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A Novel Three-Phase Converter Topology for Supplying Wide Speed Range Induction Machine Drives

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Abstract. The paper presents an inverter concept for the realization of an induction machine drive with an extremely wide constant power range. There, the stator winding is split up into two isolated and bifilar wound three-phase systems. The series operation of the two winding sets permits very high torque at low stator frequencies while keeping the machine terminal currents and the inverter DC link current relatively low. However, due to the winding arrangement field weakening has to start already at low frequencies (approximately at 50% of rated speed). Nevertheless, with the help of a transition from series to parallel winding arrangement the machine flux can be restored to its nominal value at rated machine speed. Consequently, the breakdown torque is increased by a factor of four which gives the basis for a high maximum speed with constant machine power.

1 Introduction

The considerations given in this paper are directed at the realization of a field-oriented controlled induction machine drive with high starting torque and wide constant power range [1]

$$\begin{array}{rcll} M & = & 250\,\mathrm{Nm} & (0\ldots 500\,\mathrm{r/min}) \\ P & = & 4000\,\mathrm{W} & (500\ldots 6000\,\mathrm{r/min})\,. \end{array}$$

For fulfilling these requirements basically the possibilities exist:

1. oversizing of the machine (considering the quadratic dependency of the breakdown torque on the stator frequency as valid also for the mechanical speed $P = 4000\,W@n = 6000\,\mathrm{r/min}$ results in a breakdown torque requirement of $M_K \approx 920\,\mathrm{Nm}$ at $n = 500\,\mathrm{r/min}$. Since a practicable machine dimensioning only can be performed for ratios of rated torque and breakdown torque lower up to $\frac{M_K}{M_N}|_{\max} \approx 4$, one, therefore, would have to provide a machine having a rated torque of $M_N = 230\,\mathrm{Nm}$. However, $P = 4000\,\mathrm{W}@n = 500\,\mathrm{r/min}$ only corresponds to a torque requirement of $M \approx 80\,\mathrm{Nm}$. This solution, therefore, shows clear disadvantages concerning machine size and costs.)
2. selecting $n = 1000\,\mathrm{U/min}$ as lower bound of the constant power operating range and/or for speed-proportional decrease of the output power down to $n = 500\,\mathrm{r/min}$, (for $n = 1000\,\mathrm{r/min}$ we have a relatively good correspondence of the breakdown torque requirement $M_K = 230\,\mathrm{Nm}$ due to $4000\,\mathrm{W}@6000\,\mathrm{r/min}$ with the required starting torque; as rated power one would there have to provide $P_N \approx 6000\,\mathrm{W}$, corresponding to $\frac{1}{4}M_K$ and/or for making the dimensioning of the machine feasible.)
3. reconfiguration of partial stator windings and/or constant power operation in the full mechanical speed range 500....6000 r/min (in this case the dimensioning of the machine has to be performed as for 2.)

The requirement of a high breakdown torque at $n = 500\,\mathrm{U/min}$ for allowing $P = 4000\,\mathrm{W}$ at the upper me-

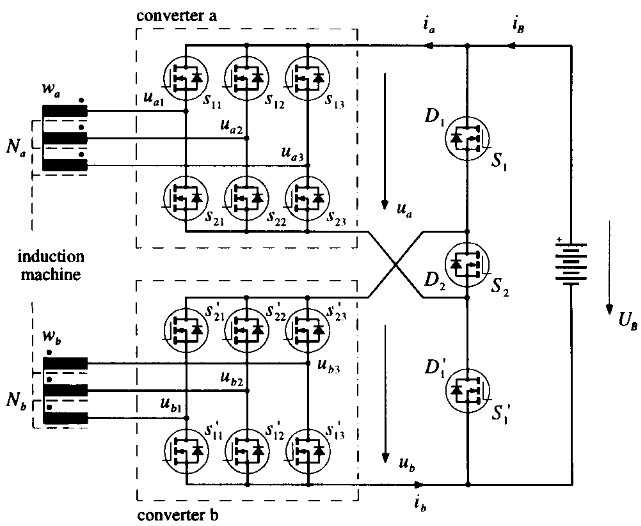

Fig.1: Basic structure of the power circuit of the proposed converter topology. For series operation S_2 is in the on-state and S_1 and S_3 remain in the off-state; accordingly S_2 is turned off and transistors S_1 and S'_1 are turned on for parallel operation. The corresponding phase windings of the partial winding systems w_a and w_b are arranged in a bifilar manner.

chanical speed boundary latterly is caused by the speed-proportional reduction of the machine flux in the field-weaking range.

Therefore, in general a changing of the winding configuration is directed to an increase of the flux level in the upper speed range. For this in the literature

- a changing of the number of poles (e.g, reducing the number of poles by a factor of 2) beyond a given speed value [2] or a
- reduction of the effective number of turns at high speeds by limitation to using tapped turns or re-connection of partial winding systems from series to parallel operation) [1], [3]–[5]

has been proposed. A very elegant concept for electronic (contactless) pole changing has been proposed and analyzed in [6]. A series-parallel reconfiguration of partial stator winding systems has been treated in the literature so far only with using electromechanical contactors, however. Drawbacks of this concept consist in the occurrence of a changeover interval with zero current and/or zero torque and in the low reliability of the change-over switches.

As this paper show, series-parallel-winding reconfiguration, however, also can be achieved by electronic means (cf. **Fig.1**). There, a PWM inverter is assigned to each partial winding system w_a and w_b and a series or parallel operation of the partial winding systems is achieved by a series or parallel operation of the converters a and b. There, for arranging the change-over devices on the DC side as opposed to an AC side arrangement only a unipolar blocking voltage stress on the turn-off power semiconductors S_i, $i = 1, 2, 3$, occurs. Furthermore, the valves S_i then can be realized by power semiconductors having only a unidirectional turn-off capability, which results in a relatively low realization effort of the change-over unit. However, then for series *and* parallel operation the main current flow is always via change-over devices (S_1, S_1' for parallel operation, S_2 for series operation). According to section 4 the thereby resulting increase of the system conduction losses remains limited to relatively low values as compared to the conduction losses of converters a and b, however.

2 Basic Principle of Operation

For describing the basic principle of operation of the novel converter system in the following the conduction states occurring for series and parallel operation are discussed briefly. There, for the sake of clearness we do not distinguish between current conduction of a transistor or of the associated antiparallel free-wheeling diode. Furthermore, the change-over from series to parallel operation and a control concept for eliminating the machine common-mode voltage are treated (cf. section 2.3).

2.1 Current- and Voltage Conditions for Series and Parallel Operation

A series operation of w_a and w_b is achieved by turning on S_2 and by applying identical control signals (i.e., control signals forming equal space vectors of the converter output voltages) to converters a and b. Thereby, the negative input voltage rail of converter a is connected with the positive input voltage rail of converter b independent of the direction of the current flow and the power transistors S_1 and S_1' are remaining in the off-state. A simultaneous conduction of S_2, S_1 and S_1' would cause a short circuit of the supplying battery voltage U_B. (The switching state of a power transistor S_i is characterized in the following by a binary switching function $s_i = 0, 1$; there, $s_i = 1$ corresponds the on-state and $s_i = 0$ to the off-state). The conduction state corresponding to $S_2 = 1$, $S_1 = S1' = 0$ is shown in **Fig.2**(a) for an active, i.e. voltage forming switching state. Considering the symmetry being given due to the bifilar arrangement of the windings of corresponding phases we, therefore, have for the input voltage of converter a and b

$$u_a = u_b = \frac{1}{2}U_B \tag{1}$$

and/or the partial windings are connected in series independent of the direction of the battery current i_B (current flow is via S_2 or the associated free-wheeling diode). Parallel operation is achieved by turning off S_2 and turning on S_1 and S_1' (cf. Fig.2(b)). Then the input voltage of converter a and converter b is defined by the total value of the battery voltage,

$$u_a = u_b = U_B \ . \tag{2}$$

According to the bidirectional current carrying capability given for the power transistors in connection with the associated free-wheeling diodes also the parallel connection is not dependent on the direction of the DC side current flow. Due to the bifilar arrangement (and/or close inductive coupling) of the corresponding phases of the partial winding systems, the synchronous switching of the converters according to

$$s_i = s_i' \tag{3}$$

has to be maintained also for parallel operation in order to avoid the occurrence of differential currents of high amplitude in the winding systems. For free-wheeling of the converters we have identical conduction states for parallel and series operation due to the missing DC side current flow (cf. Fig.2(c)).

Considering different switching times of the gating units and/or of the valves as given in practice also the conduction states occuring for not simultaneous switching of the converters are of interest. For the sake of brevity we here only would like to treat the transition from a free-wheeling state into an active voltage state as example under the assumption of a low switching delay of the valves S_{13}' and S_{23}' (cf. **Fig.3**). For parallel operation subsequent to changing the switching state the current flow is via D_1 or S_1 dependent on the sign of i_b, i_a remains at 0 due to the continuance of the free-wheeling

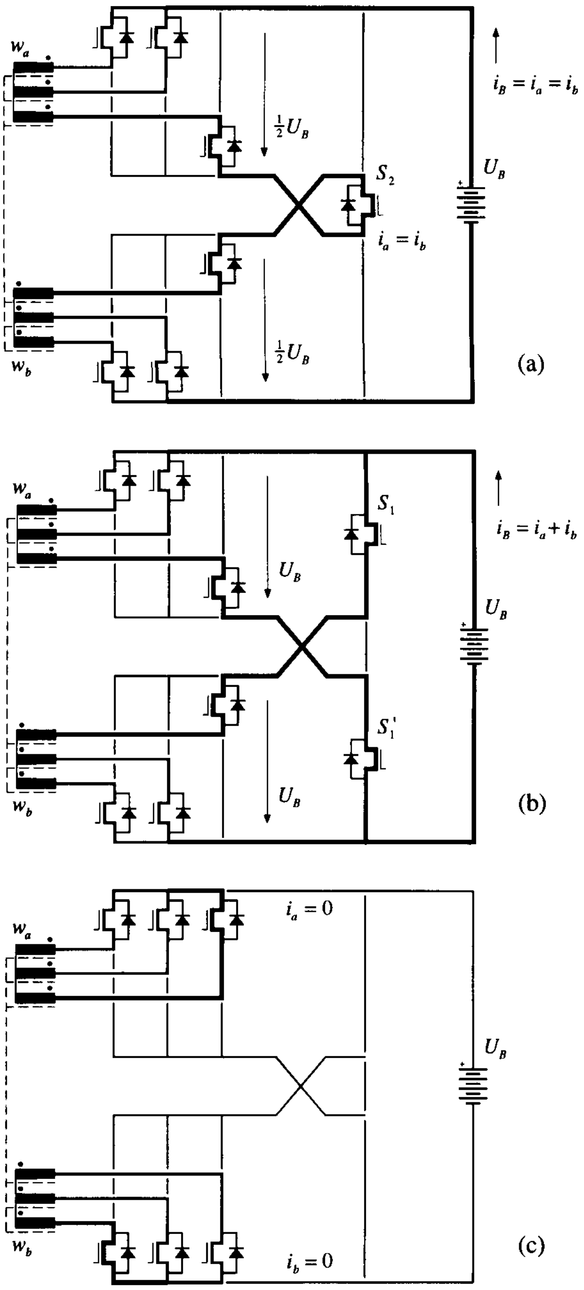

Fig.2: Conduction states of the converter system for series **(a)** and parallel operation **(b)** for an active (voltage forming) switching state, **(c)**: free-wheeling state.

of converter a. Therefore, the battery voltage is applied only to the input of converter b. This is also true for series operation, i.e., $s_2 = 1$ and $i_b < 0$ (current flow via D_1 or, alternatively, via S_2 and the free-wheeling diodes of the power transistors of converter a). Only for $i_b > 0$ converter b remains in the free-wheeling state independent of the switching state of S_2 and S_1' via D_2 and D_1' until switching over of the bridge leg S_{13}, S_{23}. As already mentioned different terminal voltage of the partial windings result in differential currents of high amplitudes due to the bifilar arrangement of corresponding phases of w_a and w_b. Therefore, one has to employ gate control cir-

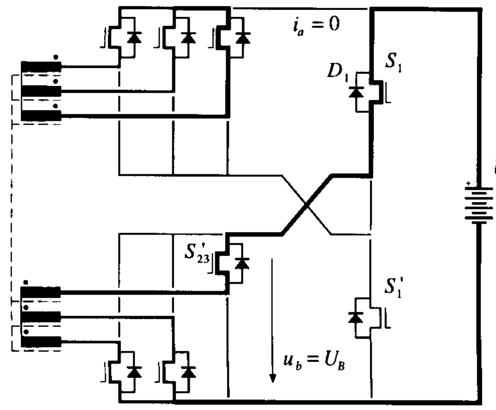

Fig.3: Conduction state for not synchronous switching of corresponding valves of converter a and b and parallel operation.

cuits and power transistors with sufficiently low switching delay times. Finally, it is important to note that the voltage at the converter inputs (ideally) due to the topology of the change-over unit is limited to the battery voltage also for not-synchronous switching of the converters.

Remark: Basically the change-over unit also could be controlled indirectly by the direction of the main power flow. Then, the switches S_i can be omitted and only the free-wheeling diodes D_1, D_1', D_2 and filter capacitors on the DC side of the converters have to be provided. These capacitors allow a transient inversion of the converter input currents as occurring for low power factor operation without changing of the direction the main current flow in the change-over unit. Thereby, a change-over between series and parallel operation with pulse frequency can be avoided. However, the series operation is limited to motor operation of the drive and parallel operation only can be achieved for feeding back of energy into the battery (generator operation). Further realization possibilities with extended controllability can be derived by adding S_2 in antiparallel to D_2 (series operation possible for motor *and* generator operation) or by adding power transistors S_1 and S_1' in antiparallel to D_1 and D_1' (parallel operation possible for generator *and* motor operation).

Furthermore, we would like to point out that for series operation and $i_a, i_b > 0$, as described above, for changing the switching state of, e.g. only converter a to free-wheeling operation also converter b is forced into the free-wheeling state independent of its actual switching state (current flow via D_1' and D_2). However, switching only converter a (or b) into the free-wheeling state and leaving b (or a) in the active switching state which again has to be applied subsequent to free-wheeling operation, no reduction of the total converter switching losses can be achieved. This is due to an increase of the turn-off voltage occuring for converter a (or b) by a factor of 2; the turn-off voltage is defined by the total battery voltage value as opposed to simultaneous switching of the converters where half the battery voltage occurs as turn-off voltage for each converter system.

2.2 Change-Over from Series to Parallel Operation

Changing-over from series to parallel operation (or vice versa) is performed advantageously at zero current, i.e. in the free-wheeling state of converters a and b (cf. Fig.2(c)). For space vector modulation or carrier-based modulation, therefore, at each beginning or end of a pulse period the possibility of changing the switching state is given.

2.3 Control of the Converters

As already mentioned in section 2.1, the converters a and b are pulsed in an inverse manner $s_i = s_i'$. Therefore, for the winding systems w_a and w_b (ideally) common-mode voltages $u_{0,a}$ and $u_{0,b}$ of equal magnitude but opposite sign

$$u_{0,a} = \frac{1}{3}(u_{a1} + u_{a2} + u_{a3}) = -\frac{1}{3}(u_{b1} + u_{b2} + u_{b3}) = -u_{0,b} \tag{4}$$

occur (the voltages are referred to a fictitious center point of the battery voltage) and/or transient changes of the common-mode voltage levels occurring for switching state changes of the converters are of equal magnitude but in opposite direction. This (ideally) results in zero shaft voltage and in a mutual cancellation of high frequency common-mode interference currents (and/or of bearing currents occuring due to the capacitive coupling between the stator windings and the rotor [7, 8]). In order to produce additive magnetic flux in the same direction the orientation of, e.g., partial winding system w_b has to be reversed with respect to w_a, as shown in Fig.1.

3 Control Concept

As **Fig.4** shows, the series connected stator winding configuration is used in the lower speed region in order to limit the battery current to low values also for operation close to the breakdown torque limit. Constant power operation also starts with series connected windings w_a and w_b (S_2 is in the turn-on state, S_1 and S_2 are switched off). However, above about 2000 r/min a constant power of 4000 W requires a machine torque beyond the breakdown limit (cf. upward pointing arrow in Fig.4). The torque capability can be regained by changing to parallel operation of w_a and w_b by turning off S_2 and turning on S_1, S_1'. Thereby, for constant converter modulation depth the volts per turn are doubled and/or the machine flux is increased by a factor of 2. Accordingly, the breakdown torque raises by a factor of 4, which permits 4000 W operation up to 6000 r/min.

For braking operation starting at high speed a reconfiguration to series connection is possible at 2000 r/min. In any case the transition has to be performed at 1000 r/min where the machine flux reaches its rated value. A transition from series to parallel connection is equivalent to cutting the machine flux in a half for constant converter

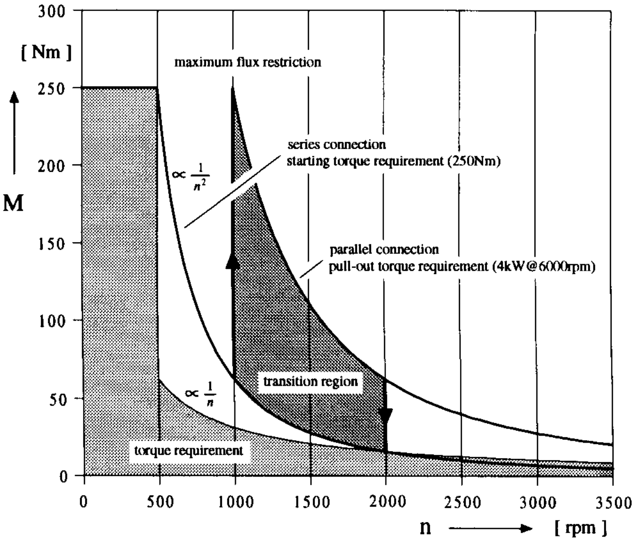

Fig.4: Breakdown torque requirements for series- and parallel operation of the partial stator winding systems according to specifications of starting torque and constant power operating range given in section 1 (pointed out by the dotted area). Transition from series to parallel operation (or vice versa) has to be performed within the transition area.

modulation depth. Going below 1000 r/min in parallel connection with limited flux level would require current values above rated current for 4000 W operation due to the increasing torque demand.

A transition from parallel to series connection is allowed only after an appropriate flux level reduction. Otherwise the machine back-emf would exceed the maximum inverter output voltage resulting in an overcurrent condition. On the other hand the transition from series to parallel connection does not require a prior flux level adjustment. However, the flux level has to be adjusted in time for high speed dynamics in order to avoid a violation of the breakdown torque restriction.

In general, any transition between the two winding configurations can be performed at any speed in between 1000 r/min and 2000 r/min. There, one has to consider that the dynamics of flux level changes for controlling the machine current are governed by the rotor time constant, which typically shows values in the range of 100 ms.

In the case at hand the adaptation of the flux level is performed whenever speed exceeds the center value of the transition range. The changeover is thereby done with limited dynamics in order to keep the necessary inverter current within a technically acceptable range. The transition is done with a ramp function with respect of time where the time constant is determined by the rotor time constant of the machine. The passing from one desired flux to another can be done seamlessly in both directions. In case of low mechanical acceleration it can be completed within a small change in speed. In case of very high acceleration and a corresponding very fast passing of the transition zone the required amount of flux change is lower compared to the first case due to the rapid change of the desired flux values. This fact leads to an advantageous characteristic with low control signal demand if the ramp time is set properly. The ramp changeover scheme then shows an almost ideal flux adaptation characteristic.

A digital simulation of the system dynamics including the inverter current controller and the induction machine has show that the change in the winding configuration can be done without any torque disturbance due to the principle of field oriented machine control [9]–[11]. The system is not stressed by an abrupt step in the flux level as a consequence of the continuous flux control during the transition. This feature is, e.g., in contrast to electronic pole changing (which is characterized by the requirement of an abrupt flux level change) and to a mechanical switch solution, which always requires a zero current gap. The instantaneous change of the DC voltage with respect to the machine fundamental after a change in the series/parallel arrangement can be handled safely by the inner current control loop. Torque and power control stay almost unaffected. However, in order to keep the distortion as small as possible the transition should be synchronized to the current controller sampling. The current controller then immediately uses an adapted switching sequence for the next sampling interval and the distortion remains ideally zero.

Remark: Aiming for low losses of the machine one would have to prefer parallel operation to series operation within the transition area. Due to the higher breakdown torque there results a lower slip frequency for a given torque and the rotor losses (being directly proportional to the slip frequency) are reduced correspondingly.

4 Stresses on the Components, System Evaluation

For a first evaluation of the proposed converter system besides control oriented aspects especially the conduction and switching losses of the partial converters a and b, the relative conduction losses of the change-over devices S_1, S_1' and S_2 (as related to the losses of the partial converters), and the current stress on the supplying battery are of interest for series and parallel operation.

In the following these characteristic values are calculated for operation of the drive system at low speed and high torque (series operation of the partial windings) and are compared to the component stresses given for application of a single converter and/or for not partitioned stator winding (this is equivalent to a parallel operation of the partial windings).

4.1 Assumptions

In order to limit the considerations to the essentials for the calculation of the component stresses in a first step all power transistors are considered as having a purely ohmic on-state behavior and/or the participation of the parasitic anti-parallel free-wheeling diodes in the current flow is neglected. This assumption is valid with good accuracy for low battery voltages U_B due to the low on-resistance of modern power MOSFETs with low blocking capability. Therefore, the on-characteristic of the valves is independent of the direction of the current flow and/or the conduction losses are not influenced by the modulation index and the phase displacement of the converter output voltage and current fundamentals. Furthermore, the battery voltage is assumed to be impressed, i.e., the inner impedance of the battery (and/or the equivalent series resistance of an electrolytic capacitor being provided for filtering of DC link current components with switching frequency) is neglected. Also, for calculating the switching losses an ideal simultaneous switching of corresponding valves of the converters is assumed.

4.2 Operation at Low Speed and High Torque

In order to make possible a direct comparison of the proposed series connection of the partial windings with supplying the machine by a single converter for the further considerations each valve of the single converter is thought to be realized by a parallel connection of two individual transistors (and/or a parallel operation of two simultaneously switching partial converters is assumed). If now for realizing a transistor of partial converter a and partial converter b by a single valve (with on-resistance $R_{DS(\mathrm{on})}$) both systems show identical realization effort of the power circuit with exception of the change-over devices. (In connection with this we, however, would like to point out that for operating the machine with a single converter a constant power operation cannot be achieved in the full speed range for using identical AC machines and equal battery voltages, cf. section 1, item 2.)

In case an operation of the drive with given torque, given mechanical speed and rated flux is assumed, we basically have equal currents in the partial windings and equal fundamental displacement factors $\cos\varphi$ for series operation and single converter and/or parallel operation of the stator winding systems. For the ratio of the modulation indices we, therefore, have according to $u_a = u_b = \frac{1}{2}U_B$,

$$M_\mathrm{s} = 2M_\mathrm{p} \qquad M_\mathrm{p} = \frac{\hat{U}_U}{\frac{1}{2}U_B}\,, \tag{5}$$

($\hat{U}_U$ denotes the amplitude of the fundamental of the voltage applied to a partial winding system).

4.2.1 Converter Conduction and Switching Losses

Independent of series and parallel operation each valve is conducting a partial winding current $i_{a,i}$ or $i_{b,i}$, therefore, we have equal total converter conduction losses

$$P_\mathrm{c} = P_\mathrm{c,s} = P_\mathrm{c,p} = \frac{3}{2} I_{N,\mathrm{rms}}^2 R_{DS(\mathrm{on})} \tag{6}$$

in both cases ($I_{N,\text{rms}}$ denotes the rms value of the total machine phase current for single-converter/parallel operation).

Regarding the switching losses P_p the series operation shows advantages, since under the assumption of a symmetrical split-up $u_a = u_b = \frac{1}{2}U_B$ of the battery voltage to the partial converters the turn-off voltage of the valves is cut in a half. We therefore have

$$P_{\text{p,s}} = \frac{1}{2}P_{\text{p,p}} \,. \tag{7}$$

4.2.2 Conduction Losses of S_2

According to [12] the current stress on S_2 can be calculated as

$$I^2_{S_2,\text{rms}} = I^2_{N,\text{rms}}[\frac{\sqrt{3}M_s}{2\pi}(\frac{1}{4} + \cos^2\varphi)] \,, \tag{8}$$

and, therefore, is dependent on the modulation index and the fundamental displacement factor. If under consideration of the voltage drop occurring across the stator resistance for high current and/or high torque $M_p \approx 0.9$ and $\cos\varphi \approx 0.7$ are assumed, there follows $I^2_{S_2,\text{rms}} = 0.09I^2_{N,\text{rms}}$. In case S_2 also is realized by a valve with on-resistance $R_{DS(\text{on})}$, therefore, the system conduction losses are increased by only about $0.06P_\text{c}$.

4.2.3 Current Stress on the Battery

We would like to point out that the current stress on the battery (and/or an electrolytic DC link capacitor) is significantly reduced for series operation as compared to single converter and/or parallel operation.

As is immediately clear the peak value of the battery current occuring in a pulse period is cut in a half

$$\hat{i}_{B,\text{s}} = \frac{1}{2}\hat{i}_{B,\text{p}} \tag{9}$$

for series operation.

Remark: The current supplied by the battery in the average over a pulse period is equal for series and parallel connection, however. For a given machine torque and given speed equal power is supplied to the machine for both modes of operation; considering the different modulation index for parallel operation the battery current is formed by current pulses showing about half width and an amplitude being higher by a factor of 2 as compared to series operation (cf. **Fig.5**).

For the rms value of the battery current we have for series operation [12]

$$I_{B,\text{rms,s}} = I_{S_2,\text{rms}} = I_{N,\text{rms}}\sqrt{\frac{\sqrt{3}}{2\pi}M_s(\frac{1}{4} + \cos^2\varphi)} \,, \tag{10}$$

for parallel operation there results a current stress [12]

$$I_{B,\text{rms,p}} = I_{N,\text{rms}}\sqrt{\frac{2\sqrt{3}}{\pi}M_p(\frac{1}{4} + \cos^2\varphi)} \,, \tag{11}$$

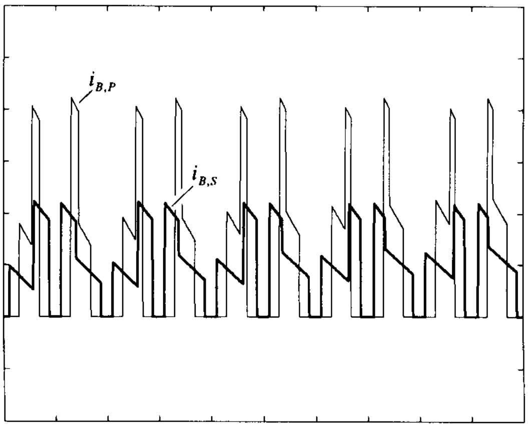

Fig.5: Time behavior of the battery current i_B for $M = 0.9$ and $M = 0.45$ and/or for series and single converter (parallel) operation of the partial winding systems ($\cos\varphi = 0.45$).

being higher by a factor of $\sqrt{2}$, resulting in a doubling of the battery losses. If again $M_\text{s} = 0.9$ ($M_\text{p} = 0.45$) and $\cos\varphi = 0.7$ are assumed and if the inner resistance of the battery is assumed being, e.g., equal to the on-resistance $R_{DS,(\text{on})}$ of a valve, battery losses of about $P_{B,\text{s}} \approx 0.13P_\text{c}$ will occur. The corresponding reduction of the losses by $0.13P_\text{c}$ results in a significant increase of the efficiency of the total system.

4.2.4 Current Stress on a DC Link Capacitor

If for filtering of switching frequency harmonics of the DC link (battery) current an electrolytic capacitor is provided we have for the capacitor current stress

$$I_{C,\text{rms,s}} = I_{N,\text{rms}}\sqrt{\frac{M_\text{s}}{2}[\frac{\sqrt{3}}{4\pi} + \cos^2\varphi(\frac{\sqrt{3}}{\pi} - \frac{9}{16}M_\text{s})]} \,, \tag{12}$$

or

$$I_{C,\text{rms,p}} = I_{N,\text{rms}}\sqrt{2M_\text{p}[\frac{\sqrt{3}}{4\pi} + \cos^2\varphi(\frac{\sqrt{3}}{\pi} - \frac{9}{16}M_\text{p})]} \,. \tag{13}$$

Therefore, for an equivalent capacitor series resistance being, e.g., equal to the on-resistance of a valve, $R_\text{ESR} = R_{DS,(\text{on})}$, for $M_\text{s} = 0.9$ and $\cos\varphi = 0.7$ capacitor losses of $P_{C,\text{s}} = 0.05P_\text{c}$ and/or a reduction of the losses by $0.12P_\text{c}$ is given for series operation as compared to single converter operation.

In summary, the increase of the converter conduction losses due to S_2 (series operation) is by far compensated by lower switching losses and lower losses in the DC link energy storage as compared to parallel operation. Furthermore, the peak current stress on the DC voltage source is reduced by a factor of 2. Therefore, the avoidance of a restriction of the constant power operation to speed values above 1000 r/min is only paid for by a higher gate drive and a higher assembly effort concerning the converter circuit but not by higher system losses or a higher stress on the power components.

5 Conclusions

The paper presents a novell inverter concept for the realization of an induction machine drive with a very wide constant power range. There, the stator winding is split up into two isolated and bifilar wound three-phase systems. The serial connection of the two winding sets permits very high machine current density and consequently high torque at low stator frequencies while keeping the machine terminal currents and the inverter DC link current relatively low. However, due to the winding arrangement field weakening has to start already at lower frequencies (approximately at 50% of rated speed). Nevertheless, with the help of a transition from series to parallel winding arrangement the machine flux can be restored to its nominal value at rated machine speed. Consequently, the breakdown torque is increased by a factor of four which gives the basis for a high maximum speed with constant machine power. Before the transition from parallel to series connection the machine flux has to be controlled to 50inverter output voltage and current control is lost. A high current peak and torque distortion would result. The transition from one winding arrangement to the other is done always via a time determined ramp function. The correspondent flux and current dynamics can be handled safely by a field oriented machine control.

In the following the advantages and drawbacks of the system shall be compiled briefly.

Advantages:

+ relatively low rated power/size of the induction machine for high starting torque and wide speed range with constant power operation
+ reduction of the peak current stress on the battery by a factor of 2 at low speed (as compared to single converter operation and/or limited constant power range)
+ no mechanical contacts for changing the winding configuration
+ low common-mode noise for inverse gating of the partial converter systems a and b
+ converter topology also applicable for extending the speed range of other types of AC machines, e.g., permanent magnet synchronous machines [13] etc.

Drawbacks:

- as compared to single converter operation and conventional winding techniques higher rated power of the converter
- relatively high effort for controlling the valves
- increased realization effort of the machine, higher number of motor terminals (of low importance for mechanical integration of converter and machine)
- lower efficiency in the upper speed range and/or for parallel operation) due to the losses occuring in the change over switches S_1 and S_1'.

Further research will be on an experimental analysis of the proposed system. There, in a first step two identical mechanically coupled individual machines (and/or two magnetically decoupled partial winding systems) shall be considered. Besides verifying the theoretical considerations given in this paper there also the potential of instabilities of the individual converter current controls [14] operating in parallel for parallel winding configuration shall be analyzed. Considering the fact that due to the bifilar arrangement of corresponding partial windings a close magnetic coupling and/or a symmetrical distribution of the phase currents between the partial winding systems is achieved (at least in the upper speed range) also the possibility of limiting the current control to a single partial system [15] shall be analyzed.

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