R)$, $\varphi_{\underline{u}_U}^*$ is controlled ideally according to its reference value. As Eq.(18) shows, a change of the reference value of $\varphi_{\underline{u}_U}^*$ does not affect φ'. Equations (18) and (19) determine the system behavior and can be used for selection of the structure and the parameters of the controller R. If a simple proportional controller ($R=k_P$) is applied, we receive $\varphi_{\underline{u}_U}=\varphi_{\underline{u}_U}^*$ and $\varphi'=\varphi_N$ due to the integrating behavior of the controlled system. The controller forms the angular speed ω of the d,q-reference frame. Therefore the controller must have an integrating portion because only then can be formed without a residual control error. However, because the mains frequency is known relatively well (e.g., 50Hz), a simple proportional controller can be used if an estimate $\tilde{\omega}$ is added to the controller output (precontrol, anticipatory control).

b) DC link voltage controller

According to the cascade structure the DC link voltage controller 13 generates the reference value of the mains current vector (or of its components i_{Nd} and i_{Nq}). This is performed by application of Eqs. (20) and (21):

$$i^* = \frac{3}{2} \cdot \frac{1}{u_Z} \cdot (u_{U_d}^* \cdot i_{N_d}^* + u_{U_q}^* \cdot i_{N_q}^*) \qquad (20)$$

$$i^*_{N_q} = i^*_{N_d} . \tan\Phi^* \quad (21a)$$

$$i^*_{N_q} = - u_{N_d} / \omega_N . L^* \quad (21b)$$

$$i^*_{N_q} = u_{N_d} . \omega_N . C^* \quad (21c)$$

Equations (21) determine the desired reactive power behavior. Control according to Eq.(21a) means constant power factor for the mains, whereas application of (21b) or (21c), respectively, would control the reactive component according to a capacitance C or an inductance L, respectively. Relationship (20), which has been derived in section 3 constitutes the transformation of the reference value i* (given by the controller) of the converter output current to the mains side. In the stationary case i is equivalent to the load current i_L; therefore it is precontrolled by the load current via block 14. The controller itself only has to generate that contribution which is necessary to control the capacitor voltage.
If the FCR would be part of a drive system, the precontrol can be done also by the motor controller used in such drives, rather than by the measured current i_L. This means an improvement of the dynamic behavior of the DC voltage link controller because the recharging of the capacitor is initiated before the actual load pulse occurs.
If, however, the DC link capacitor has to keep away load pulses from the mains (which of course requires a sufficiently large capacitance), the precontrol via i_L possibly has to be dynamically weakened - depending on the time constant of the current control loop.

c) Current Controller

The mains current vector is formed similary to the voltage vector by Eqs. (22) and (23) based on the measured values of the phase currents i_{N_R} and i_{N_S} (with $i_{N_R}+i_{N_S}+i_{N_T}=0$) and by transformation into the d,q-system:

$$i_{N_a} = i_{N_R}$$
$$i_{N_b} = \frac{1}{\sqrt{3}} (i_{N_R} + 2.i_{N_S}) \quad (22)$$

$$i_{N_d} = i_{N_a} . \cos\psi + i_{N_b} . \sin\psi \quad (23)$$
$$i_{N_q} = -i_{N_a} . \sin\psi + i_{N_b} . \cos\psi$$

The controllers 19 and 22 give the reference values of the converter voltage components based on the d,q-components of the current control errors. Thereby precontrol via the estimates $\tilde{u}_{U_d}$ and $\tilde{u}_{U_q}$ is performed. $\tilde{u}_{U_d}$ and $\tilde{u}_{U_q}$ have to be determined from the reference values according to the differential equations (Eqs.(12)) of the FCR power circuit (see Fig.30b). The coordinate transformer 24 computes absolute value and phase angle of the converter voltage according to Eq.(24):

$$|\underline{u}^*_U| = \sqrt{u^{*2}_{U_d} + u^{*2}_{U_q}} \quad (24)$$

$$\psi_{\underline{u}_U} = \arctan u^*_{U_q} / u^*_{U_d}$$

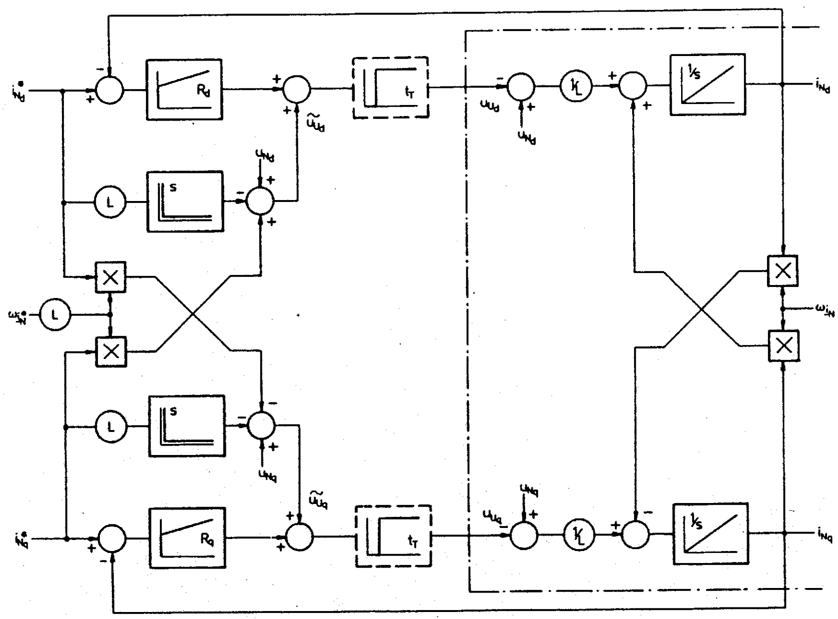

Fig.32: Equivalent circuit (system) of the inner current control loop; dotted line: system to be controlled.

d) Treatment of Limitations

The limiters 25-28 prevent the control loop elements against saturation, "wind-up" and overflow (especially of the differentiators in case of reference value steps). For this purpose they act upon the generation of the reference values by 13 and 15. The reference value changes can then be only that fast that saturation is prevented.

e) Dimensioning of the controllers

For dimensioning of the d,q-axis current controller we start from the structure shown in Fig.32. It consists of the controlled system, of the two current controllers and of the part which serves for the precontrol of the converter voltage. If one neglects the converter delay time t_T (indicated in Fig.32) and if one assumes an ideal precontrol, the couplings cancel. Then the structure is split up into two independent structures as shown in Fig.33. Calculating the system transfer function for reference value variations for these two control system parts one can note that the controllers R_d and R_q are not contained therein any more. This can be explained by the fact that the controller does not have to make a contribution to the quantities controlling the system any more; this is due to the fact that in the ideal case the system is controlled only by the precontrol feature. Therefore the controller basically only serves to control disturbances and control errors which are caused by the neglections and simplifications mentioned. Then the controller can be optimized with respect to control of external disturbances. Therefore it seems to be advantageous to apply the "symmetric optimum" as optimization basis. Thereby one can do without the usual reference value smoothing because the system will be guided by the precontrol feature, the control error then will remain zero in this ideal case. For the controller dimensioning one has to define a sum of the small time constants: for this sum one can use a time constant in the order of magnitude of the converter delay time; to this a time constant corresponding to the current smoothing (if smoothing is applied) has to be added. As a first approximation one can represent the current control loop by a first order delay; this has to be considered in the outer (master) control loop for the DC link voltage. Figure 34 shows the equivalent circuit of this control loop. If one assumes that the precontrol - this time by the load current i_L - is ideal, the influences caused by i_L can be neglected. It is advantageous to dimension the DC link voltage controller also based on the "symmetrical optimum" where the current controller equivalent time constant T_{SR} has to be used as the required sum of small time constants. Homever, if the reference value u_z^* can change immediately, one has to provide reference value smoothing with $4.T_{SR}$.

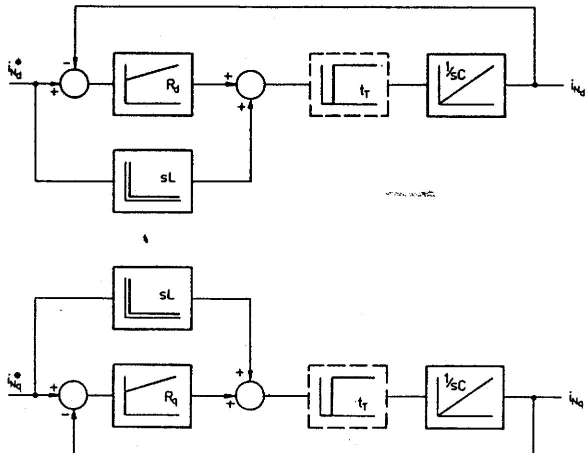

Fig.33: Equivalent circuit of the current control circuits for the d,q-coordinate system as decoupled by the voltage precontrol

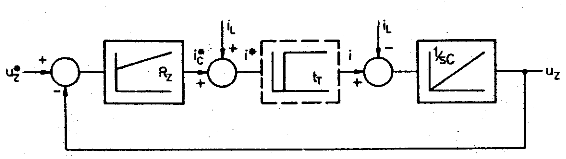

Fig.34: Equivalent circuit of the DC link voltage control circuit

7. C O N C L U S I O N S

As the research results of this paper show, by application of a PWM inverter working as rectifier (under consideration of the high development status of power electronics and information electronics) an avoidance of the drawbacks of the classical rectifier principle seems to be possible, whereby operational safety and economical operation are maintained. In general for the developement of new converter systems, one has to stress mainly the effort connected with practical realization. Furthermore it is not reasonable, to derive the system behavior based on idealizations (which are necessary for a clear system description) and to compare this (ideal) system behavior with that of a system which has been practically realized. It is also not permissible for studying the cooperation of (in general) a power electronic system with other systems (e.g., energy distribution systems) to assume the power electronic system as ideal and to derive its non-interaction based upon this. Under these "additional conditions" (which should gain closer attention in general) the proposal of an "ideal converter system" seems to be made substantially more difficult. Also under consideration of all the points mentioned here (especially concerning the conditions of practical realization) the FCR system proposed represents - for the power range above 20 kVA - the almost optimum input system for an electric drive with DC voltage link.

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Some of the references are not especially mentioned in the text, but had to be listed here because they constitute an important addition to some of the points of this paper.

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