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Conference Paper**Author(s):**

Bringezu, Thilo; Biela, Jürgen

Publication date:

2020

Permanent link:

<https://doi.org/10.3929/ethz-b-000495880>

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Originally published in:

<https://doi.org/10.23919/epe20ecceeurope43536.2020.9215873>

Comparison of optimized motor-inverter systems using a Stacked Polyphase Bridge Converter combined with a 3-, 6-, 9-, or 12-phase PMSM

Thilo Bringezu and Jürgen Biela

Laboratory for high power electronic systems (HPE), ETH Zürich

E-Mail: bringezu@hpe.ee.ethz.ch

Keywords

«Modular converter», «Multiphase drive», «Permanent magnet motor», «System integration»

Abstract

Stacked Polyphase Bridge Converters (SPB-C) are well suited for integrated motor drives, i.e. for systems that combine an electric motor with its supplying power electronic converter in one compact unit. A SPB-C requires a motor with multiple 3-phase winding systems. In the literature, the SPB-C is typically combined with motors whose winding systems are phase-aligned. This type of winding is magnetically equivalent to a 3-phase winding (with parallel branches). In this paper, the SPB-C with either 2, 3, or 4 modules is combined with a motor that has non-phase-aligned winding systems what results in a 6-, a 9-, or a 12-phase system, respectively. These multiphase systems are compared to 3-phase systems with the same number of modules with regard to efficiency, power density, and the possible degree of modularity and the benefits of the multiphase systems are evaluated.

1 Introduction

One solution to face the demand for more compact and efficient drive systems is to integrate the power electronic converter into the motor [1]. An example for such an integrated motor drive is given in Fig. 1 which consists of a tooth-coil wound PMSM for high-torque applications and a modular converter cooled by an end plate with cooling fins. Modular converter topologies, such as Stacked Polyphase Bridge Converters (SPB-C), are the preferred choice for this integration concept as the converter modularity ensures a very compact design. The SPB-C is introduced in [2] as a modular converter topology consisting of series connected three-phase converters for hybrid electric vehicles. The topology of the SPB-C is shown in Fig. 2a for 2, 3, and 4 three-phase modules which are called SPB-modules in the following. Each SPB-module has a dc link capacitor and is connected to a 3-phase winding system of a motor in multi-star point connection. The main advantages of the SPB-C are: Low voltage semiconductor devices with low $R_{ds,on}$ can be used due to the series connection, fault-tolerance, and modularity ensuring a compact integrated system. The latter advantage is presented in different publications that analyze converter topologies suitable for an Integrated Modular Motor Drive (IMMD) [3], [4]. The IMMD-concept is introduced in [5] as a motor with segmented stator and axially integrated converter units leading to a compact modular system.

Figure 1: Basic concept of an integrated motor drive.

Figure 2: a) SPB-C with 2, 3, and 4 SPB-modules connected to motor windings [2]. b) For each SPB-C, two configurations of the magnetic winding axes are shown with m phases and a_p phase-aligned winding systems.

Publications [3], [6] consider the SPB-C in combination with modular three-phase motors, i.e. motors having as many phase-aligned three-phase winding systems as SPB-modules. However, the SPB-C is also suitable for motors with non-phase-aligned three-phase winding systems. In [4], the possibility of using either phase-aligned or non-phase-aligned winding systems is mentioned but the effect on the system performance is not analyzed. Therefore, this analysis is presented in this paper. Fig. 2b shows the magnetic winding axes for different possible alignments. The motor-converter configurations ①, ③, and ⑤ have phase-aligned three-phase winding systems resulting in 3-phase motors, whereas configurations ②, ④, and ⑥ have non-aligned three-phase winding systems, resulting in a 6-, a 9-, and a 12-phase motor, respectively.

The motivation for considering higher phase numbers is that higher phase numbers result in higher winding factors and consequently in systems with higher power density and/or efficiency as will be shown based on the design procedure described in section 2. The motor related models and design steps for the design procedure are presented in section 3. Thereafter, the converter related models and design steps are presented in section 4. Finally, the results of the design procedure, applied to all system configurations shown in Fig. 2, are presented and discussed for specifications given in Tab. IV-VI in section 5.

2 Design procedure

The main goal of this paper is to determine and to compare the system performance of the motor-inverter configurations ① to ⑥ (c.f. Fig. 2) in terms of the performance indicators efficiency, power density, and the possible degree of motor-converter modularity that is defined in section 3.2. A focus is put on the comparison of configurations ① & ②, ③ & ④, and of ⑤ & ⑥ because these pairs have the same amount of semiconductor devices at the same ratings and thus comparable costs.

The procedure used for an optimal system design in terms of the aforementioned performance indicators is shown in Fig. 3. First, the system specifications are defined which consist of the rated torque, the rated speed, the dc-link voltage, the selected semiconductor devices, and the considered system configuration ①/.../⑥ shown in Fig. 2. For these specifications, a motor is designed in the steps S1 and S2. In **S1**, the number of stator slots and the number of rotor pole pairs are selected based on the criteria described in 3.2. This selection already determines the possible motor-converter modularity as will be explained later. In **S2**, the motor dimensions are optimized and electrical motor parameters are calculated using the optimization routine presented in 3.3. In the next steps S3-S8, the switching frequency is optimized. For each switching frequency selected in **S3**, the motor/converter currents are calculated in **S4** using the equations for sinus-PWM given in [7] and the motor model equations given in 3.1, where the electrical motor parameters resulting from S2 are inserted. In **S5**, the dc link is dimensioned using the method described in 4.1. In **S6** and **S7**, the system losses and the system volume are calculated, respectively. The used volume models and the loss models are described in 3.4 for the motor and in 4.2 for the converter. The sum of all volumes and the sum of all losses are finally used to calculate the system efficiency and power density in **S8**.

Figure 3: Design procedure for an optimized motor-converter system using the SPB-C and a tooth-coil wound PMSM for given system specifications. Optimization goals: high efficiency, high power density, and high degree of motor-converter modularity. Yellow: Motor modelling and design steps. Green: Converter modelling and design steps.

Table I: Used motor variables.

$A_{c,skin}(f)$	Effective conductor cross sectional area	m	Number of phases (distinct winding axes)
A_δ	Air gap area ($2\pi r_g l_m$)	μ_0	Vacuum permeability
α	Angular position on the stator	N	Number of stator slots
$\alpha_{N[-\pi,\pi]}$	El. angle between adjacent slots in $[-\pi, \pi]$	$N_i(\alpha)$	Winding function of the i -th phase
α_{sys}	Angle between winding axes (cf. Fig. 2)	N_l	Number of tooth-coils
α_z	Angle between phasors in the star of slots	n_l	Number of winding layers (1 or 2)
a_p	Number of phase-aligned winding axes	n_n	Rated speed
$\hat{B}_{\delta,1}$	Fundamental air gap flux density	n_t	Number of turns per tooth-coil
b_s	Slot opening	ω	Electrical angular frequency
D	Outer stator diameter	$p_{Cu,loss}$	Copper loss density (related to motor surface)
δ	Air gap length	PF	Power factor
h_M	Length of rotor magnets in radial direction	p	Number of pole pairs
$\vec{i}_{dqk}$	Motor current in the k -th dq-subspace	$\vec{\psi}_{dqk}$	Motor flux linkage in the k -th dq-subspace
J	Current density in the stator winding	$\vec{\psi}_{PMk}$	Magnet flux linkage in the k -th dq-subspace
k_{fill}	Copper fill factor	$R(f)$	AC phase resistance
k_p	Pole embrace of rotor magnets	r_g	Air gap radius
k_w	Winding factor	ρ_{Cu}	Copper conductivity
L_{dqk}	Motor inductance in the k -th dq-subspace	T	Electromagnetic torque
L_{ij}	(i, j) -th motor inductance matrix entry	T_n	Rated torque
l_m	Motor length	t	Winding periodicity number
$\bar{l}_t$	Mean length of one winding turn	$\vec{v}_{dqk}$	Motor voltage in the k -th dq-subspace
M	Modulation index		

3 Motor modelling and design

In this section, the motor model and the motor design steps of the system design procedure in Fig. 3 are explained. The used motor variables are summarized in Tab. I.

3.1 Electromechanical model for multiphase non-salient PMSMs

An electromechanical motor model comprising a voltage, a flux, and a torque equation in dq-coordinates is given in [8] in (5.9) for electrically excited salient multiphase synchronous machines. This model is adapted to non-salient PMSMs by replacing the product of the excitation current and of the main inductance with the magnet flux linkage $\vec{\psi}_{PMk}$ and by setting the inductance in the d-axis equal to the inductance in the q-axis. These adaptations result in the following set of equations

$$\vec{v}_{dqk} = R \vec{i}_{dqk} + jk\omega \vec{\psi}_{dqk} + \frac{d\vec{\psi}_{dqk}}{dt} \quad (1)$$

$$\vec{\psi}_{dqk} = L_{dqk} \vec{i}_{dqk} + \vec{\psi}_{PMk} \quad (2)$$

$$T = \frac{m}{2} p \sum_{k=1,3,\dots}^{k_{\max}} k \psi_{PMk} i_{qk} \stackrel{!}{=} \frac{m}{2} p \psi_{PM1} i_{q1} \quad (3)$$

for $k = 1, 3, \dots, k_{\max}$ with $k_{\max} = m - 2$, if m is odd and $k_{\max} = m - 1$, if m is even. In multiphase machines ($m > 3$), different harmonics of phase voltages, currents and flux linkages are mapped into $k_{\max} > 1$ dq-subspaces [8], in contrast to 3-phase machines where $k_{\max} = 1$. This paper does not consider harmonic current injection ($i_{qk} \approx 0$ for $k > 1$) so that only the first summand in (3) remains.

3.2 Selection of a slot-pole combination

In this section, the criteria used for selecting a slot-pole combination, i.e. a combination of the number of stator slots N and of the number of rotor pole pairs p are described (step S1 in Fig. 3). There, tooth-coil windings (double-layer and single-layer) are considered because of their shorter end windings compared to distributed windings, which is important for high torque motors whose diameter is large compared to its length. The considered criteria are summarized in Tab. II and explained in the following. Criteria C1 to C5 are commonly used in motor design whereas criterion C6 is particularly added to determine how suitable a slot-pole combination is for an integrated modular motor-converter system.

C1: An m -phase, n_l -layer tooth-coil winding for a motor with N stator slots, p rotor pole pairs, and a_p phase-aligned winding systems is feasible if the two conditions C1a and C1b are fulfilled. **C1a** ensures that the number of stator slots can be equally shared among all phase windings and that α_{sys} matches an integer multiple of the electrical distance of slots along the stator circumference. **C1b** ensures that a_p phase-aligned winding systems are possible. The analytical expressions for C1a and C1b in (4) are valid for single and for double layer tooth-coil windings. The expressions are derived based on [9] and [10] by defining the number of layers n_l (1 for single layer,

Figure 4: a) Star of slots for an example winding with $m = 3$, $n_l = 2$, $N = 27$ and $p = 12$. b) Derivation of the winding factor from the star of slots.

2 for double layer windings) and the number of tooth-coils $N_l := n_l N / 2$.

$$\mathbf{C1a}: \frac{\alpha_{\text{sys}} N_l}{2\pi t} \in \mathbb{N} \quad \wedge \quad \mathbf{C1b}: \frac{a_{p,\text{max}}}{a_p} \in \mathbb{N} \quad \text{with } a_{p,\text{max}} = \begin{cases} 2t, & \text{if } \frac{N_l}{mt} \text{ is even} \\ t, & \text{if } \frac{N_l}{mt} \text{ is odd} \end{cases} \quad (4)$$

The winding periodicity number t in (4) is given by the greatest common divisor (gcd) of N_l and p . This number indicates how often a so-called base winding is repeated along the stator circumference.

C2: The fundamental winding factor $k_{w,1}$, further called winding factor k_w , can be defined as $\psi_{\text{PM1}} / w \phi_{\text{PM1}}$, where w is the number of turns in series per phase of the stator winding and ϕ_{PM1} is the fundamental air gap flux of one rotor magnet [11]. Therefore, the product $k_w \cdot w$ describes the effective number of turns linked to the complete rotor flux ϕ_{PM1} . In the considered design approach where torque, speed and available motor voltage are imposed, a higher winding factor can be used to reduce the number of turns or to reduce the flux by decreasing the motor size. These changes have an impact on efficiency and/or power density as will be shown later. Here, the winding factor is determined using the method of the star of slots [9], [10], illustrated in Fig. 4 for one example configuration. The "star" in Fig. 4a consists of so-called slot phasors indicating the voltages induced by the rotating excitation field in the coil sides of one layer of one base winding. The phasors are placed at an angular distance of $\alpha_z = 2\pi t / N_l$ resulting in a number of $n_e := N_l / t$ phasors. Each of the slot phasors, denoted by $\vec{e}_i$, is assigned to one phase with a positive or negative sign s_i according to the zones labeled +R, -T, etc. that are defined by the considered winding axes configuration (cf. Fig. 2b). Then, the winding factor is calculated as the ratio of the total induced phase voltage to the induced phase voltage of an ideal reference winding whose slot phasors lie on the same axis, as shown in Fig. 4b. There, the coil pitch and the winding distribution are inherently taken into account [9].

C3: Eddy currents in the rotor magnets and in the rotor iron due to parasitic stator magnetomotive force harmonics should be negligible. In this paper, a relative indicator $p_{r,\text{loss}}$ derived in [3], [12] is used to take this loss component into account. Slot-pole combinations with $p_{r,\text{loss}} > 100$ are discarded in the analysis.

C4 & C5: As a further criterion, $\text{gcd}(N_l, 2p) > 1$ is applied in order to avoid unbalanced magnetic pull that can lead to motor vibration and noise [13]. Furthermore, a high value of the lowest common multiple (lcm) of N and $2p$ indicates low cogging torque [13]. Here, $\text{lcm}(N, 2p) > 50$ is chosen.

C6: To evaluate how suitable a winding is for an integrated modular motor-converter system, three degrees of modularity for motor-converter systems are defined (cf. Fig. 5). The three degrees A, B and C define to what extent the stator and the converter can be modularized into motor-converter modules. In Fig. 5, coils of the same color are connected to the same SPB-module. Modularity degree A means that only converter modules but no combined motor-converter modules are possible because the coils connected to one SPB-module (i.e. one color in Fig. 5) are spread around the stator. Degree B means that combined motor-converter modules are possible where each module consists of one stator segment and one half-bridge of the SPB-C. Degree C means that motor-converter modules are possible where each module consists of one stator segment and one complete SPB-module. The disadvantage of degree B compared to C is that half-bridges of the same three-phase SPB-module are spread around the stator, which requires wiring across converter modules. For each degree of modularity, conditions on

Table II: Criteria C1-C6 for selecting a slot-pole combination.

C1	Winding feasibility	C4	$\text{gcd}(N_l, 2p) > 1$ to avoid unbalanced magnetic pull
C2	High winding factor k_w	C5	High $\text{lcm}(N, 2p)$ for low cogging torque
C3	Low rotor loss indicator $p_{r,\text{loss}}$	C6	High degree of modularity

Figure 5: Defined degrees of modularity A, B, and C of motor-converter systems using the SPB-C and tooth-coil windings for exemplary motors with $N = 27$, $a_p = a_{p,\max} = 3$, $n_l = 2$ and A: $p = 1$ & $m = 9$, B: $p = 13$ & $m = 9$, and C: $p = 12$ & $m = 3$. Windings of the same color (green/red/black) are connected to the same SPB-module as also indicated in Fig. 2.

the slot-pole combination are analyzed. For modularity degree A, no condition applies. For modularity degree B, the conditions in (5) are derived based on the method of the star of slots where $\alpha_{N[-\pi,\pi]} := 2\pi p/N_l$ is the electrical angle between two mechanically adjacent slots in case of a double layer winding and between every second slot in case of a single layer winding. For modularity degree C, it is found that (5) must be fulfilled and that the winding axes of phases connected to one SPB-module must be adjacent. Fig. 2 shows that this is only the case if $m = 3$. Hence, the considered multiphase-motors cannot reach the highest degree of modularity.

$$\begin{cases} a_p = a_{p,\max} \wedge \left(\frac{N_l}{m_l} = 2 \vee \alpha_{N[-\pi,\pi]} = \pm \alpha_z \vee \alpha_{N[-\pi,\pi]} = \pm \left(\frac{N_l}{2l} - 1 \right) \alpha_z \right), & \text{if } \frac{N_l}{m_l} \text{ is even} \\ a_p = a_{p,\max} \wedge \left(\frac{N_l}{m_l} = 1 \vee \alpha_{N[-\pi,\pi]} = \pm \left(\frac{N_l}{2l} - \frac{1}{2} \right) \alpha_z \right), & \text{if } \frac{N_l}{m_l} \text{ is odd} \end{cases} \quad (5)$$

3.3 Optimization of motor dimensions and of electrical motor parameters

In the following, step S2 of the design procedure shown in Fig. 3 is explained. In this step, the motor length l_m and split ratio χ , as well as the electrical motor parameters R , L_{dqk} , and ψ_{PM1} are determined for each of the system configurations ①–⑥ (cf. Fig. 2). The split ratio is defined as the ratio of rotor diameter to outer stator diameter. The following approach is chosen to compare the configurations with regard to power density and efficiency: The three-phase motors (systems ①/③/⑤ in Fig. 2) are designed for maximum power density. For the multiphase motors (systems ②/④/⑥ in Fig. 2), two variants, "a" and "b", are defined. In variant "a" (②_a/④_a/⑥_a), the multiphase motors are designed for maximum power density. In variant "b" (②_b/④_b/⑥_b), the multiphase motors are designed with the same outer dimensions as the three-phase motors in order to compare the achievable efficiencies when the same volume/power density is imposed.

The motor design routine for maximum power density is illustrated in Fig. 6. First, motor specifications are defined. The specifications that are used for all system configurations are summarized in Tab. IX. Therein, T_n , n_n , and D are selected based on application requirements. For the constraints ($J_{\max}$, $p_{Cu,loss,\max}$ & $PF_{\min}$) and for the other specifications in Tab. IX, reasonable choices are made based on literature [14] in order to limit the calculation effort. For the configuration dependent specifications (dc link voltage and slot-pole combination) it is referred to Tab. V and Tab. VIII.

After defining the specifications, the design routine starts in **M1** (cf. Fig. 6) by selecting values for l_m and χ . In **M2**, $\hat{B}_{\delta,1}$, A_δ , and $n_t \cdot i_{q1}$ are calculated using the Carter factor concept [14] and (3), where sinusoidal currents and $B_{fe} \leq B_{fe,sat}$ (no saturation) are assumed. In **M3–M5**, the number of turns per tooth-coil n_t is determined in an internal iteration ensuring that the modulation index $M := 2\hat{V}_{dq1}/V_{dc,mod}$ matches the reference value M_{ref} in steady state, where $\hat{V}_{dq1}$ is calculated using (1). In each iteration step (**M4**), the electrical model parameters R , L_{dqk} , and ψ_{PM1} are recalculated using (6)–(8).

$$R(f) = \frac{n_t N_l}{m a_p^2} \rho_{Cu} \frac{\bar{l}_t}{A_{c,skin}(f)} \quad (6)$$

$$L_{ij} = \frac{\mu_0 r_g l_m}{\delta} \int_0^{2\pi} N_i(\alpha) N_j(\alpha) d\alpha + L_{ij,slot} \quad (7)$$

$$\psi_{PM1} = k_w \frac{n_t N_l}{m a_p} l_m \frac{2r_g}{p} \hat{B}_{\delta,1} \quad (8)$$

Figure 6: Optimization of the motor length and split ratio for a PMSM with maximum power density and specifications given in Tab. V, VIII & IX.

L_{ij} in (7) is the (i, j) -th entry of the winding inductance matrix. The integral term corresponds to the main and the leakage harmonic inductance [15]. The slot leakage inductance $L_{ij, \text{slot}}$ is calculated based on [16]. Inductance L_{dqk} is obtained from the k -th diagonal entry of the matrix $[T]_m [L_{ij}] [T]_m^{-1}$ where $[T]_m$ is the dq-transformation matrix for an m -phase system derived from [17]. In **M6**, the constraints are checked and the motor with minimum length among all valid results is returned. This motor is used for the further design steps S4-S8 (cf. Fig. 3).

3.4 Motor volume and losses

In steps S6 and S7 of the design procedure shown in Fig. 3, the motor volume and the motor losses are calculated. The motor volume is approximated by the cylindrical stator of length l_m and diameter D . Concerning losses, the copper losses are calculated using $\sum R(f)I(f)^2$. For the iron loss calculation in the stator, a reluctance model is used to estimate the magnetic flux density $\hat{B}$ in the stator teeth and in the yoke elements. Then, iron losses are calculated in the frequency domain using the Bertotti equation $p = K_{\text{hyst}} \hat{B}^2 f + K_{\text{ec}} (\hat{B} f)^2 + K_{\text{ex}} (\hat{B} f)^{1.5}$ [18] where the loss coefficients are fitted to loss curves provided by the data sheet of the used laminated steel.

4 Converter modelling and design

In this section, the converter model and the converter design steps of the system design procedure in Fig. 3 are explained.

4.1 DC link dimensioning

For the dc link, film capacitors are considered as they are well suited for the high temperatures occurring in integrated motor-inverter systems [4]. In general, dc link capacitors are selected based on the rated voltage, the capacitance required to limit the dc link voltage ripple, and the current rating. The minimum required voltage rating is determined by the module dc link voltage $V_{\text{dc,mod}}$. The capacitance required to keep the dc link voltage ripple below a predefined maximum $\Delta v_{\text{dc,max}}$ is separately determined for static load conditions ($C_{\text{dc,min,s}}$) and for dynamic load conditions ($C_{\text{dc,min,d}}$) with the methods described in sections 4.1.1 and 4.1.2, respectively. The current rating requirement is also transformed into a minimum capacitance ($C_{\text{dc,min,th}}$) as is explained in section 4.1.3. Hence, the overall condition on the capacitance rating results in $C_{\text{dc,min}} = \max \{C_{\text{dc,min,s}}, C_{\text{dc,min,d}}, C_{\text{dc,min,th}}\}$.

4.1.1 Capacitance required for steady state

In steady state, the voltage ripple is caused by the PWM. Analytical formulae for calculating $C_{\text{dc,min,s}}$ are derived in [7] for various PWM techniques and a current-voltage phase shift angle of $\phi = 0$. As inductive loads with $\phi > 0$ are considered here, a similar procedure is applied to derive (9), valid for arbitrary ϕ and sinus-PWM:

$$C_{\text{dc,min,s}} = \frac{I_{\text{mod}}}{\Delta v_{\text{dc,max}} f_s} \left| \frac{3}{\sqrt{8}} M \cos(\phi) (\delta_{000}(\alpha_{\text{crit}}) + \delta_{100}(\alpha_{\text{crit}})) - \sqrt{2} \cos(\alpha_{\text{crit}} - \phi) \delta_{100}(\alpha_{\text{crit}}) \right| \quad (9)$$

$$\delta_{000}(\alpha) = \frac{1}{2} - \frac{M}{2} \cos(\alpha), \quad \delta_{100}(\alpha) = \frac{\sqrt{3}}{2} M \cos\left(\frac{\pi}{6} + \alpha\right) \quad (10)$$

I_{mod} is the phase rms-current of one SPB-module, f_s is the switching frequency, and $\delta_{000}(\alpha_{\text{crit}})$ and $\delta_{100}(\alpha_{\text{crit}})$ are the duty cycles of the switching states <000> and <100> at the angle α_{crit} of the reference voltage vector. The values for α_{crit} are determined numerically and are shown in Fig. 7. To derive an analytical expression, these angles are least squares fitted to a 2D polynomial, given in (11) with coefficients given in Tab. III. The duty cycle functions in (10) are already known from [7].

$$\alpha_{\text{crit}} = \begin{cases} p(M, \phi) = c_1 \phi^3 + c_2 \phi^2 + c_3 M \phi + c_4 \phi + c_5 M + c_6, & \text{if } p(M, \phi) \in [0, 60] \\ 0, & \text{if } p(M, \phi) \notin [0, 60] \end{cases} \quad (11)$$

Fig. 8 shows the resulting normalized dc link capacitance $C_{\text{dc,min,n}} := C_{\text{dc,min,s}} / (I_{\text{mod}} / (\Delta v_{\text{dc,max}} f_s))$. The result shows that the minimum required dc link capacitance increases towards the maximum modulation index and towards the phase shift angle $\phi = 90^\circ$. Therefore, dimensioning the dc link capacitance for the angle $\alpha_{\text{crit}} = 0$ is not sufficient but the presented formula in (9) should be taken into account.

Figure 7: Critical voltage angle α_{crit} vs. modulation index M and phase shift ϕ .

Figure 8: Normalized capacitance $C_{\text{dc,min,n}}$ vs. modulation index M and phase shift ϕ .

Table III: Coefficients of (11).

c_1	0.03408
c_2	-0.1606
c_3	-0.1852
c_4	0.7286
c_5	0.2910
c_6	-0.3567

Figure 9: Model of a 2-module SPB-C with converter modules modelled as current sources.

Figure 10: Module dc link voltage $V_{dc,mod}$ for a load step from 0 A to 2 A, $L_c = 7.5 \mu\text{H}$, $C_{dc,mod} = 10 \mu\text{F}$, $ESR = 10 \text{ m}\Omega$.

$$C_{dc,min,d} = \frac{4L_c \Delta i_{dc,out}^2}{N_{SPB} \Delta v_{dc,max}^2} \quad (12)$$

4.1.2 Capacitance required for load steps

If cables are used to connect the motor-inverter system to the dc source, load steps lead to an additional ripple on the dc link voltage due to the cable inductance. This effect is analyzed using the model of the 2-module SPB-C shown in Fig. 9 where the SPB-modules are modelled with controlled current sources $i_{dc,out}$. Here, the PWM induced current ripple is neglected so that $i_{dc,out} = \frac{3}{4} M \hat{i}_{mod} \cos(\phi)$. In Fig. 10, the module dc link voltage is shown for a load step of 2 A and a cable inductance of $7.5 \mu\text{H}$. The equation for the minimum required dc link capacitance $C_{dc,min,d}$ is given in (12) for the maximum voltage ripple $\Delta v_{dc,max}$ and load step $\Delta i_{dc,out}$. There, L_c is the cable inductance and N_{SPB} is the number of SPB-modules.

4.1.3 Capacitor current rating requirement

The current rating requirement is based on the thermal condition $T_c < T_{c,max}$ with $T_c \approx T_{amb} + ESR(f_s) I_c^2 R_{th,c}$ where T_c , $T_{c,max}$ and T_{amb} are the capacitor operating, maximum rated, and ambient temperature, ESR and $R_{th,c}$ are the equivalent series and thermal resistance of the capacitor, and I_c is the rms capacitor current. I_c is calculated analytically based on [19]. The thermal condition is translated into a required capacitance $C_{dc,min,th}(I_c, f_s, T_{amb}, T_{c,max})$ which is numerically determined based on data sheet values of a selected capacitor series.

4.2 Volume and loss estimation

Fig. 1 shows the considered motor-converter integration concept. The converter volume is approximated as the sum of the populated PCB volume and the heat sink volume where the populated PCB volume is defined here by the smallest possible cylindrical box around the board with all its components. The height of this box is imposed by the highest component on the PCB which is usually the dc link capacitor. For small dc link capacitors, a minimum height of the populated PCB $h_{PCB,min}$ is defined. The available dc link footprint area is estimated in Fig. 11 for a converter with 2, 3, or 4 SPB-modules. The capacitor height results from this area and the volume that is given by (13) [7]. The coefficients in (13) are given in Tab. X for the selected high energy density capacitor series.

$$Vol_c(C, V_c) = k_1 C V_c^2 + k_2 C V_c + k_3 V_c + k_4 \quad (13)$$

The heat sink shown in Fig. 1 is considered to be thermally isolated from the motor. The heat sink volume is estimated based on the semiconductor losses, their thermal resistances given in Tab. VI, and a cooling system performance index CSPI of $4.4 \text{ W dm}^{-3} \text{ K}^{-1}$ that was already experimentally achieved for naturally convective cooled heat sinks [20]. The semiconductor losses (conduction and switching) are calculated using the equations from [7]. The transistor switching losses are derived based on data sheet values and the method described in [21].

5 Application and results

The system design procedure in Fig. 3 is applied to all system configurations ① to ⑥ (cf. Fig. 2). The system specifications are presented in section 5.1. The results of the design steps S1, S2, and S3-S8 (cf. Fig. 3) are presented in sections 5.2, 5.3, and 5.4, respectively.

Figure 11: PCB-topside of a SPB-C with 2, 3, and 4 SPB-modules for estimating the PCB area available for dc-link capacitors. Transistors are mounted on the bottom-side. The used gate-driver dimensions result from gate-driver *1EDS5663H* (Infineon).

Table IV: System specifications used for all system configurations ①-⑥ (cf. Fig. 2) in the design procedure (Fig. 3).

Rated torque	T_n	(Nm)	50
Rated speed	n_n	(rpm)	300
Heat sink	$CSPI$	(W K ⁻¹ dm ⁻³)	4.4
Cable inductance	L_c	(μH)	7.5
Min. PCB height	$h_{PCB,min}$	(mm)	10
Max. DC link voltage ripple	$\Delta v_{dc,max}/V_{dc,mod}$	(-)	0.02
Max. load step	$\Delta i_{dc,out}/I_{mod}$	(-)	1
Max. junction temperature	$T_{jc,max}$	(°C)	125
Ambient temperature	T_{amb}	(°C)	55

Table V: Considered dc link voltages in V for each system configuration ①-⑥ (cf. Fig. 2) and for device blocking voltages of 200 V/600 V.

Config	$V_{ds,max}$	V_{dc}	$V_{dc,mod}$
① & ②	200	266	133
③ & ④		400	133
⑤ & ⑥		533	133
① & ②	600	800	400
③ & ④		1000	333
⑤ & ⑥		1000	250

Table VI: Specifications of the selected transistors (dimensions: $w \times h \times l$).

Transistor	$V_{ds,max}$ (V)	$R_{ds,on}$ (mΩ)	$R_{th,jc}$ (K/W)	w (mm)	l (mm)	h (mm)
IGOT60R070D1	600	70	1	15.9	14.2	3.5
EPC2034C	200	8	0.3	4.6	2.2	0.79

5.1 System specifications

The system specifications used for all system configurations ①-⑥ (cf. Fig. 2) are summarized in Tab. IV. GaN-HEMTs are used as semiconductor devices to achieve low losses. To evaluate the system performance at different dc link voltage levels, device blocking voltages of 200 V and 600 V are considered. Tab. VI lists the specifications of the selected GaN-HEMTs with these blocking voltages and lowest available $R_{ds,on}$ values. Tab. V summarizes the dc link voltage V_{dc} and the module dc link voltage $V_{dc,mod}$ for each configuration ① to ⑥. There, V_{dc} is limited to 1000 V to stay within low-voltage standards.

5.2 Selection of slot-pole combinations

Tab. VII shows the comparison of slot-pole combinations based on the criteria described in 3.2 for all configurations ①-⑥, slot numbers $N \in [1, 40]$, and pole pair numbers $p \in [1, 20]$, for single and for double layer windings ($n_l = 1, 2$). Columns of pole pair numbers p and rows of slot numbers N with only grey entries (not feasible or $k_w < 0.9$) are not shown for the sake of brevity. The rows are ordered in such a way that the results of the configuration pairs ① & ②, ③ & ④, and ⑤ & ⑥ lie next to each other to enable a direct comparison.

From Tab. VII, the following results: In all (N, p) -combinations where both configurations are feasible (marked with a star), the multiphase configuration (②/④/⑥) has a higher winding factor than the 3-phase configuration (①/③/⑤). On average, only considering the relevant/green combinations, the increase in k_w is 3.7%. As expected, only three-phase configurations can reach the highest modularity degree C. In all cases where the multiphase configurations ensure the modularity degree B, the slot-pole combination suffers from unbalanced magnetic pull (orange) or high rotor losses (red). Therefore, combined motor-converter modularity is practically not feasible for multiphase motors with the considered slot-pole combinations. Furthermore, there are significantly less feasible slot-pole combinations for single layer than for double layer windings. This is due to condition C1a given in (4) that requires an even number of stator slots N for single layer windings.

For the next design steps, only double layer windings are considered and one recommended (green) slot-pole combination is chosen for each configuration ①-⑥. The selected combinations are summarized in Tab. VIII.

5.3 Motor dimensions and electrical motor parameters

The motor length and the electrical motor parameters resulting from step S2 of the system design procedure (Fig. 3) are summarized in Tab. XI, for all system configurations ①-⑥_{a/b} (cf. section 3.3). As explained in 3.3, the multiphase motors of variant "b" are derived by choosing the same dimensions as those of the three-phase motors, i.e. I_m and χ are fixed for system configurations ②_b, ④_b, and ⑥_b in the optimization routine shown in Fig. 6. The motor parameters in Tab. XI are inserted into (1)-(3) in S4 (Fig. 3) of the design procedure. S4 is part of the switching frequency optimization (S3-S8 in Fig. 3), the results of which are presented in the following section.

Table VIII: Selected slot-pole combinations for system configurations ①-⑥ (cf. Fig. 2).

Configuration	①	②	③	④	⑤	⑥
N	36	36	27	27	24	24
p	17	17	12	12	10	11
m	3	6	3	9	3	12
k_w	0.953	0.986	0.945	0.985	0.933	0.991

Table VII: Winding factor k_w and degree of modularity (A/B/C) (cf. Fig. 5) of tooth-coil windings for slot-pole combinations with $N \in [1, 40]$, $p \in [1, 20]$, and system configurations ①-⑥ (cf. Fig. 2). Other criteria of Tab. II are color-coded.

Double layer windings															
N	p Config	4	5	7	8	10	11	12	13	14	15	16	17	19	20
Configurations ① & ②															
12	①		0.933-C	0.933-C									0.933-C	0.933-C	
	②		0.966-A	0.966-A									0.966-A	0.966-A	
18	①			0.902-A	0.945-C	0.945-C	0.902-A								
	②														
24	①					0.933-A	0.949-C		0.949-C	0.933-A					
	②					0.966-A	0.983-A		0.983-A	0.966-A					
30	①								0.936-A	0.951-C		0.951-C	0.936-A		
	②														
36	①									0.902-A	0.933-A	0.945-A	0.953-C	0.953-C	0.945-A
	②										0.966-A		0.986-A	0.986-A	
Configurations ③ & ④															
9	③														
	④	0.985-B	0.985-B						0.985-B	0.985-B					
18	③														
	④			0.940-A	0.985-A	0.985-A	0.940-A								
27	③							0.945-C			0.945-C				
	④					0.914-A	0.954-A	0.985-A	0.994-B	0.994-B	0.985-A	0.954-A	0.914-A		
36	③										0.933-A				
	④								0.903-A	0.940-A		0.985-A	0.992-A	0.992-A	0.985-A
Configurations ⑤ & ⑥															
24	⑤					0.933-C				0.933-C					
	⑥						0.991-A		0.991-A						
36	⑤									0.902-A		0.945-C			0.945-C
	⑥														
Single layer windings															
N	p Config	4	5	7	8	10	11	12	13	14	15	16	17	19	20
Configurations ① & ②															
12	①		0.966-C	0.966-C									0.966-C	0.966-C	
	②														
24	①					0.966-A	0.958-C		0.958-C	0.966-A					
	②						0.991-A		0.991-A						
36	①									0.902-A	0.966-A	0.945-A	0.956-C	0.956-C	0.945-A
	②														
Configurations ③ & ④															
18	③														
	④			0.940-B	0.985-B	0.985-B	0.940-B								
36	③										0.966-A				
	④								0.906-A	0.940-A		0.985-A	0.996-A	0.996-A	0.985-A
Configurations ⑤ & ⑥															
24	⑤					0.966-C				0.966-C					
	⑥														
Colorlegend															
C1-C5 fulfilled			$\gcd(N_1, 2p) == 1$			$p_{r,loss} > 100$			Not feasible			$k_w < 0.9$			

Table IX: Motor specifications used for all system configurations ①-⑥_{a/b} (cf. section 3.3) in the optimization routine shown in Fig. 6.

T_n	(Nm)	50	b_s	(mm)	3
n_n	(rpm)	300	k_{fill}	(-)	0.35
M_{ref}	(-)	0.9	J_{max}	(A mm ⁻²)	7
D	(mm)	250	$p_{Cu,loss,max}$	(kW m ⁻²)	1.1
δ	(mm)	1	PF_{min}	(-)	0.9
h_M	(mm)	4	Iron material	(-)	M330-35a
k_p	(-)	0.9	Magnet material	(-)	NdFeB

Table X: Coefficients of (13) for capacitor series EPCOS B3277x [7].

k_1	cm ³ F ⁻¹ V ⁻²	0.7
k_2	cm ³ F ⁻¹ V ⁻¹	1700
k_3	cm ³ V ⁻¹	0.0016
k_4	cm ³	3.6

Table XI: Electrical motor parameters (R , L_{dqk} & ψ_{PM1}), rated motor current for one SPB-module (I_{mod}), and motor length (l_m) resulting from the design routine in Fig. 6 for system configurations ①-⑥_{a/b} (cf. section 3.3) and $V_{ds,max} = 600$ V.

Configuration		①	② _a	② _b	③	④ _a	④ _b	⑤	⑥ _a	⑥ _b
R	(Ω)	2.05	3.99	3.86	1.41	4.07	3.92	0.82	3.06	2.78
L_{dq1}	(mH)	20.4	41.2	40.3	20.0	60.6	62.0	12.5	49.9	51.0
L_{dq5}	(mH)	-	32.1	31.4	-	36.3	36.8	-	38.0	38.5
L_{dq7}	(mH)	-	-	-	-	23.4	23.4	-	29.2	29.3
L_{dq11}	(mH)	-	-	-	-	-	-	-	17.3	16.8
ψ_{PM1}	(mWb)	280	279	281	328	328	330	301	269	269
I_{mod}	(A)	2.47	2.48	2.47	2.00	2.00	1.99	1.96	1.99	1.99
l_m	(mm)	54	52	54	56	54	56	58	54	58

Figure 12: System efficiency and power density of the 3-phase and of the multiphase system configurations ① – ⑥_{a/b} (cf. section 3.3) for transistor blocking voltages $V_{ds,max} = 600\text{ V}/200\text{ V}$ and switching frequencies $f_s = 1..100\text{ kHz}$. Considered losses: Motor copper and stator core losses, converter switching and conduction losses (cf. sections 3.4/4.2). Considered volumes: Motor, populated PCB and heat sink (cf. sections 3.4/4.2).

5.4 Efficiency vs. power density

Fig. 12 shows the resulting power densities and efficiencies of all considered system configurations for switching frequencies between 1 kHz and 100 kHz where each graph compares one of the 3-phase configurations ①, ③ or ⑤ to its corresponding multiphase configuration ②_{a/b}, ④_{a/b} or ⑥_{a/b} (cf. 3.3). As expected, all configurations show a similar dependency on the switching frequency. At low switching frequencies f_s , large capacitance values are required to limit the voltage ripple on the dc link, leading to non-optimal power densities. Also, the efficiency is not maximum for small values of f_s as could be expected from common pareto curves. This is due to the motor inductance that is held constant in the applied design procedure (Fig. 3) which leads to a high current ripple and high harmonic losses in the motor for low switching frequencies. Increasing the switching frequency increases the motor efficiency but decreases the converter efficiency due to the switching losses. Also, the required capacitor volume is reduced but the heat sink volume to dissipate more switching losses increases. The curve shape near the point of maximum power density (marked with a star) is relatively abrupt for the 2-module configurations and becomes smoother with increasing amount of SPB-modules. This is due to the capacitance limit $C_{dc,min,d} \sim 1/N_{SPB}$ (cf. (12)) that is reached earlier (i.e. for smaller f_s) for a small number of modules compared to a large number of modules. Overall, Fig. 12 shows that multiphase configurations on average enable an increase of 0.70% in maximum efficiency or an increase of 3.8% in maximum power density compared to 3-phase configurations.

6 Conclusion

In this paper, the SPB-C combined with a 3-, 6-, 9-, or 12-phase tooth-coil wound non-salient PMSM is analyzed and compared with respect to efficiency and power density based on a comprehensive system optimization. The results show that the winding factors of the proposed 6-, 9-, and 12-phase (multiphase) motors are on average 3.7% higher than the winding factors of 3-phase motors with 2, 3, or 4 phase-aligned winding systems. The higher winding factor can be used to increase the maximum efficiency by 0.70% or the maximum power density by 3.8%. Furthermore, the possible motor-converter modularity is analyzed. As a result, 3-phase motors allow a higher degree of motor-converter modularity compared to multiphase motors.

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