34th IEEE Industry Applications Society Annual Meeting, Phoenix, Oct. 3–7, 1999

Analysis of a Wide Speed Range Starter/Alternator System Based on a Novel Converter Topology for Series/ Parallel Stator Winding Configuration

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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 of the induction machine is split up into two isolated and quasi-bifilar wound three-phase winding systems. The series operation of the two winding sets permits very high torque at low stator frequencies while keeping the machine terminal 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 given for 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]

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

for future automotive starter/high power generator applications. For fulfilling these requirements basically the following 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 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 $n = 1000\,\mathrm{r/min}$ as lower bound of the constant power operating range and/or speed-proportional decrease of the output power down to $n = 500\,\mathrm{r/min}$. (For $n = 1000\,\mathrm{r/min}$ 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 $M = 250\,\mathrm{Nm}$; as rated power one would have to provide there $P_N \approx 6000\,\mathrm{W}$, 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 $P = 4000\,\mathrm{W}$ in the entire 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 $n = 500\,\mathrm{r/min}$ for allowing $P = 4000\,\mathrm{W}$ at the upper mechanical speed boundary latterly is caused by the speed-proportional reduction of the machine flux in the field-weakening 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 region (limitation to using tapped turns or reconnection 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 change-over interval with zero current and/or zero torque and in the low reliability of the change-over switches.

As this paper shows, a series-parallel-winding reconfiguration, however, also can be achieved by electronic means (cf. **Fig.1**).

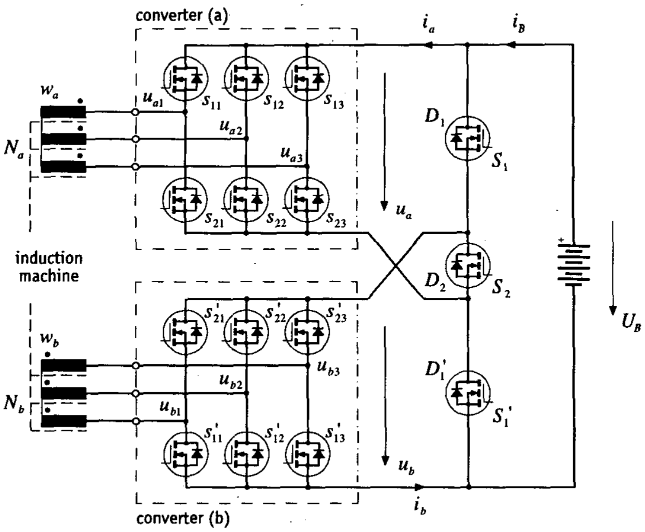

Fig.1: Basic structure of the power circuit of the novel converter topology. For series operation S_2 is in the on-state S_1 and S_1' remain in the off-state; accordingly S_2 is turned off and transistors S_1 and S_1' are turned on for parallel operation. The corresponding phase windings of the partial winding systems w_a and w_b are arranged in a quasi-bifilar manner (cf. section 2.1).

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 respective 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 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 (i.e., power MOSFETs with antiparallel free-wheeling diodes) which results in a relatively low realization effort of the change-over circuit. 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 5 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.

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. There, for the sake of clearness we do not distinguish between current conduction of a transistor or the associated antiparallel free-wheeling diode. Furthermore, the change-over from series to parallel operation and a control concept for the elimination of the common-mode component of the machine voltage are treated (cf. section 2.4).

2.1 Geometrical Placement of the Partial Winding Systems

Induction machines with a six-phase winding arrangement have been widely used in high power current source inverter drives. The two three-phase winding systems are thereby shifted by 30 degrees which allows the elimination of the 6th harmonic in the output torque and higher inverter utilization [7].

However, such a winding arrangement would show significant drawbacks in the case at hand with a voltage source inverter. Although the stator windings are electrically isolated from each other, the two systems are linked closely. The mutual magnetic coupling between the two stator winding systems becomes clear from the space vector equivalent circuit shown in **Fig.2**(b) [7]. The stator leakage flux is split into two parts; part one is solely linked to the individual winding system while for the second part there is a mutual coupling of the winding system. The figure is only complete with the definition of the geometrical placement of the windings in the complex plane as shown in Fig.2(a) for a 30° shift. Fig.2 forms the basis for the space vector equations of each system (cf. Eqs.(1)–(5) in [8]). As a result even identical stator voltages on each system terminals (i.e. parallel connected terminals) do not form the same voltage space vector at the input of the equivalent circuit due to the geometrical shift of the phases (cf. Eq.(6) in [8]). This difference in the voltage vectors is applied across the very low impedance formed by the stator resistance and a fractional part of the stator leakage inductance only (cf. Fig.2(b)) and causes a significant differential current flow between the systems a and b.

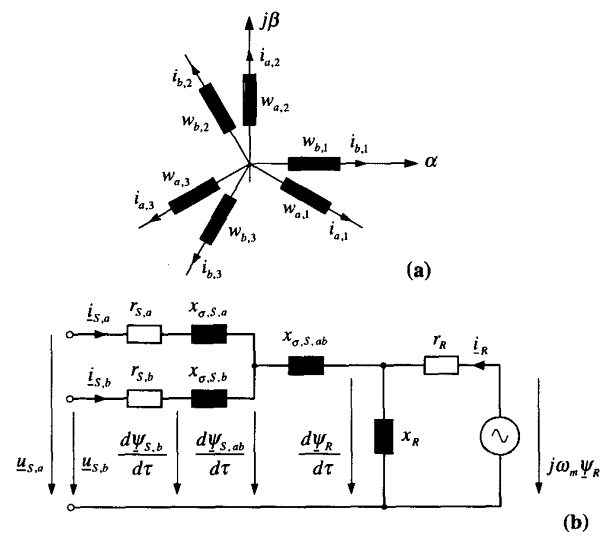

Fig.2: Geometrical placement of the winding systems w_a and w_b defined by the orientation of the phase winding axes in the complex plane **(a)** (shown for 30° angular displacement, as, e.g., given for six-phase induction motors for current source inverter drives [8]) and associated machine space vector equivalent circuit **(b)**. In the case at hand due to placing corresponding phase windings of the systems w_a and w_b in same slots there is no angular displacement of the phase winding axes.

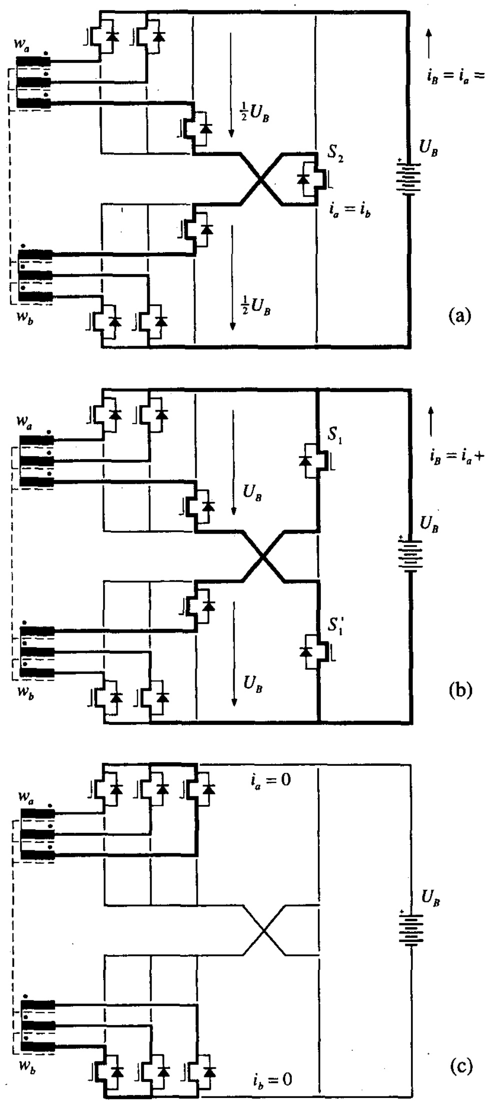

Fig.3: Conduction states of the proposed converter system for **(a)** series and **(b)** parallel operation for an active (voltage forming) switching states; **(c)**: free-wheeling state.

In case two voltage DC link inverters are connected to such a machine with only 7 discrete voltage vectors available a significantly higher current ripple is observed as compared to a conventional machine [9]. Also, if the two winding systems are not shifted and different inverter voltage patterns are applied to the winding systems w_a and w_b one has to expect a high current distortion [10]. Usually an explicite inverter output reactance is used to limit the current distortions in parallel connected voltage DC link inverter arrangements.

The two winding arrangement does not affect the current waveform only if the two winding systems are not shifted and if identical voltage space vectors are applied to the individual systems at any time. However, it is not desirable to have bifilar-wound windings in order to keep some rest of leakage inductance. This small inductance limits the differential currents in case of non-ideal inverter output voltage matching as, e.g., caused by different switching times of the valves of the individual inverters. The two winding systems therefore should be placed in same slots but with an isolation layer in between, i.e., the partial windings should be placed in a quasi-bifilar manner.

2.2 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 voltage space vectors at the converter outputs) 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 to the on-state and $s_i = 0$ to the off-state). The corresponding conduction state of the system is shown in **Fig.3**(a) for an active, i.e., voltage forming switching state. Considering the symmetry being given due to the quasi-bifilar arrangement of the windings of corresponding phases we therefore have for the input voltage of converters 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 (the 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.3(b)). There, 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 the parallel connection is not dependent of the direction of the DC side current flow. Due to the quasi-bifilar arrangement (and/or close magnetic 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 (cf. section 2.1).

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.3(c)).

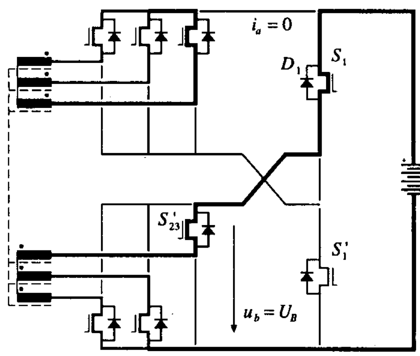

Fig.4: Conduction state for not synchronous switching of corresponding valves of converters a and b and parallel operation.

Considering different switching times of the gating units and of the valves as given in practice also the conduction states occurring for not synchronous switching of the converters are of interest. For the sake of brevity we here would like to only 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.4**). 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 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}. Different terminal voltages of the partial windings result in differential currents of high amplitudes despite to only quasi-bifilar arrangement of corresponding phases of w_a and w_b. Therefore, one has to employ gate control circuits and power transistors with sufficiently low switching delay times.

It is important to note that the voltage at the converter inputs (ideally) due to the topology of the change-over circuit is limited to the battery voltage also for not-synchronous switching of the converters, i.e., no overvoltages do occur.

Remark: Basically the change-over 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 (cf. **Fig.5**(a)). 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 circuit. Thereby, a change-over between series and parallel operation with pulse frequency can be avoided. However, than 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, cf. Fig.5(b)) 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, cf. Fig.5(c)).

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) converter 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 the increase of the turn-off voltage occurring for the valves of 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.

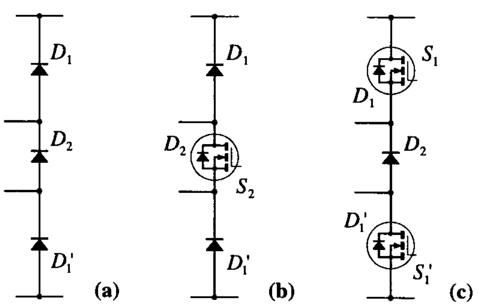

Fig.5: Variations of the series-parallel change-over circuit structure. **(a):** change-over of the winding arrangement dependent on the direction of the main average power flow (series arrangement for motor operation of the induction machine, parallel arrangement for generator operation); **(b):** series winding arrangement possible for motor and generator operation; **(c):** parallel winding arrangement possible for motor and generator operation.

2.3 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 half period the possibility of changing the switching state is given.

2.4 Control of the Converters

As already mentioned in section 2.2 (cf. Eq.(3)), the converters a and b are pulsed in an inverse manner $s_i = s'_i$. Therefore, for winding system w_a and w_b (ideally) common-mode voltages $u_{0,a}$ and $u_{0,b}$ of equal magnitude but opposite sign

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

occur (the inverter output phase voltages $u_{a,i}$ and $u_{b,i}$ 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 cancelation 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) [11], [12]. 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 of the Drive System

For the sake of clarity for the following considerations motor operation of the drive is assumed, also in the upper speed range where the change-over from series to parallel arrangement of the partial winding systems has to be performed. (For applications of the proposed drive as automotive starter/alternator the change-over of the windings arrangement usually would occur in the generator mode).

As **Fig.6** 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_1' 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.6). The machine torque capability can be regained by changing to parallel operation of w_a and w_b by turning S_2 off and turning on S_1 and 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 to continue 4000 W operation up to 6000 r/min.

For deceleration 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 halving the machine flux for constant converter 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 and, therefore, would result 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

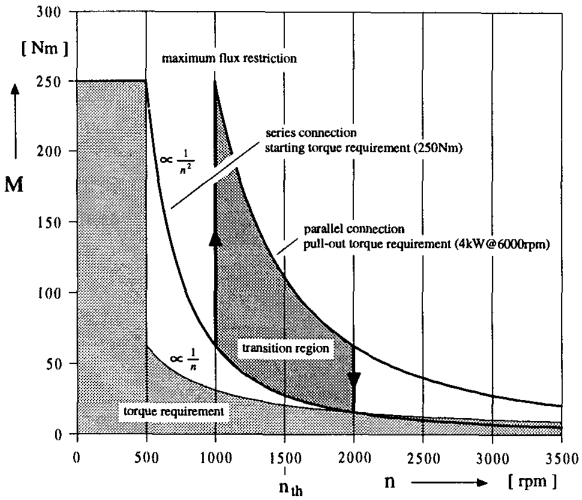

Fig.6: Breakdown torque requirements for series- and parallel operation of the partial stator winding systems according to specifications of starting torque and constant power operating range given in section 1 (pointed out by the dotted area). Transition from series to parallel operation (or vice versa) has to be performed within the transition region.

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 a threshold value $n_{th} \approx 1500$ r/min in the center of the transition region. The change-over 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 ramp rise time is set according to the rotor time constant of the machine. The passing from one desired flux–speed curve to another can be done seamlessly in both directions. In case of low mechanical acceleration the transition can be completed within a small change in speed. In case of very high acceleration and a corresponding very fast passing of the transition region the required amount of flux change is lower as compared to the first case due to the rapid change of the desired flux values. This fact leads to an advantageous characteristic with low action signal demand if the ramp rise time is set properly. The ramp change-over scheme then shows an almost ideal flux adaptation characteristic.

3.1 Control Block Diagram

The control structure equals very much the standard field oriented control of an induction machine [13]-[15] with the exception of the calculation of the stator flux reference value (cf. **Fig.7**).

The stator flux reference value is derived from the actual DC link voltage u_B and the rotor flux angular velocity $\omega_{\underline{\psi}_R}$. Field

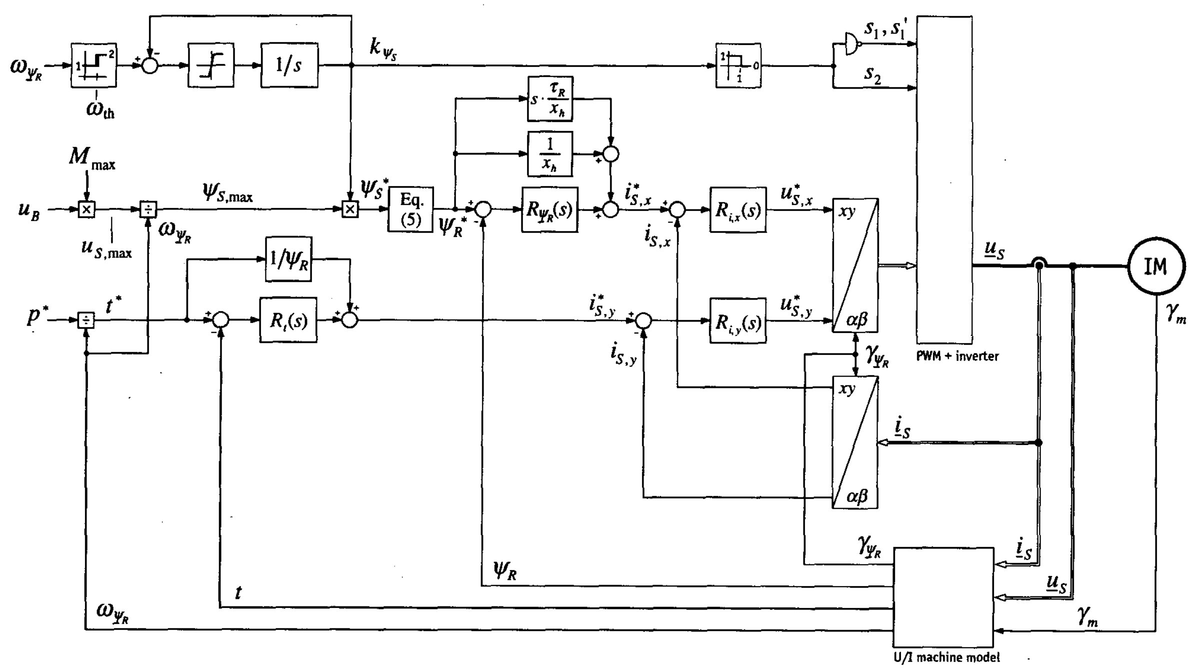

Fig.7: Block diagram of a rotor flux-oriented control (valid only in the field weakening region $n_m > 500\,\mathrm{r/min}$; the stator flux reference value $\psi^*_{S,\max}$ is derived by dividing $u_{S,\max}$ by $\omega_{\underline{\psi}_R}$, for the definition of $M_{\max}$ see [16]) of the proposed induction machine drive; direct axis, which is defined by the orientation of the rotor flux, denoted by index x, quadrature axis denoted by index y. For the sake of clarity the control of the currents in the partial windings systems is not shown in detail. In practice, a current control for each partial winding system has to be implemented in order to guarantee a symmetric split-up of the total stator current for parallel operation. By giving a current reference value for each partial winding system, no adaption of the reference value has to be performed for changing from series to parallel winding arrangement (or vice versa).

weakening starts at 500 r/min. However, the series/parallel selection also changes the stator flux reference as described above (cf. k_{ψ_S} in Fig.7). The rotor flux reference value is derived from the stator flux reference by

$$\psi_R^* = \sqrt{{\psi^*_{S,\max}}^2 - (x_\sigma i_{S,y,\max})^2} - x_\sigma i_{S,x}\ . \qquad (5)$$

The rotor flux controller utilizes a static and dynamic feedforward path according to the dependency of the rotor flux on the magnetizing stator current component $i_{S,x}$,

$$\frac{\mathrm{d}\psi_R}{\mathrm{d}\tau} + \frac{r_R}{x_R}\psi_R = r_R\frac{l_h}{x_R}i_{S,x} \qquad \tau_R = \frac{x_R}{r_R}\ . \qquad (6)$$

The power set value p^* determines the required machine torque t^*. The torque controller also uses a feedforward path and outputs the reference value $i^*_{S,y}$ of the torque producing stator current component. Both stator current components are controlled by a synchronous vector current controller which defines the converter output voltage to be applied in the average over a pulse period. For deriving the control signals for the power transistors of the converters, e.g., space vector modulation could be used advantageously.

A machine model estimates the rotor flux vector $\underline{\psi}_R$ and calculates the transformation angle $\gamma_{\underline{\psi}_R}$ being required for the rotor flux orientation of the reference frame. In the lower speed range a current model (requiring a rotor position encoder) is employed in order to guarantee the high starting torque. Operation at higher speed is based on a voltage model showing lower parameter sensitivity.

The instant for a change of the machine winding configuration is derived from the actual rotor flux angular speed $\omega_{\underline{\psi}_R}$ (which differs to the mechanical speed only by the slip frequency) determining the back-emf of the machine. Whenever a certain threshold ω_{th} is reached an integration process starts that ramps the stator flux reference value up to two times its regular value (as given for series operation) with respect to speed (cf. multiplication of $\psi^*_{S,\max}$ by k_{ψ_S}). Please note, that this is always done in the field weakening region and, therefore, the rated machine flux will never be exceeded. On the other hand, the change back from parallel to series operation of the winding systems at lower speed only occurs when the flux has been guided back to its regular value. For the control structure given in Fig.7 a high control accuracy and/or high dynamic quality of the flux control is assumed, i.e., the change-over is directly determined by k_{ψ_S} and/or the flux reference value.

Remark: A similar machine control characteristic with the very same series to parallel change-over strategy could be realized using a stator flux oriented scheme.

4 System Simulation

4.1 Simulation Parameters

The system control has been analyzed by digital simulations based on parameters gained from the dimensioning of a 2-kW prototype of in induction machine which has been designed for high nominal to breakdown torque ratio and also is planned to be employed for an experimental verification of the system. The drive was assumed to be operated at a DC link/battery voltage $U_B = 220V$ being available at the electrical machine lab of the Technical University of Vienna. The rated (index N) technical parameters of the machine are listed in the following:

$$\begin{array}{lcllcl} U_N & = & 150\,\mathrm{V}/50\,\mathrm{Hz} & n_K & = & 770\,\mathrm{r/min} \\ p & = & 3 & R_1 & = & 0.20\,\Omega \\ I_N & = & 17.1\,\mathrm{A} & R_2' & = & 0.14\,\Omega \\ n_N & = & 985\,\mathrm{r/min} & X_{1,\sigma} & = & 0.37\,\Omega \\ M_N & = & 19.4\,\mathrm{Nm} & X_{2,\sigma}' & = & 0.41\,\Omega \\ M_K & = & 104\,\mathrm{Nm} & X_{1,h} & = & 5.81\,\Omega. \end{array}$$

The rise time of the ramp function adapting the flux reference value for each change-over of the winding arrangement has been set to 100ms in close correspondence with the rotor time constant τ_R of the machine.

4.2 Normalization

For easing the comparison and for simplifying a transfer of the obtained results to different power levels it is advantageous to use a normalized system description. All voltages are referred to the amplitude of the rated phase voltage. Currents are referred to the rated amplitude of the phase current. The reference value of the flux is determined by the reference voltage and the synchronous angular speed of the flux space vector at rated frequency. ω_m denotes the normalized rotor (mechanical) angular speed. Note, that the time is normalized using $\tau = \omega_N t_{\mathrm{real}}$, where ω_N denotes the rated stator frequency and t_{real} denotes the time in seconds. Torque values are referred to a reference torque value defined by rated apparent power divided by the rated synchronous mechanical angular speed.

4.3 Assumptions

The simulations are based on the following assumptions:

- constant DC link voltage $u_B = U_B = 220\,\mathrm{V}$, i.e., neglection of the inner impedance of the battery;
- generator operation and/or impressed mechanical speed (the speed is assumed to be determined by the internal combustion engine)
- torque controller $R_t(s)$ not implemented (a torque controller is useless in a simulation where no parameter uncertainty exists, therefore, the torque control is only by the feed-forward path)
- control of the phase currents by a dead-beat current controller (the current controller defines a local average of the stator voltage vector which ideally would result in zero current control error at the end of the following pulse interval)
- space vector modulation.

The battery voltage is applied to the machine phase terminals as shown in **Fig.8**(a) for series connection of the partial winding systems. The winding configuration which is shown in Fig.8(c) is used for the simulation of the parallel operation. The winding connection there remains as in Fig.8(a) but the machine terminal voltage is artificially doubled. As a consequence the winding system voltage is equal as actually given for parallel connection and the winding system current is also represented correctly. As the machine normalized values are equal for series and parallel operation (for normalization to the respective rated values) the integration of the flux and current values can be continued seamlessly, which significantly simplifies the modeling of the machine.

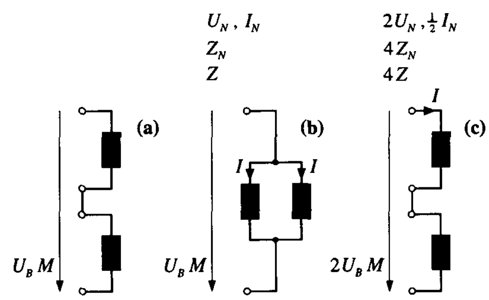

Fig.8: Simulation of the parallel operation **(b)** of the partial winding systems w_a and w_b based on the consideration of an equivalent series connection **(c)** being supplied by a converter having DC link voltage level being higher by a factor of 2 as compared to regular series operation **(a)**.

4.4 Simulation Results

Figure 9 shows an acceleration/deceleration process of the drive for generator operation, i.e., for feeding a constant amount of power into the DC link (the stator winding resistance is neglected and/or the machine output power is considered to be equal to the air-gap power). The angular acceleration/deceleration is thereby set to a constant value $\pm 5000\,(\mathrm{r/min})/\mathrm{s}$ as it might typically occur during acceleration/deceleration of an internal combustion engine for declutched operation. Shortly after t_0 the drive enters the field weakening region at t_1. Now the flux has to be reduced quickly, which is achieved by a negative magnetizing current component $i_{S,x}$ The time behavior of rotor and stator flux magnitude is shown in Fig.9 by traces ψ_S and ψ_R, respectively.

At time t_2 the winding arrangement is changed to parallel. Accordingly, the available voltage per machine winding system doubles and the stator flux reference value can be ramped up, which can be observed in time interval $t_2 \ldots t_3$. Please note

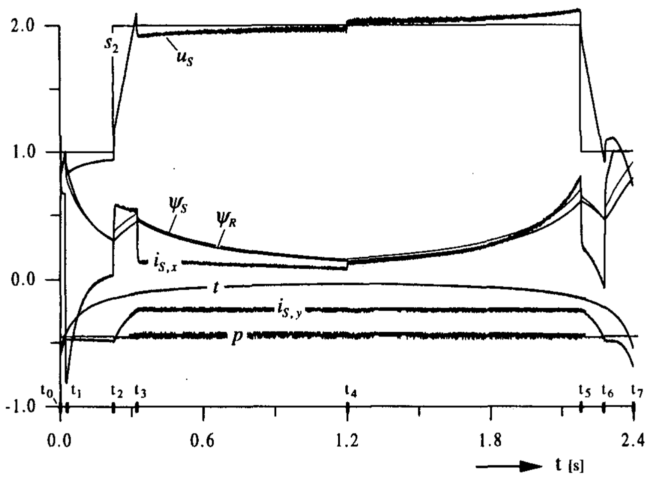

Fig.9: Simulation results for operating the proposed drive system with impressed mechanical speed ω_m, i.e., constant acceleration in time interval (t_0, t_4) and constant deceleration in time interval (t_4, t_7).

that the change-over does not cause any air-gap torque or air-gap power discontinuity. The magnetizing current $i_{S,x}$ stays well below the limits due to the flux reference ramp. The required fundamental stator voltage can bee seen from trace u_S; its value change from 1.0 to 2.0 with slight overshoots due to the magnetizing current demand. The stator current ripple in the system shows higher amplitudes for parallel than for series operation due to the higher inverter voltage applied to the partial winding systems for parallel operation.

The machine accelerates up to a maximum speed of 6300 r/min and is forced into deceleration at t_4. Now the machine flux has to be built up again, which can be observed in the time behavior of the magnetizing stator current component $i_{S,x}$. Again torque t and power p stay unaffected. The machine decelerates till t_5 where the preparation for changing over form parallel to series arrangement of the partial winding systems starts. The machine flux is reduced to its original lower value, which is reached in t_6. Now the change-over of the winding arrangement can be performed and the decrease of the mechanical speed continues until t_7 with series connected partial winding systems.

According to the digital simulation (including the inverter current controller and the induction machine, cf. Fig.9) the change in the winding configuration can be done without any torque disturbance due to the principle of field oriented machine control. As a consequence of the continuous flux control during the transition the system is not stressed by an abrupt step in the flux level. This feature is, e.g., in contrast to electronically changing the number of poles [2] being characterized by an abrupt flux level change and also with the 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 series/parallel change of the winding arrangement can be handled safely by the inner current controller loop. Torque and power control stay almost unaffected. In order to keep the distortion as small as possible the transition should be synchronized to the current controller cycle time. The current controller uses then already a new switching sequence for the next sampling interval and the distortion ideally remains zero.

Remark: Aiming for low losses of the machine one would have to prefer parallel operation to series operation within the transition region. 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.

5 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 (related to the losses of the partial converters, and the current stress on the supplying battery as occurring for series and parallel operation are of interest.

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).

5.1 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 voltage 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 partial converters is assumed.

5.2 Operation at Low Speed and High Torque

In order to make possible a direct comparison of the proposed series connection of the partial windings to 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 considered). If now for realizing each valve of partial converter a and partial converter b a single power MOSFET (with on-resistance $R_{DS(\mathrm{on})}$) is employed, both systems show an 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, 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 (cf. **Fig.10**). For the ratio of the modulation depths we, therefore, have according to $u_a = u_b = \frac{1}{2}U_B$,

$$M_\mathrm{s} = 2M_\mathrm{p} \qquad M_\mathrm{p} = \frac{\hat{U}_U}{\frac{1}{2}U_B} , \tag{7}$$

($\hat{U}_U$ denotes the amplitude of the fundamental of the voltage applied to the machine).

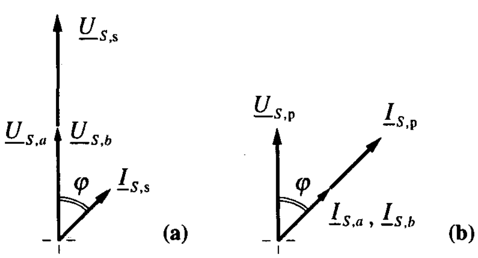

Fig.10: Phasor diagram of stator phase voltage and stator current fundamental for series **(a)** and parallel **(b)** operation of the partial stator winding systems. For parallel operation the stator voltage phasor shows half the absolute value of the voltage phasor given for series operation. Accordingly the current phasors differ in absolute value by a factor of 2 for equal output power.

5.2.1 Converter Conduction and Switching Losses

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

$$P_\mathrm{c} = P_\mathrm{c,s} = P_\mathrm{c,p} = \frac{3}{2} I^2_{N,\mathrm{rms}} R_{DS(\mathrm{on})} \tag{8}$$

in both cases ($I_{N,\mathrm{rms}}$ 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 half. We therefore have

$$P_\mathrm{p,s} = \frac{1}{2} P_\mathrm{p,p} . \tag{9}$$

5.2.2 Conduction Losses of S_2

According to [17] the current stress on S_2 can be calculated as

$$I^2_{S_2,\mathrm{rms}} = I^2_{N,\mathrm{rms}} [\frac{\sqrt{3}M_\mathrm{s}}{2\pi}(\frac{1}{4} + \cos^2\varphi)] , \tag{10}$$

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 $M_\mathrm{s} \approx 0.9$ and $\cos\varphi \approx 0.7$ are assumed, there follows $I^2_{S_2,\mathrm{rms}} = 0.09 I^2_{N,\mathrm{rms}}$. In case S_2 also is realized by a valve with on-resistance $R_{DS(\mathrm{on})}$, therefore, the system conduction losses are increased only by about $0.06P_\mathrm{c}$.

5.2.3 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 half

$$\hat{i}_{B,\mathrm{s}} = \frac{1}{2}\hat{i}_{B,\mathrm{p}} \tag{11}$$

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 and/or 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.11**).

For the rms value of the battery current we have for series operation [17]

$$I_{B,\mathrm{rms,s}} = I_{S_2,\mathrm{rms}} = I_{N,\mathrm{rms}}\sqrt{\frac{\sqrt{3}}{2\pi}M_\mathrm{s}(\frac{1}{4} + \cos^2\varphi)} , \tag{12}$$

for parallel operation there results a current stress [17]

$$I_{B,\mathrm{rms,p}} = I_{N,\mathrm{rms}}\sqrt{\frac{2\sqrt{3}}{\pi}M_\mathrm{p}(\frac{1}{4} + \cos^2\varphi)} , \tag{13}$$

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

5.2.4 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

$$I_{C,\mathrm{rms,s}} = I_{N,\mathrm{rms}}\sqrt{\frac{M_s}{2}[\frac{\sqrt{3}}{4\pi} + \cos^2\varphi(\frac{\sqrt{3}}{\pi} - \frac{9}{16}M_s)]}\ , \quad (14)$$

or

$$I_{C,\mathrm{rms,p}} = I_{N,\mathrm{rms}}\sqrt{2M_p[\frac{\sqrt{3}}{4\pi} + \cos^2\varphi(\frac{\sqrt{3}}{\pi} - \frac{9}{16}M_p)]}\ . \quad (15)$$

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

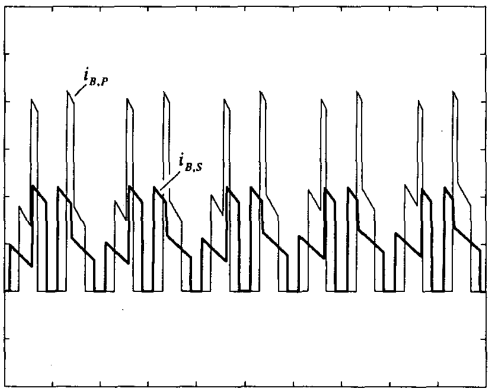

Fig.11: Time behavior of the battery current i_B for $M = 0.9$ and $M = 0.45$ and/or for series and single converter (parallel) operation of the partial winding systems ($\cos\varphi = 0.45$).

5.3 Discussion

The results of the foregoing calculations are compiled in **Tab.1**. As a comparison of the characteristic values resulting for series and parallel operation clearly shows the increase of the converter conduction losses due to S_2 is by far compensated by lower switching losses and lower losses in the DC link energy storage as compared to parallel operation. Furthermore, the maximum 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 (cf. section 1, item 2.) is only paid for by a higher gate drive and assembly effort concerning the converter circuit but not by higher system losses or a higher stress on the power components.

6 Conclusions

The paper presents a novel inverter concept for the realization of an induction machine drive with a very wide constant power range. There, the stator winding of the induction machine is split up into two isolated and quasi-bifilar wound three-phase winding 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 sweries 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 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 50% of the maximum possible value. If this condition would be violated the machine back-emf would exceed the maximum inverter output voltage and current control would be lost. Accordingly, a high current peak and a torque disturbance would occur. 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 the 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)

Mode of operation	modulation index	converter conduction losses	converter switching losses	conduction losses in S_2	battery current stress	battery losses	DC link capacitor losses
series	0.9	P_c	P_p	$0.06P_c$	$\hat{i}_{B,s}$	$0.13P_c$	$0.05P_c$
parallel	0.45	P_c	$2P_p$	—	$2\hat{i}_{B,s}$	$0.26P_c$	$0.17P_c$

Tab.1: Comparison of the loss contributions of the power components and of the current stresses on the DC link energy storage devices as occurring for series or parallel (single converter) operation. Parameters: $cos\varphi = 0.7$; turn-on resistance of S_2 and inner resistance of the battery and equivalent series resistance of the DC link capacitor are assumed to be equal to the turn-on resistance $R_{DS(\mathrm{on})}$ of a power transistor of a partial converter; for parallel/single converter operation a valve is assumed to be formed by parallel connection of two power transistors having each a turn-on resistance of $R_{DS(\mathrm{on})}$ in order to ensure about equal realization effort of the converter power circuit in both cases.

+ 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 [15] etc.

Drawbacks:

- as compared to single converter operation and conventional winding techniques higher rated converter power

- 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 occurring in the change-over switches S_1 and S_1'.

Further research will be on an experimental analysis of the proposed system. Besides verifying the theoretical considerations given in this paper there the potential of instabilities of the individual converter current controls operating in parallel for parallel winding configuration [10] shall be analyzed. Considering the fact that due to the quasi-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 shall be analyzed.

Furthermore, the performance of the rotor flux oriented control shall be investigated in detail; in particular the susceptibility of the system to variations of the DC link voltage (as caused by the battery inner impedance in connection with the load current) and parameter uncertainties (e.g., of the rotor time constant) shall be considered. Also, a stator flux oriented control shall be implemented for the control of the wide constant power range. This control may show a higher insensitivity to DC link voltage variations.

ACKNOWLEDGMENT

The authors are very much indebted to the Jubiläumsfonds der OESTERREICHISCHEN NATIONALBANK which generously supports the work of the Department of Electrical Drives and Machines of the Technical University of Vienna.

References

[1] **Miller, J.M., McCleer, P.J., Lang, J.H., Gale, A.R., and Leonardi, F.**: *Starter-Alternator for Hybrid Electric Vehicle: Comparison of Induction and Variable Reluctance Machines and Drives.* Rec. of the 33rd IEEE Ind. Applications Soc. Annual Meeting, Oct. 12-15, St. Louis, Vol. 1, pp. 513-523 (1998).

[2] **Osama, M., and Lipo, T.A.**: *Modeling and Analysis of a Wide-Speed-Range Induction Motor Drive Based on Electronic Pole Changing.* IEEE Transactions on Industry Applications, Vol. 35, No. 5, pp. 1177-1184 (1997).

[3] **Kume, T., Iwakane, T., Sawa, T., Yoshida, T., and Nagai, I.**: *A Wide Constant Power Range Vector Controlled AC Motor Drive Using Winding Changeover Technique.* IEEE Trans. on Ind. Appl., Vol. 27, No. 5, p p. 934–939 (1991).

[4] **Naidu, M., Boules, N., and Henry, R.**: *A High-Effciency High Power Generation System for Automobiles.* Record of the 30st IEEE Industry Applications Society Annual Meeting, Oct. 8-12, Orlando (FL), U.S.A., Vol. 1, pp. 709-716 (1995).

[5] **Boglietti, A., Lazzari, M., Pastoreilli , W., and Vigli-etti, W.**: *Induction Motor Winding Commutation with Direct Field Oriented Control for Wide Constant Power Range Operation.* Record of the 31st IEEE Ind. Appl. Soc. Annual Meeting, Oct. 6-10, San Diego (CA), Vol. 1, pp. 551-557 (1996).

[6] **Osama, M., and Lipo, T.A.**: *A New Inverter Control Scheme for Induction Motor Drives Requiring Wide Speed Range.* IEEE Trans. on Ind. Appl., Vol. 32, No. 4, pp. 938–944 (1996).

[7] **Lipo, T.A.**: *A d-q Model for Six Phase Induction Machines.* Proceedings of the International Conference on Electrical Machines, Athens, Greece, Sept. 15–17, pp. 860–867 (1980).

[8] **Wieser, R.S., and Lechner, A.**: *Accurate On-Line Model for a 6-Phase Current Source Inverter Fed Induction Machine Without Machine Current Measurements.* Proceedings of the International Conference on Electrical Machines, Istanbul, Turkey, Sept. 2–4 (1998).

[9] **Stemmler, H., and Deplazes, R.**: *Multi-Star Induction Motors Fed by Voltage Source Inverters.* Proceedings of the Power Conversion Conference, Aug. 3–6, Nagaoka, Japan, Vol. I, pp. 119–126 (1997).

[10] **Thunes, J., Kerkman, R.J., Schlegel, D., and Rowan, T.M.**: *Current Regulator Instabilities on Parallel Voltage Source Converters.* IEEE Transactions on Industry Applications, Vol. 35, No. 1, pp. 70–77 (1999).

[11] **Oriti, G., Julian, L., and Lipop, T.A.**: *An Inverter/Motor Drive with Common Mode Voltage Elimination.* Record of the 32nd IEEE Industry Applications Society Annual Meeting, New Orleans, Oct. 5–9, Vol. 1, pp. 587–592 (1997).

[12] **von Jouanne, A., and Zhang, H.**: *A Dual Bridge Inverter Approach to Eliminating Common Mode Voltages and Bearing and Leakage Currents.* IEEE Transactions on Power Electronics, Vol. 14, No. 1, pp. 43–48 (1999).

[13] **Moreira, J.C., and Lipo, T.A.**: *A New Method for Rotor Time Constant Tuning in Indirect Field Oriented Control.* Proceedings of the IEEE Power Electronics Specialists Conference, San Antonio, June 10–15 (1990).

[14] **Seibel, B.J., Rowan, T.M., and Kerkman, R.J.**: *Field Oriented Control of an Induction Machine with DC Link and Load Disturbance Rejection.* Proc. of the 11th IEEE Applied Power Electronics Conf., San Jose, March 3–7, pp. 387–393 (1996).

[15] **Bunte, A., Grotstollen, H., and Kafka, P.**: *Field Weakening of Induction Motors in a Very Wide Region with Regard to Parameter Uncertainties.* Record of the 27th IEEE Power Electonics Specialists Conference, Baveno, Italy, June 23–27, Vol.1, pp. 944–959 (1996).

[16] **Wieser, R.S.**: *Optimal Rotor Flux Regulation for Fast Accelerating Induction Machines in the Field Weakening Region.* Record of 32nd IEEE Industry Applications Society Annual Meeting, New Orleans, Oct. 5–9, Vol. 1, pp. 401–409 (1997)

[17] **Kolar, J.W., Ertl. H., and Zach, F.C.**: *Calculation of the Passive and Active Component Stress of Three-Phase PWM Converter Systems with High Pulse Rate.* Proc. of the 3rd European Conf. on Power Electr. and Applications, Aachen, Germany, Oct. 9–12, Vol. III, pp. 1303–1311 (1989).

[18] **Caricchi, F., Credcimbini, F., Mezzetti, F., and Santini, E.**: *Multi-Stage Axial-Flux PM Machine for Wheel Direct Drive.* Record of the 30st IEEE Ind. Appl. Soc. Annual Meeting, Oct. 8-12, Orlando, U.S.A., Vol. 1, pp.679-684 (1995).