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Advances in Converter Control and Innovative Exploitation of Additional Degrees of Freedom for Multiphase Machines

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***Abstract*–Multiphase variable-speed drives and generation systems (systems with more than three phases) have become one of the mainstream research areas during the last decade. The main driving forces are the specific applications, predominantly related to the green agenda, such as electric and hybrid electric vehicles, locomotive traction, ship propulsion, 'more-electric' aircraft, remote offshore wind farms for electric energy generation, and general high-power industrial applications. As a result, produced body of significant work is substantial, making it impossible to review all the major developments in a single paper. This paper therefore surveys the recent progress in two specific areas associated with multiphase systems, namely power electronic supply control and innovative ways of using the additional degrees of freedom in multiphase machines for various non-traditional purposes.**

***Index Terms*–Multiphase variable-speed drives, ac-ac converters, inverters, PWM, multiphase power electronics, integrated on-board battery chargers.**

I. Introduction

THE research in the area of multiphase machines and variable-speed drive/generation systems has significantly accelerated since the publication of the last two major tutorial/survey papers [1, 2]. Important new results have been reported in a large number of subareas of this research field, making it practically impossible to provide a meaningful survey in a single paper. Hence the recent advances are surveyed in two separate papers.

This paper predominantly looks at the progress in topologies and control of multiphase power electronic converters for multiphase drive/generation systems. Section II at first examines recent additions to the knowledge in relation to PWM control of two-level voltage source inverters, using star connected load. This is followed by a review of the achievements in the PWM control of multilevel voltage source inverter supply for multiphase drives, using both single-sided multilevel voltage source inverters (VSIs) and dual-inverter supply with the machine's stator in open-end winding (OeW) configuration. Next, the work related to the evaluation of the current and voltage quality, based on the use of total harmonic distortion (THD), is revisited.

An alternative to the VSI supply is the ac-ac matrix converter and substantial progress has been achieved with regard to the PWM control of such converters with a multiphase output or input. Both single-sided supply and OeW topologies have been scrutinised and the most important recent research results are surveyed in the fifth sub-section of Section II. This section concludes with a sub-section containing a brief review of progress in other converter topologies and control.

Multiphase machines are characterised with existence of additional degrees of freedom, since independent flux and torque control of any machine always requires two independent currents, regardless of the phase number. These additional degrees of freedom can be used for various purposes. Some of these date back to the early days of multiphase machines and drives (termed here 'classical') [1,2], while some others are of newer date and have appeared in this century. The latter category is surveyed in Section III of the paper. Progress in series-connected multi-motor multiphase drives since [1, 2] is reviewed first. This is followed by two very specific uses of the additional degrees of freedom that were not known at the time of publication of [1, 2]. The first one relates to capacitor voltage balancing in machines with multiple three-phase windings and multiple three-phase converters that are connected in series, giving an elevated dc-link voltage level. The second one is of potentially huge interest for the future, since multiphase machines enable easy integration of the existing power electronics and the machine into the charging process and vehicle-to-grid operation in electric vehicles (EVs), eliminating the need for installation of a separate battery charging unit and thus providing a saving in the cost, space, and weight.

Advances in the 'classical' use of additional degrees of freedom, such as torque enhancement using low-order stator current harmonic injection and fault-tolerant operation, are surveyed in the companion paper, which also looks at the progress in multiphase machine modelling and control, multiphase generation systems, and multiphase machine design.

II. Multiphase Power Supply Options and Their Control

Since multiphase supply is not readily available, it has to be created from the available three-phase grid. The simplest solution is the design of three-phase to *n*-phase

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transformers, such as for example three-phase to five-phase [3, 4] or three-phase to seven-phase [5] transformer. However, the usefulness of such transformers is limited and they are only a viable solution for laboratory testing of multiphase machines under sinusoidal supply conditions. In real-world applications any multiphase machine would be operated under variable-speed conditions, meaning that a multiphase power electronic converter is required.

Forming a multiphase power electronic converter is simple from the hardware point of view. It is only necessary to make the number of converter legs equal to the number of phases (denoted as n further on) in the case of an inverter, regardless of the number of levels, or to increase the number of switches in accordance with the output phase number in the case of an ac-ac matrix converter. However, the existence of more than three phases significantly affects the PWM control of the converter since the number of 2D planes that has to be considered in the PWM process increases from one in the three-phase case to $(n–1)/2$ (assuming, for the sake of simplicity, that n is a prime number and there is therefore a single neutral point). PWM control of multiphase power electronic converters is surveyed in the subsequent sub-sections.

A. PWM Control of Two-Level Inverters

The principles of PWM control for multiphase two-level VSIs can be considered as well-known and have been surveyed in detail in [1, 2]. The statement applies to continuous PWM techniques for operation in the linear modulation region, of carrier and space vector type, used to generate a single-frequency output voltage. The emphasis has therefore shifted towards other aspects in recent years.

Since a multiphase system can be resolved into $(n–1)/2$ independent planes, it is possible to control independently the same number of sinusoidal voltage references of different frequencies, one per plane. While realization of such multi-frequency output voltage is simple when carrier-based PWM is used, corresponding space vector PWM algorithms are much more involved and are described in [6, 7] for the five-phase and the seven-phase VSIs. A further contribution in the area of multi-frequency output voltage generation is given in [8], where the concept of the space partitioning is applied. The important characteristic of these PWM techniques is that there is no *a priori* limitation on the maximum achievable magnitudes of any of the sets of sinusoidal references. The VSI can operate in the linear modulation region as long as the modulation indices in all the planes satisfy certain relationships, discussed in detail in [9]. Taking standard definition of the modulation index as the ratio between the peak reference in a plane and one half of the dc bus voltage, and using a seven-phase VSI as an example, the resulting linear modulation region is entirely contained within the volume shown in Fig. 1. There are three independent planes in the seven-phase system and hence there are three modulation indices, used as axes in Fig. 1. The values of individual modulation indices for specific points of the body shown in Fig 1 are summarized in Table I. If the output is single-frequency (points A, B and C), the maximum modulation index in the linear region (achieved using the zero-sequence injection) is 1.0257. However, if the output contains three sets of sinusoidal signals of different frequencies, one per plane, the total sum of modulation indices can go up to app. 1.37 (point G in Fig. 1).

A rather general approach to PWM of multiphase VSIs has been developed in [10, 11]. The idea is to split the reference into an integer part and a fractional part; hence the algorithm is applicable to both two-level and multilevel inverters. The algorithm does not rely on vector space decomposition (VSD) technique and therefore operates within a multidimensional (n-dimensional) space. The algorithm has been extended to the multi-frequency output voltage generation in [12] for the case of two-level operation and its performance has been shown to be identical to the one obtainable using the space vector algorithm of [6], which is based on the VSD approach.

Operation of the multiphase two-level inverters can be extended beyond the linear modulation region, into over-modulation. This inevitably leads to an output that will contain low-order harmonics. Hence the idea is always to keep the low-order harmonics for any modulation index in the non-linear region at values that are as low as possible. This can be achieved using different approaches [13-15]. Space vector PWM scheme is developed in [13] and [14] for this purpose, for the five-phase case and the general odd number of system phases, respectively, in which the amplitudes of low-order harmonics are minimized. The effort, described in [15] and based on the space vector approach, attempts to not only improve the operation in the over-modulation region by controlling the low-order voltage harmonics, but also to reduce simultaneously the common mode voltage (CMV). By avoiding the use of zero vectors, a 40% CMV reduction is claimed for the standard four active vector PWM scheme for five-phase VSIs.

CMV has the same cause and the same consequences in all drive systems, regardless of the phase number. Attempts to reduce this unwanted voltage for five-phase drives have

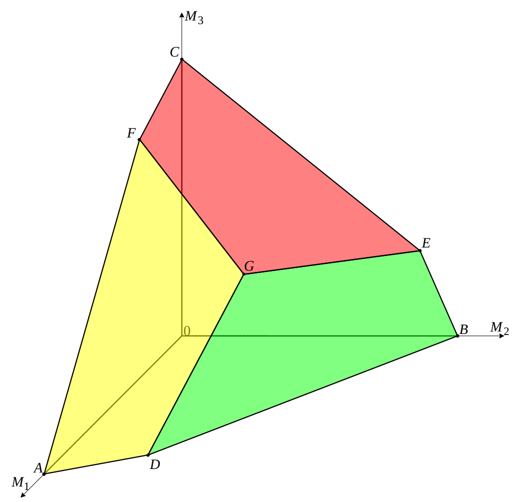

Fig. 1. Volume that defines dc bus utilisation for a seven-phase VSI with excitation in all three planes [9].

TABLE I. CHARACTERISTIC OPERATING POINTS SHOWN IN FIG. 1 [9].

	A	*B*	*C*	*D*	*E*	*F*	*G*
M_1	1.0257	0	0	0.8851	0	0.3159	0.4565
M_2	0	1.0257	0	0.3159	0.8851	0	0.4565
M_3	0	0	1.0257	0	0.3159	0.8851	0.4565

been reported in [16, 17] for operation in the linear modulation region. The approach in [16] is the same as in [15] and, in essence, [15] is an extension of [16] into over-modulation region. Another example of a PWM strategy, aimed at CMV reduction, this time in an asymmetrical six-phase drive with a single neutral point, is described in [17]. The idea of [18] is however very different and it is suggested to incorporate a criterion related to the CMV as a part of the cost function that is used in conjunction with model predictive control (MPC) based current controllers.

Correlation between carrier-based PWM and space vector PWM for two-level multiphase inverters has been explored in detail in [19]. An analytical methodology for the determination of the modulation signals for carrier-based PWM modulation schemes is presented in [20] for the five-phase inverters operated with either balanced or unbalanced loads requiring a single-frequency output. The procedure can be extended to determination of the generalized discontinuous PWM for other phase numbers. A further detailed study of discontinuous PWM strategies for multiphase VSIs is given in [21].

Further advances have also been made in relation to space vector PWM development. For example, a PWM method based on 24 sectors is proposed for an asymmetrical six-phase machine in [22], instead of the standard one that utilizes 12 sectors. Another example, again related to the asymmetrical six-phase machine, is the one of [23], where two three-phase space vector modulators, equipped with over-modulation scheme, are used. A classification algorithm is used for the implementation of the space vector modulation in both linear and over-modulation modes. As is obvious from these considerations, space vector PWM continues to be widely studied, but its implementation is typically based on the carrier equivalent. Nevertheless, implementation of true space vector PWM is of course also feasible and FPGA appears as a viable platform for this purpose [24].

A further recent interesting study includes development of a PWM technique for an unbalanced five-phase system supplied from a six-leg inverter, with the neutral point of the load connected to the sixth VSI leg [25]. Next, space vector based synchronized PWM schemes are reviewed for five-phase and six-phase drives in [26]. Last but not least, reduced-switch-count topologies for multiphase systems require very specific PWM techniques that are only applicable to a particular converter structure. An example is [27], where a six-phase drive system has been investigated.

B. PWM Control of Multi-Level Inverters

Multilevel multiphase supply appears as the natural solution for high power drives. Use of multilevel inverters enables increase of the drive voltage rating (when compared to the two-level inverter supply), while increase in the number of phases enables an increase in the total drive current rating for the given per-phase current rating. It is therefore only too natural that numerous studies have appeared recently related to the various aspects of PWM control for multiphase multilevel drive systems [28].

Multilevel voltage output can be obtained in two different ways. In the first case, discussed in this sub-section, the load is star-connected and there is a single multilevel VSI used as the supply ('single-sided supply' in what follows). It can be of any type, e.g. neutral-point clamped (NPC), flying capacitor (FC), or based on cascaded H-bridge (CHB) topology. The second possibility is to dispense with the star connection of the stator winding and leave the stator winding terminals open, which gives open-end winding configuration [2]. An inverter can then be connected at each side of the winding ('dual-inverter supply' in what follows). By using two inverters with lower number of levels it becomes then possible to achieve exactly the same output voltage as with one inverter in single-sided supply mode, but with a higher number of levels. This is surveyed in the next sub-section.

As far as the multilevel multiphase VSIs in single-sided supply mode are concerned, most of the recent work relates to the three-level NPC inverter type. There are two basic forms of the carrier-based PWM techniques that are typically utilized in conjunction with three-phase three-level VSIs. The phase-shifted PWM (PS-PWM) and the level-shifted PWM (LS-PWM) are the natural extension of the traditional carrier-based two-level PWM. In the PS-PWM a phase shift is introduced between the carrier signals of the cells belonging to the same inverter leg, causing them to switch at different instants. In the LS-PWM the shift is introduced in the vertical positions of the carrier signals. Phase disposition PWM (PD-PWM) and alternating phase opposition disposition PWM (APOD-PWM) are variations of LS-PWM, with the carriers being in phase or in phase opposition, respectively. The extension of carrier-based PWM from three-phase to multiphase case is straightforward, since the same carrier dispositions are possible, and the only difference is related to the level of zero-sequence injection into the modulating signals, which depends on the number of phases [1, 2]. PD-PWM leads to the least distorted voltages for not only three-phase but multiphase systems as well, and is therefore the preferred choice [29]. A modified carrier-based PWM, aimed at three-level five-phase NPC VSIs and claimed to provide an improved performance over the one achievable with PD-PWM, is described in [30].

The number of switching states and the number of available space vectors increase when the number of phases n and the number of levels of the inverter m increase, according to the law m^n. This is summarized in Table II for values of m and n up to 9 where the total numbers of switching states and space vectors are given [31]. Such a situation makes development of space vector PWM for multilevel multiphase inverters a formidable task.

The situation is further exacerbated by existence of (n–1)/2 independent planes [32]. For single-frequency output the non-zero reference exists only in the first plane, while reference in all the other planes is zero, just the same as for the two-level PWM. Hence one again needs to apply n space vectors with (n+1) states. However, finding the most appropriate space vectors is now less than straightforward. The first successful report of a VSD based approach to space vector PWM, for a three-level five-phase NPC VSI, is given in [33]. The important general principle, introduced in [33], is that some switching states can be omitted from further analysis using the 'order-per-sector' rule. In this way the number of switching states of interest is reduced from 243 (Table II) to 113. The same principles are further applied in [34] to develop a VSD based space vector PWM for a three-level seven-phase inverter. 'Order-per-sector' law now enables reduction of the total number of switching states that have to be considered from 2,187 (Table II) to 297. Hence the problem becomes manageable.

However, there are further complications imposed by the multiphase configuration of the system. In principle, each phase number has to be considered separately for each level number. Next, the number of sub-sectors with which one ends up in each sector of the first plane (and hence the number of switching state sequences that need to be considered) significantly increases as the number of phases increases. In a three-phase case there are only four sub-sectors formed within each 60º sector, which are equilateral triangles (Fig. 2a). However, in the five-phase and seven-phase three-level VSI PWM, the numbers of sub-sectors increase to 10 and 18, respectively, within one sector of 36º and 25.7º [33, 34]. Disposition of sub-sectors in the first sector of the first plane for three-level five-phase and seven-phase VSIs is shown in Figs. 2b and 2c. In each sub-sector there is a single space vector sequence that realises the desired reference in the first plane with simultaneous zeroing of the average voltage in all the other planes.

An alternative approach to the space vector PWM, that does not rely on VSD and approaches the problem from the multidimensional point of view, is the already mentioned methodology described in [10, 11]. An approach to the implementation of this technique is given in [35]. A VHDL module is developed, enabling optimized implementation of

TABLE II. NUMBER OF SWITCHING STATES AND SPACE VECTORS (IN BRACKETS) FOR MULTIPHASE MULTILEVEL INVERTERS.

$m\backslash n$	3	5	7	9
2	8 (7)	32 (31)	128 (127)	512 (511)
3	27 (19)	243 (211)	2,187 (2,059)	19,683 (19,171)
4	64 (37)	1024 (781)	16384 (14,197)	262,144 (242,461)
5	125 (61)	3,125 (2,101)	78,125 (61,741)	1,953,125 (1,690,981)
9	729 (217)	59,049 (26,281)	4,782,969 (2,685,817)	~3.9e+8 (~2.5e+8)

TABLE III. NUMBER OF SWITCHING STATES AND SPACE VECTORS (IN BRACKETS) FOR DUAL-INVERTER SUPPLY USING OEW TOPOLOGY.

$m\backslash n$	3	5	7	9
2	64 (19)	1024 (211)	16,384 (2059)	262,144 (19,171)
3	729 (61)	59,049 (2101)	4,782,969 (61,741)	~3.9e+8 (1,690,981)
4	4096 (127)	1,048,576 (9031)	~2.7e+8 (543,607)	~6.9e+10 (~3.0e+7)
5	15,625 (217)	9,765,625 (26,281)	~6.1e+9 (2,685,817)	~3.8e+12 (~2.5e+8)

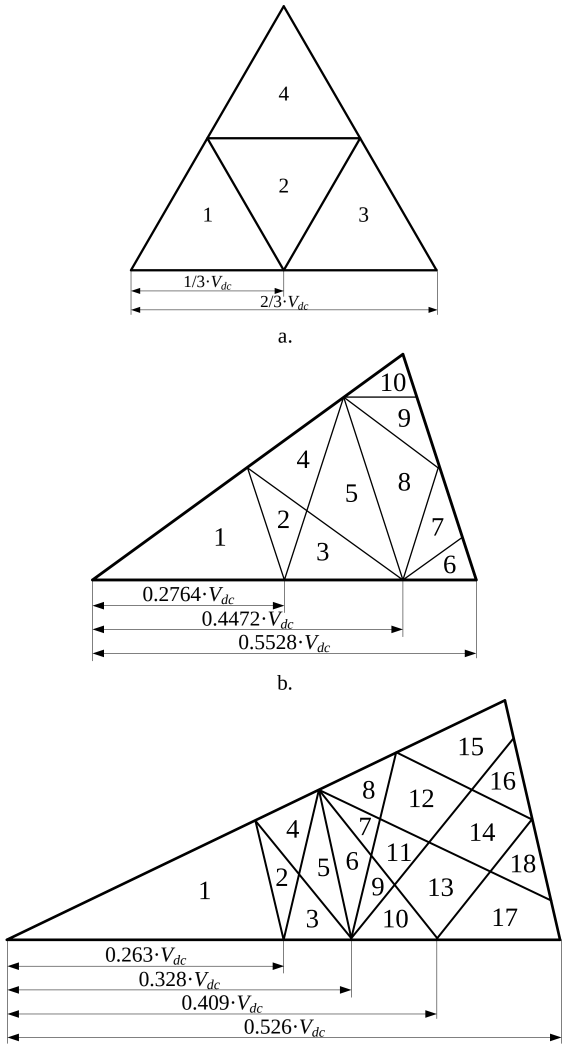

Fig. 2. Sub-sectors of the first sector in the first plane of a three-phase (a.), five-phase (b.) and seven-phase (c.) three-level VSI.

the PWM algorithm and its testing in the FPGA environment. A modified feed-forward version of the multilevel multiphase PWM of [10] is further reported in [36]; it aims at reducing the distortion under balanced or unbalanced dc voltage conditions.

An entirely different approach to space vector PWM of multilevel multiphase VSIs is the one of [37]. Instead of using either multidimensional approach or VSD approach, it relies on the use of single-phase modulators. It is demonstrated that single-phase multilevel modulators can be applied to a VSI with any number of levels and phases.

Another general approach to the modulation for topologies with any number of phases and any number of levels is developed in [38]. The approach is, similar to [10, 11], highly analytical. By applying the eigenspace decomposition of the system matrix, the *n*-phase two-level VSI control problem is solved first analytically. The PWM strategy for a multilevel multiphase VSI system is then obtained by using a concept similar to the integer and fractional reference parts in [10, 11] and is based on the switching strategy of a two-level multiphase VSI. The same eigenspace decomposition based approach is utilized in [39] and, in contrast to [38], finite pulse width resolution is this time taken into account.

In [40] several possible modulation techniques are compared using a cascaded multilevel inverter as the topology. The power quality and semiconductor switching frequency are taken as figures of merit and an attempt is made to reduce the switching loss in conjunction with multilevel sinusoidal carrier-based PWM schemes.

By and large, multilevel multiphase PWM has been predominantly looked at in conjunction with either NPC or CHB inverters. A very different topology is used in [41] where a combination of coupled inductors and a two-level inverter enables three-level like operation of the five-phase system. Similarly, in [42] a PWM control scheme is developed for a three-level five-phase system composed of a dual-buck stage and again a two-level VSI.

C. PWM Control of the Dual-Inverter Supply

Use of dual-inverter supply in conjunction with a machine in OeW configuration offers advantages when compared to the single-sided supply, which include improved fault tolerance (since one inverter can be shut down and operation continued with the healthy inverter only) and lower dc-link voltages (since the output is created using two rather than one dc source). Also, additional diodes/capacitors and capacitor voltage balancing techniques are not required (when both VSIs are two-level). Such a supply topology is well suited to autonomous power systems with multiple dc sources (e.g. electric vehicles) and operation is possible using either isolated dc sources or a single dc source. In the former case it becomes also possible to realize power sharing between the dc sources.

Although three-phase OeW drives have been considered for a variety of cases with differing number of inverter levels at the two winding sides, the existing body of work is in the case of multiphase drives restricted to the use of a two-level VSI at each side of the winding. An OeW topology comprising two VSIs with isolated dc sources and m_1 and m_2 levels, respectively, can create a waveform obtainable in single-sided supply mode with up to m_1m_2–level inverter, depending on the ratio of the dc-bus voltages. With some dc-bus voltage ratios the distribution of the step between two adjacent leg voltage levels is not uniform. In the most frequently analyzed scenario, with two isolated and equal dc-bus voltages ($V_{dc}/2$ each) and $m_1 = m_2 = 2$, the system output is equivalent to the one obtainable with a three-level inverter with the dc-bus voltage V_{dc}.

When compared to the single-sided multilevel supply, development of suitable PWM techniques becomes even more involved with the dual-inverter supply. While the number of semiconductor switches is the same in the two topologies for the same output level number, in single-sided supply there is one state that is not allowed, so that the total number of states is as given in Table II. However, in the dual-inverter topology, all state combinations at two sides of the winding are permissible, so that the number of states further increases. Table III summarizes the number of states and the number of space vectors for *n*-phase OeW topologies assuming that two isolated dc sources are used, each having $V_{dc}/2$ (with V_{dc} being the dc voltage of the single-sided topology) and each with *m* levels. It is simple to establish that the number of states is now $m^n\, m^n$, while the number of space vectors corresponds to the single-sided supply with ($2m$–1) levels [31]. To emphasize this, corresponding rows in Tables II and III are given in the same font color.

The OeW topology has so far been studied predominantly in conjunction with five-phase and asymmetrical six-phase machines. While in the five-phase case one uses two five-phase inverters, the asymmetrical six-phase configuration can be based on the use of two six-phase inverter (and two dc sources) or four three-phase inverters and four isolated (and typically equal) dc sources. The advantage of the latter solution is that one ends up with in essence two three-phase windings, each supplied with two three-phase VSIs. Hence for each of the two three-phase windings one can directly apply the principles of three-phase three-level PWM. Such a solution has been studied in detail, including the aspects of fault-tolerant operation and power sharing, in [43-45]. A space vector based synchronized PWM strategy, in both continuous and discontinuous form, is proposed in [44] for the purposes of control of a six-phase drive supplied from four three-phase inverters fed from four isolated dc sources.

Use of two isolated dc supplies of equal voltage value enables utilization of various PWM strategies. Carrier-based PWM techniques can be applied by simple alteration of the gating signals produced for the three-level FC topology [46, 47] and the same possibilities are at disposal as in the single-sided supply mode. In the case of a five-phase drive, PD-PWM offers the best performance [46], this being the same as for a three-phase drive. However, this is not valid for an asymmetrical six-phase drive, where PD-PWM offers the best performance only insofar as the voltage THD is concerned [47], while the APOD-PWM and the PS-PWM offer the same and superior performance in terms of the current THD. CMV of the PD-PWM also attains considerably higher instantaneous values, when compared with the PS- and APOD-PWM [47].

As far as space vector PWM for dual-inverter supplied drives is concerned, one problem that arises is the way in which the reference voltage is split between the two inverters. A simple solution [48] uses only one two-level inverter as long as the required output is up to half of the maximum achievable in the linear modulation region, while the other inverter is kept in zero state. When the demand exceeds maximum output of one inverter, the second VSI becomes operational as well and provides the rest of the required voltage. The realization is simple, because each two-level VSI can be operated using principles of two-level space vector PWM. The quality of the output voltage is considerably better than in the single-sided supply mode, as only half of the total dc voltage is initially switched.

An alternative space vector PWM method, analyzed in [49], has the same principle of operation as in [48] up to half of the maximum achievable output voltage. Hence the converter is in the two-level mode of operation. Once when the magnitude of the reference voltage exceeds the maximum value obtainable with one inverter, one inverter is operated in ten-step mode while the second inverter is space vector modulated. The second inverter must be able to not only control the fundamental but also eliminate the unwanted low order harmonics. Hence, the multi-frequency modulation technique of [6] is applied to this inverter.

Use of a single dc source as the supply means that the PWM now has to ensure operation with zero average voltage in all the planes except for the first one, and to additionally zero the CMV. A subset of available switching states therefore has to be used, which do not create common mode voltage [50]. Out of these switching states two different switching sequences can be selected. These switching sequences are redundant switching states of the same space vectors. While the average switching frequency is the same for both methods, one provides half as high maximum switching frequency within a fundamental period as the other. This is so because one method switches both inverters throughout the fundamental period while the other switches only the left VSI during the positive and only the right VSI during the negative half-cycle.

As noted in sub-section II-A, the issue of CMV has attracted a lot of attention recently. A comprehensive analysis of the CMV in multilevel multiphase inverter supplied drives, for a variety of specific PWM techniques, has been reported in [51]. A proposal for CMV quantification is set forth, both single-sided and dual-inverter supplied topologies are encompassed by the considerations, and comparison between the two topologies is provided, as related to the CMV.

A modified OeW topology for an asymmetrical six-phase induction motor drive is analyzed in [52]. Two isolated dc sources are used to power two main three-phase inverters, connected to the beginnings of the two three-phase windings, which in essence provide active power to the machine. To eliminate/reduce the unwanted harmonics of the second plane, the other set of terminals of the two three-phase windings is connected to either a common capacitor or to two separate capacitors through the other pair of low-power (and low dc voltage) three-phase inverters. This low-power VSI set in essence takes the role of an active filter in this configuration, and the dual-inverter system operation corresponds to what is usually termed 'bulk' and 'conditioning' mode.

As already noted, the single-sided and the dual inverter supply yield, under the appropriately selected conditions, the same output voltage and the same quality of performance. This is confirmed in Fig. 3 [53], which shows the characteristic waveforms obtained using a three-level five-phase NPC VSI with V_{dc} = 600V in single-sided supply mode and using two two-level five-phase VSIs in dual-

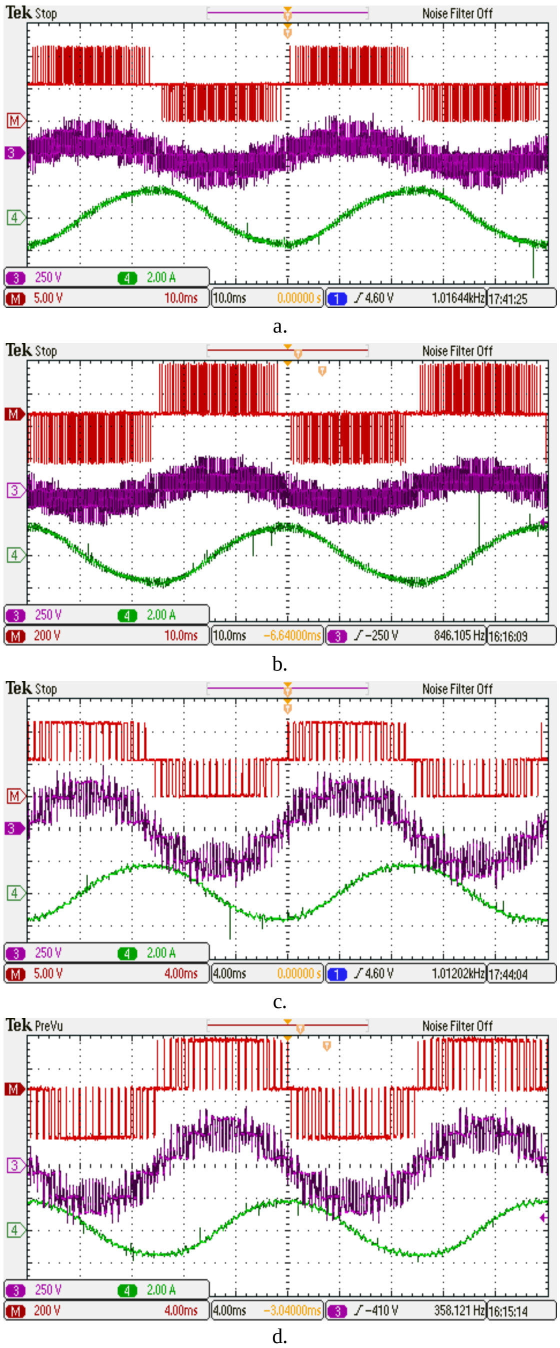

Fig. 3. Time-domain waveforms of the leg voltage, phase voltage and load current obtained using a five-phase induction machine supplied from a three-level five-phase NPC VSI (a. and c.) and using dual-inverter two-level supply (b. and d.) for modulation indices 0.4 (a. and b.) and 1 (c. and d.) [53]. Legend: M-leg voltage, Ch3-phase voltage; Ch4-phase current.

inverter supply mode, with isolated dc sources of $V_{dc}/2$ value. Results are obtained using carrier-based PD-PWM and are shown for two modulation indices, 0.4 and 1. They are, for all practical purposes, identical.

An advantage of the OeW topology over the single-sided supply mode is an improvement in the fault tolerance. Advanced means for fault-tolerant operation of a five-phase OeW motor drive are examined in [54], where the case of a short circuit of an inverter switch is studied and various modes for continued operation are proposed.

An interesting solution for a six-phase system, which combines the use of the bridge converter for one three-phase winding and the use of an OeW system for the second three-phase system (in the form of three H-bridge single-phase converters) has been developed in [55]. Using an appropriate PWM technique, detailed in [55], multilevel-like operation is achieved.

D. Evaluation of the Current and Voltage Quality

Each PWM technique, regardless of the phase number, leads to the switching harmonics, which in turn produce additional losses in the machine. The additional losses can be characterized using the notions of the 'harmonic loss factor' or the 'current ripple rms squared'. A significant effort has been put recently into evaluation of these factors, using various analytical approaches, for numerous PWM schemes aimed at two-level VSIs. Analyses, based on the output voltage harmonics in multiphase drives, have been conducted in [56-58] at a theoretical level for continuous [56] and discontinuous [57] PWM techniques in the linear modulation region, and in the over-modulation region [58].

An alternative approach, which does not require information on the output voltage harmonic content and conducts the analysis in the time-domain instead, typically starts with the derivation of the flux harmonic distortion factor (HDF). Current ripple rms squared is related in three-phase systems to the HDF through a constant and an inductance that is relevant for the switching harmonics. HDF for a specific PWM scheme can be obtained in two ways. The first one follows the space vector approach and, in the case of multiphase machines, it means that HDF has to be determined for each plane separately [59]. Thus it is necessary to deal with, in general, $(n-1)/2$ HDF factors, one per plane. The total HDF is then governed by the sum of all the individual plane HDFs. However, since switching harmonics map into different planes of a multiphase system according to the certain rules [32, 60], and since inductances of a multiphase system that relate to switching harmonics are in principle not of the same value in different planes, calculation of the total current ripple rms squared from the plane HDFs inevitably has to involve additional scaling related to unequal plane leakage inductances [60]. This is a peculiarity of multiphase systems that makes them behave rather differently from the three-phase systems.

Derivation of the total HDF, using space vector approach, for a five-phase system in the linear modulation region, for sinusoidal, sinusoidal with the fifth harmonic injection, and space vector PWM is detailed in [59]. It is shown that HDF of the second plane is the same for all three methods, while the difference caused by the nature (or lack of) zero-sequence injection manifests itself in the different HDF values in the first plane. Corresponding flux HDF and current ripple rms squared determination for asymmetrical six-phase drives with two neutral points has been reported in [22], while a further analysis for two specific space vector PWM schemes for a five-phase drive, using again the space vector approach, is reported in [60, 61]. Switching ripple characteristics are further studied in [62] for the discontinuous PWM techniques for a five-phase VSI. Last but not least, a detailed evaluation of the current ripple rms squared for asymmetrical six-phase drives with two isolated neutral points is undertaken in [63], where the techniques covered in [22] are included in the comparison. The emphasis is however placed on the simplest and therefore frequently used PWM consisting of two three-phase modulators operated with the 30º phase shift. It is shown that, in contrast to all the other available PWM methods, this one utilizes five active vectors in each switching period (instead of four). As a consequence, the double zero-sequence injection method leads to rather high current ripples.

An alternative to the space vector approach for current ripple rms squared determination is an extension of a method frequently used for three-phase systems, known as 'delta' approach, since considerations are based on the analysis of the flux HDF between two inverter output terminals. This approach does not require resolution of the phase variables into their components by means of decoupling (real or complex) transformation. An attempt to extend the delta approach to five-phase drives has been reported in [64]. Unfortunately however, considerations of [64] failed to recognize that a multiphase system is characterized with more than one polygon connection (to be precise, with $(n-1)/2$ different polygon connections). Hence the ripple was in essence considered only for the adjacent polygon connection. In contrast to this, the need to account for all the possible polygon connections in multiphase systems has been recognized in [65], where the approach to flux HDF and subsequent current ripple rms squared determination has been discussed for a variety of phase numbers for sinusoidal PWM with and without the n-th harmonic injection. It is shown in [65] that the polygon approach yields identical total HDF results, by summing individual polygon HDFs, as those obtained using the space vector approach, provided that all the possible polygon connections are encompassed by the analysis. However, if the inductances in different planes of a multiphase system, relevant for the switching harmonics, are not equal (as usually the case is) determination of the current ripple rms squared is only possible using polygon approach by subsequent application of the space vector analysis, in order to relate HDFs of different polygons with corresponding HDFs of different planes (for which the relevant inductances are known). This involves a rather complicated relationship between harmonics in the phase voltages and the line-to-line voltages (since there are $(n-1)/2$ different ones in a multiphase system) [56, 65].

A study of current ripple rms squared and current THD for discontinuous PWM technique is reported in [66] for the five-phase drive system using the polygon approach. The study relates to two different selections of the active space vectors that can be used in the PWM process, with the conclusion that the standard choice of two large and two medium is the best and that discontinuous PWM only improves the switching characteristics at rather high modulation indices (above 0.8).

By far the most important conclusion of the studies related to the HDF and the current ripple rms squared determination is that zero-sequence injection reduces the current ripple in the linear modulation region only in the three-phase systems. For all the other phase numbers the situation worsens if zero-sequence injection is used in the PWM. With regard to the other reason for using zero-sequence injection, improvement in the dc-bus voltage utilization, this becomes practically negligible for higher phase numbers (it is already only just over 5% for the five-phase system). Hence for higher phase number systems the best PWM strategy appears to be pure sinusoidal PWM. Discontinuous modulation is less advantageous for multiphase systems than for three-phase systems [57, 62, 66] and, for phase numbers above nine, any advantages over the continuous PWM methods disappear [57].

Total HDF (and hence the current ripple rms squared, i.e. the harmonic loss, as well) is proportional to the number of phases. An illustration of the per-phase HDF value for multiphase systems with n = 3,5,7,9 is given in Fig. 4 for the sinusoidal PWM (SPWM). As can be seen, the value is the lowest for the three-phase system and it quickly saturates for higher phase numbers, with very little difference between the values for, say, n = 5 and n = 9. Finally, in the over-modulation region, the 3rd voltage harmonic is of substantial value for multiphase systems and harmonic losses are therefore significantly higher than for the three-phase systems. This is so since the 3rd harmonic maps into one of the planes with a very low inductance, thus causing a substantial 3rd harmonic current [58].

The survey in this sub-section has so far been related to output current THD and ripple rms squared value, which are important in order to predict the additional losses that will take place in a drive due to the inverter switching. A rather different type of study, related to the output current ripple peak-to-peak value, has been reported recently in [67-69]. Peak-to-peak current ripple amplitude is analytically determined as a function of the modulation index in [67] for the seven-phase two-level VSI with space vector PWM control. It has been found that the value is practically linearly dependent on the value of the modulation index in the linear region. A similar study, but for the five-phase VSI controlled again using symmetrical centered PWM, is reported in [68] and a thorough experimental verification is also included. A similar idea is pursued in [69] as well, where an attempt is made to provide a general prediction method for the current ripple amplitude using an equivalent single-phase converter.

All the current ripple related considerations so far have applied to the inverter output currents. It is however also possible to evaluate the inverter input dc current and dc voltage ripples. Such a study has been reported in [70] for a five-phase two-level VSI. It was concluded that the input dc current ripple is independent of the modulation technique used, while the input dc voltage ripple is insignificantly lower with the pure sinusoidal PWM than with other techniques that use zero-sequence harmonic injection.

Closely related to the issue of the output current ripple rms squared and current THD is the output voltage quality, which can be expressed using various figures of merits, such as normalized mean square criterion [71] or THD [72]. Frequently used figures of merit in three-phase systems are the weighted THD (WTHD) and normalized THD/WTHD. However, definition of the WTHD requires an assumption that all the voltage harmonics see the same inductance. This unfortunately does not apply to multiphase motor drives, since different planes present different inductances for different harmonics. Thus the THD appears to be the most appropriate indicator of the multiphase inverter output voltage quality [72]. The voltage THD can of course be easily calculated using numerical approach and FFT of the generated inverter output voltage. However, it is possible to find analytical solutions, as is demonstrated in [72] for the pure sinusoidal PWM. Formulae for phase voltage THD for any number of phases are derived for two- and three-level cases, for the most commonly used carrier-based methods. As an example, Fig. 5 shows the output phase voltage THD of the five-phase VSI in two-level and three-level configuration against the modulation index. The continuous lines are the analytical expressions derived in [72], while the discrete values labelled with individual markers are the experimental values. It is obvious that the three-level PD-PWM gives by far the lowest THD values among the considered PWM cases. This of course fits with the discussion of the carrier-based PWM techniques for multilevel VSIs in Section II-B.

E. Control of Multiphase AC-AC Converters

In a general case, a matrix converter can be viewed as an array of $l \times n$ bidirectional switches that are able to transform l-phase input voltages into n-phase output voltages of

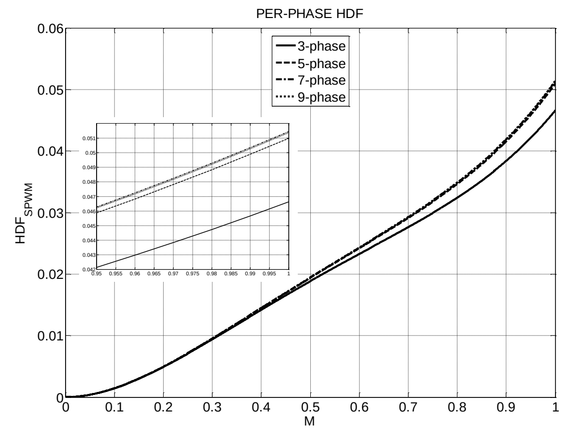

Fig. 4. Per-phase HDF of the sinusoidal PWM for phase numbers n = 3, 5, 7 and 9 with zoomed extract around M = 1 as inset [65].

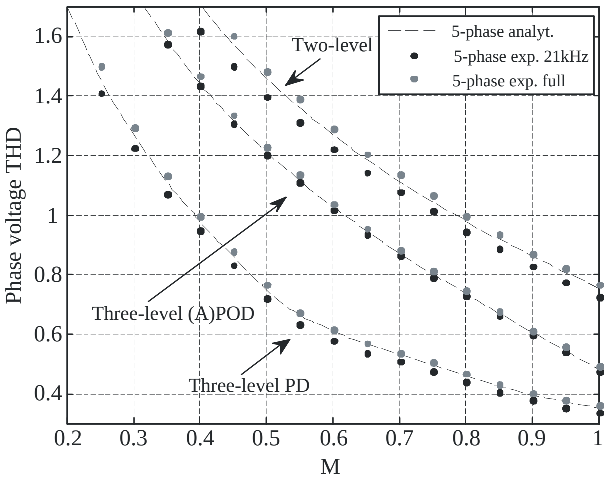

Fig. 5. Comparison of the analytical curves (continuous lines) for the phase voltage THD with experimental results using the full and up to 21kHz spectrum for THD calculation (discrete values labelled with corresponding markers), for the five-phase configuration [72].

variable magnitude, phase and frequency, typically with no energy storage element and with the avoidance of the conversion via a dc-link circuit, as with inverters. For the considered l-phase to n-phase configuration, the theoretical number of switching states amounts to $2^{l \cdot n}$, out of which not all are useful. The actual number of useful states is l^n. The matrix converter has the advantage over VSIs since, instead of a fixed number of levels, it can apply to the output any of the values of the voltages of the input l phases. The result is a complex switching logic.

Probably the first paper ever to address multiphase matrix converters was [73], but the interest in developing PWM algorithms for three-to-n-phase matrix converters has increased substantially since [1, 2] were published. The existing PWM techniques for three-to-three-phase matrix converters are typically modified appropriately to accommodate the higher number of output phases. Most of the existing work in this area relates to direct ac-ac matrix converters. A carrier-based PWM method is adapted to the three-to-nine-phase topology in [74]. Operation with sinusoidal input currents and unity input power factor is feasible irrespective of the load power factor. The main drawback is that the limit of the linear modulation region is reached with an output voltage equal to 76.2% of the input. The same approach to the PWM has also been used in conjunction with a three-phase to five-phase direct matrix converter [75].

An approach with space vector modulation technique for three-phase to five-phase converters is reported in [76]. The resulting input current space vectors form a hexagon, while the output voltage space vectors form three concentric decagons. Out of the 3^5 allowed switching states, which satisfy imposed constraints, 93 are considered useful. This set comprises a zero space vector with triple redundancy and 30 each of small, medium and large vectors, located at the boundaries of the sectors. These space vectors do not have fixed positions in time, as with VSIs. Instead, they oscillate along axes passing through the origin. The rest of the allowed 3^5 switching states result in space vectors with elliptical paths. The modulation uses large vectors only or large and medium space vectors, following the reasoning adopted for space vector modulation of five-phase VSIs. The output voltage is limited to 78.86% of the input in the linear modulation region. In principle, the highest output voltage in the linear region is available for the three-phase output (86.6%) and, as the number of output phases increases, the maximum achievable output voltage reduces [77]. The space vector PWM scheme of [76] is further adapted to the three-to-seven-phase direct matrix converter in [78].

Another modulation technique for three-to-n-phase matrix converters, where n is an odd number, is presented in [79]. In this work the authors extend a PWM method recently introduced for converters with three-phase output to multiphase outputs, using a continuous carrier and the predetermined duty ratio signals to directly generate the gating signals. This explains the origin of the name, the 'direct duty ratio PWM'. A duty ratio is calculated for each switch, and since this applies to each output phase independently, it can be used to supply multiphase loads. The input number of phases is restricted to three, since the modulation requires identification of the values of the input voltages as the highest, medium and the lowest.

An alternative approach to PWM is discussed in [80] for the case of a three-phase to five-phase direct matrix converter. The modulation is developed by using indirect approach, i.e. as a combination of modulations for a three-phase rectifier and a five-phase VSI.

Similar to a dual-inverter supply system it is also possible to use dual matrix converter supply for a multiphase machine, in conjunction with the open-end winding topology. An available study, which discusses a dual direct matrix converter application, using a single three-phase supply and a five-phase load in the OeW configuration, is given in [81]. A carrier-based PWM is utilized and the output reference voltage is equally shared between the two converters. One advantage of the OeW solution, when used in this manner, is that the available output voltage significantly increases in the linear modulation region, compared to what is available in the single-sided supply mode. Since a single input is used for both matrix converters, the modulation has to ensure operation with the zero CMV.

As noted already, with the three-phase input the achievable output voltage reduces as the number of phases increases [77]. However, in the inverse case where the output is three-phase while the input is multiphase, the situation is exactly the inverse one. This means that the output voltage can be higher than the input [77]. The voltage transfer ratio is summarised in Table IV, adapted from [77], for input and output phase numbers up to nine. The situations where it may be advantageous to have a multiphase input while the output is three-phase are related predominantly to interfacing of the renewable electric energy generation sources (e.g. wind, small-scale hydro, and similar) to the grid. Such a direct multiphase matrix converter is considered for wind and hydro power generation in [82], where there are five input phases and three output phases. A carrier-based PWM algorithm is developed for the converter control in [82], space vector PWM methods for the same topology are discussed in [83, 84], and a 27-phase to three-phase matrix converter, aimed at 100 MW power range, is discussed in [85, 86].

Last but not least, in addition to studies related to the direct matrix converters, with either multiphase output or input, there are also similar PWM related investigations for the indirect matrix converter topology, which has a rectifier and an inverter stage. The two stages, rectifier and inverter, are controlled independently but are synchronised. The rectifier control aims to create maximal dc-bus voltage and minimal distortion of the input current, which is controlled to be in phase with the input voltage. The inverter control relies on accepted modulation techniques for multiphase VSIs. A carrier-based PWM for a three-to-five indirect matrix converter is elaborated in [87].

TABLE IV
OUTPUT TO INPUT VOLTAGE RATIO FOR VARIOUS MATRIX CONVERTER TOPOLOGIES [77].

V_o/V_i (%)		Output phase number						
		3	4	5	6	7	8	9
Input phase number	3	86.6	75	78.8	75	76.9	75	76.2
	4	81.6	70.7	74.3	70.7	72.5	70.7	71.8
	5	104.4	90.4	95.1	90.4	92.8	90.4	91.8
	6	100	86.6	91	86.6	88.8	86.6	87.9
	7	109.8	95	99.9	95	97.5	95	96.5
	8	106.7	92.4	97.1	92.4	94.8	92.4	93.8
	9	112	97	102	97	99.5	97	98.5

F. Other Converter Control Related Issues and Topologies

Inverter dead time is a well-known problem in practical realizations. The impact of dead time on inverter operation is more pronounced in multiphase systems than in three-phase systems [88, 89]. This is so since the largest voltage harmonic, generated by the dead time, is the third; this harmonic is a zero-sequence harmonic in three-phase systems. In multiphase systems this harmonic is not zero-sequence (unless the machine is designed with *k* three-phase windings with isolated neutral points) and it maps in a non-flux/torque producing plane with, typically, very low impedance, leading to a potentially substantial third harmonic current flow.

Although multiphase drive systems are primarily aimed at high-power applications and a well-known method of selective harmonic elimination (SHE) is often used in high power region for three-phase drives, there is little evidence of research related to SHE in conjunction with multiphase inverter supplied drives. The same also applies to other approaches aimed at improving the drive behaviour at low switching frequencies, such as for example THD minimisation. One known example is SHE for an eleven-phase drive, elaborated in [90].

A frequently used supply solution for high-power three-phase drives is a current source inverter (CSI), which can be, in the case of synchronous motor drives, of the load commutated type. Such a supply configuration is compared against the predominantly used multilevel VSI option in [91] and it is shown that the CSI based solution can be a better choice for multiphase drives with three or more three-phase windings (i.e. nine phases or more). A specific CSI fed drive is developed in [92] for ship propulsion and is based on the use of two 7 MW synchronous machines, each of which is equipped with two three-phase windings that are either in phase or shifted by 30º. The principle of SHE is used to eliminate the 5^{th} and 7^{th} motor current harmonics. Test results are reported for the machine with two three-phase windings that are spatially in phase and, from the point of view of the converter control, this comes down to a three-phase CSI control. As far as PWM control of true multiphase CSIs is concerned, there is little evidence of progress at present. An exception is [93], where a carrier-based PWM technique for a five-phase CSI is described.

Finally, some less traditional multiphase inverter structures and their control are elaborated in [94, 95]. A Z-source six-phase inverter is analysed in [94], while [95] introduces a novel six-phase topology aimed at medium-voltage drives. The inverter is called hexagram, and it is composed of six two-level VSI modules. The output is multilevel and the inverter enables reduction of the current and voltage stress, while offering modular structure.

III. INNOVATIVE WAYS OF EXPLOITING THE ADDITIONAL DEGREES OF FREEDOM

As explained in Section I, only the non-traditional uses of additional degrees of freedom that exist in multiphase machines are surveyed here. This includes progress in multi-motor drives with a single inverter supply, as well as two completely original and novel applications of the additional degrees of freedom that have emerged in the last few years: capacitor voltage balancing in series-connected voltage source converters and integrated on-board battery charging for electric vehicles.

A. Multi-Motor Multiphase Drive Systems with Single VSI Supply

As explained in [1, 2], it is possible to realise a multi-motor drive system, with independent field oriented control of all machines, by using a single multiphase VSI and an opportune series connection of the stator windings of multiphase machines. The number of machines that can be connected in series depends on the phase number (two for a five-phase system, three for a seven-phase system, etc.). The concept was invented in the beginning of this century and has been further advanced since publication of [1, 2] in various ways.

Detailed steady-state modelling of series-connected two motor five-phase and six-phase drive systems is described in [96], where it is shown how the equivalent circuits in different planes combine to yield an overall equivalent circuit representation of the complete drive system. Further relevant insights into the properties of the two-motor six-phase drive are given in [97], while [98] discusses a convenient PWM method for the same drive structure. Advanced control aspects are covered in [99].

A filter is designed for an asymmetrical six-phase machine in [100], to reduce the low order harmonics in non-PWM operation. The filter is inserted in series with motor and the work is inspired by the series-connected multi-motor drive systems with phase transposition. Basically, the filter is stator-like, but has different phase sequence either spatially or in the connection to the motor, so that the impedance of the first plane of the filter is put in series with the impedance of the second plane of the motor, and vice versa.

An application-related example, describing a positioning system with two series-connected five-phase tubular PM actuators supplied from a single five-phase inverter, is given in [101]. Although the complete concept of the series-connected multi-motor multiphase drive systems has been originally developed for the machines with near-sinusoidal magneto-motive force (MMF), it is shown in [102] that the idea can be successfully extended to the machines with non-sinusoidal MMF.

Finally, the well-known general equivalence between series and parallel connections in electrical engineering suggests that the concept of series-connected multiphase multi-motor drive systems should have its parallel-connected equivalent. This is indeed the case, as elaborated in [103]; however, in contrast to the series connection, which is a potentially useful way of building a multi-drive system, parallel connection is without any prospect for real world applications [103].

B. Capacitor Voltage Balancing with Additional Degrees of Freedom

The most frequently considered topologies of multiphase drive and generation systems are those that are based on use of multiple three-phase windings (e.g., six-, nine-, twelve-phase). This is so since in such a case it becomes simple to use the existing three-phase technology and, importantly, it is also simple to produce a modified stator winding since typically there is no need to produce new stator laminations – rewinding suffices. When such machines are used, it is customary to parallel dc sides of the three-phase converters to the same capacitor (i.e. dc link), so that the additional degrees of freedom can be used for post-fault control.

Another possibility is to connect the three-phase converter dc sides in series. This means that the natural means for fault-tolerant operation, which are inherent in multiphase drives, are lost. However, series connection opens up the possibility of using a machine and three-phase converters of low-voltage rating for operation in conjunction with a dc link of elevated voltage rating. The concept is in principle applicable to any motor/generator with k three-phase windings ($k \geq 2$) and k three-phase voltage source converters. It has been explored in detail in [104] for the case of k =2 (i.e. for an asymmetrical six-phase topology). A problem observed in the actual operation is the drift between the voltages of the capacitors on the converter dc sides, caused purely by the inherent asymmetries between the two three-phase windings. It is shown in [104] that this drift can be eliminated by using an additional dc-link voltage controller, which is cascaded with the current controller for the second plane of the six-phase system. This is an entirely different use of the additional degrees of freedom in multiphase systems from all the other currently known applications and it illustrates the versatility of the concept. Fig. 6 illustrates system operation before and after activation of the dc-link voltage controller (V_{dc1} and V_{dc2} are voltages of two dc-link capacitors). As can be seen, significant initial voltage drift is quickly compensated once when the control is activated.

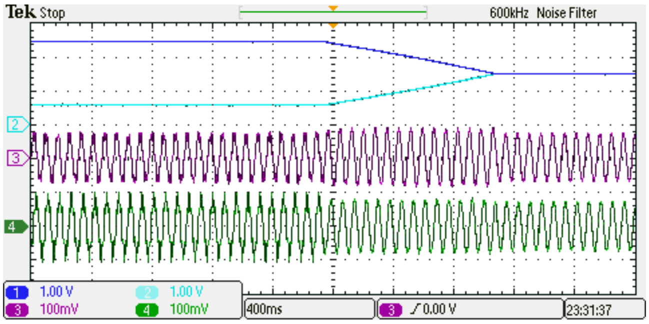

Fig. 6. Experimental recording of the effect of the activation of the dc-link voltage balancing controller at t = 2.0s (mid-point along time axis) for no-load operation at 250rpm: Channel 1: V_{dc1} (100V/div), Channel 2: V_{dc2} (100V/div), Channel 3: phase-$a1$ current (1A/div), Channel 4: phase-$a2$ current (1A/div), Horizontal: Time (400ms/div). (Markers for Channels 1 and 2 are overlapped) [104].

C. Integrated On-Board Battery Chargers for EVs Based on Multiphase Machines and Converters

Current generation of EVs utilize by and large three-phase machines of either permanent magnet or induction motor type, in conjunction with a three-phase voltage source converter. Both the motor and the converter are typically used in propulsion mode only, while in the charging mode a separate charger unit is utilized. This is not economical and a lot of effort has recently been put into investigating the means for integration of the propulsion mode components into the charging and vehicle-to-grid (V2G) modes. The idea is to use the motor's stator windings as filters and to use the propulsion inverter as the rectifier/inverter in the charging/V2G modes. This in essence means that the machine is to be used in the OeW configuration during charging/V2G operation and the current should flow through the motor stator winding; but, the machine should not generate any torque. If the machine is a three-phase squirrel cage induction or permanent magnet synchronous, then such a solution is only possible in conjunction with single-phase (slow) charging; for fast charging a three-phase supply is to be used and it is not possible to avoid rotating field (and hence torque) generation in the three-phase machine when the three-phase currents are passed through it.

The situation however changes fundamentally if the propulsion motor is a multiphase machine. The discussion further on is equally applicable to permanent magnet (i.e. synchronous in general) and induction machines. Recent surveys of integrated battery chargers [105, 106] clearly show the advantages of such solutions. Probably the most developed one at present is based on the concept exposed in [107]. In propulsion mode the machine operates as a standard three-phase one and the only difference, compared to the standard solutions, is that an H-bridge is used to supply each phase, meaning that there are effectively six inverter legs. The machine is however non-standard, since mid-points of each stator phase have to be brought to the terminal box. For charging/V2G, the three-phase mains are connected to the phase mid-points. Hence half-windings of each phase, which are in spatial opposition, carry the same current in charging, providing complete field cancellation. The system thus uses the same converter and motor in both propulsion and charging/V2G, there is no torque developed during charging, and there is no reconfiguration required. Effectively, during charging the machine is used as a symmetrical six-phase machine and the zero torque production is ensured by diverting stator currents from the flux/torque producing plane into the non-flux/torque producing plane. The excitation of the first plane is in this system for charging/V2G identically equal to zero.

The same system is further studied in [108, 109] with the emphasis placed on the propulsion mode. Although the machine is three-phase in motoring, three rather than two current controllers are required due to the use of the H-bridge for each phase, which means that zero-sequence current flow is possible. Detailed considerations are given for the operation in the field-weakening region as well. Overall control for all the operating modes of this system is discussed in [110].

An alternative solution for three-phase charging, based on the asymmetrical six-phase machine, is discussed in [111]. For propulsion, each phase pair (e.g. *a*1, *a*2) is connected in series, giving three-phase system like operation. For charging one three-phase winding is connected to the grid, while the other gets connected to the battery through the inverter. This gives galvanic isolation but is rather complicated and the charging power is at most 50% of the propulsion power. The machine however rotates and power is transferred from one to the other three-phase winding by induction during charging. This is a drawback of the solution since it requires an additional clutch.

Any machine with multiple three-phase windings yields a simple and straightforward method for integrated on-board single-phase charging. All that is needed is to connect the single-phase grid to the two isolated neutral points of two (out of *k*) three-phase windings. Since each phase of a three-phase winding carries the same current, the field in the machine is zero. The concept is illustrated for an asymmetrical six-phase machine in [112]. With a symmetrical six-phase machine one can use either two neutral points or, if the winding structure is as in [107], midpoints of two stator phases [113].

The idea of connecting the grid to the neutral points of isolated three-phase windings is extended from single-phase to three-phase charging in [114]. The principle outlay of the system is shown in Fig. 7, where the machine can be either symmetrical or asymmetrical nine-phase, of either induction or permanent magnet type. It should be noted that, in principle, the same idea can be extended to 12-phase, 15-phase, etc. systems. Nine is the minimum number of phases with which the three-phase charging becomes possible in this manner. If there are more than three three-phase windings, then some neutral points are not used in the charging process.

As can be seen, each phase in each of the three three-phase windings will carry the same current in charging/V2G modes; this means that the resulting field of each three-phase winding will be zero at all times, giving overall zero current and flux components in the flux/torque producing plane of the nine-phase system. Instead of mapping into the first plane, grid currents map into the zero-sequence component and a pair of components in one of the non-flux/torque producing planes in the case of an asymmetrical nine-phase machine [114]. As the first plane is not excited, the torque during charging/V2G modes is at all times equal to zero. Moreover, the system does not require any hardware reconfiguration and, due to its modularity, is very well suited for future EV applications. The system gives full functional integration and it seems as potentially ideal to combine this with the full design and constructional integration of a nine-phase drive for EVs reported in [115].

Experimental verification of all operating modes for the system of Fig. 7 is offered in [114]. As an example, Fig. 8 illustrates transient that follows initial operation in the V2G mode, with the final steady state corresponding to the battery charging. Unity power factor operation is evident, since grid current changes from being in phase opposition to being in phase with the grid's phase voltage.

If the machine of choice is a six-phase one (symmetrical or asymmetrical), there are only two neutral points and the idea of Fig. 7 cannot be used. However, it is now possible to combine the isolation requirements with the need to have a six-phase supply and to again divert either all or a part of the currents from the flux/torque producing plane to the non-flux/torque producing plane [116]. In the case of asymmetrical six-phase machine, a transformer with two secondaries, connected in star and delta, respectively, is used. For a symmetrical six-phase machine one can use either a transformer with two secondaries giving voltages in phase opposition, or a transformer with a single secondary

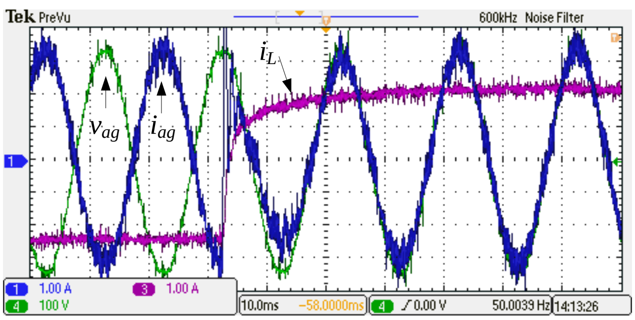

Fig. 8. CH1 - grid phase voltage v_{ag} (100V/div), CH3 - dc-bus current i_L (1A/div), and CH4 - grid phase current i_{ag} (1A/div) waveforms for transition from V2G to charging mode [114].

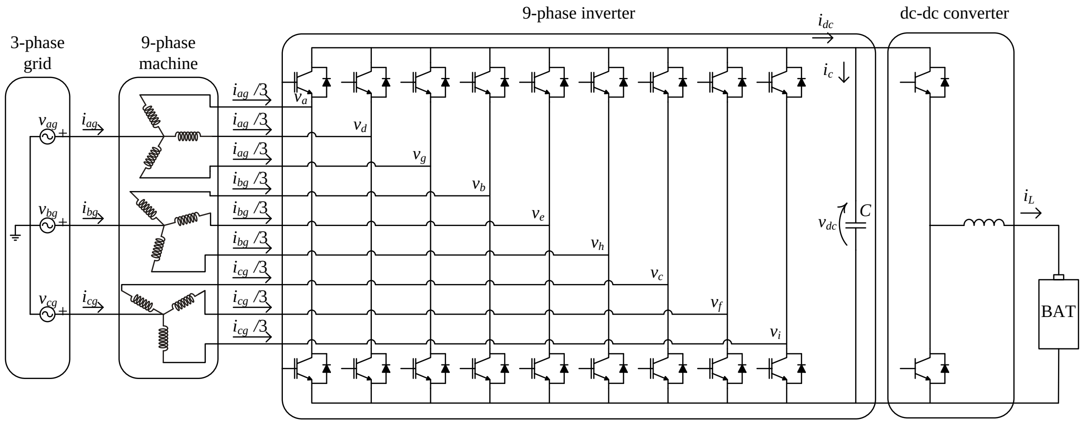

Fig. 7. Topology of the nine-phase integrated on-board battery charger for three-phase (fast) charging. Dc-dc converter may or may not exist – it does not for the results shown in Fig. 8.

in OeW configuration (the transformer is of course off-board). In all scenarios additional switches are required for hardware reconfiguration between propulsion and charging/V2G modes. The principle of phase transposition [1, 2], developed in conjunction with series-connected multiphase multi-motor drives, is always used to divert grid currents into non-flux/torque producing plane.

It should be noted that solutions of [107, 114] and the one with asymmetrical six-phase machine of [116] lead to zero excitation in the first plane. However, the solution of [116] for a symmetrical six-phase machine does cause excitation in the first plane. However, the excitation is along one axis only and it gives rise to the pulsating rather than rotating field, so that no torque can be produced at zero speed. This idea can be used to develop charging schemes for other phase numbers as well. For example, three-phase grid can be used directly in conjunction with an asymmetrical six-phase machine (in contrast to [116]) if the connection scheme of [117] is used: the excitation in the first plane is now along a straight line (i.e. pulsating field), transformer of [116] is not required, and one only needs four additional relays to perform hardware reconfiguration.

The solution which produces pulsating field during charging/V2G operation turns out to be feasible for machines with prime phase numbers as well, in conjunction with three-phase grid charging. Regardless of the phase number (as long as it is a prime number), it is only necessary to connect grid phase a to machine's phase 1, grid's phase b to machine's phases 2 and n, and grid's phase c to machine's phases 3 and (n–1). This will give cancellation of one of the components in the first plane, making the field in the first plane again pulsating [118]. If the number of phases is more than five, multiple solutions are possible. Charging/V2G operation again takes place with zero torque and two switches are typically required for hardware reconfiguration between the charging and the propulsion modes.

IV. Conclusion

A significant growth has taken place in the areas of multiphase machines, drives and generators during the last decade. An attempt has been made in this paper to survey at first the progress in the area of multiphase power electronic converter control. The emphasis is placed on PWM control of two-level and multi-level VSIs, and ac-ac matrix converters with multiphase input/output. Two supply options, single-sided and dual-converter supply, are addressed.

It is obvious from the presented survey that, although significant progress has been made in relation to topologies and control of multiphase power electronic converters, there is a significant lag when compared to the progress in three-phase converters [119]. This applies, first of all, to the multilevel inverter topologies in single-sided supply mode that have been scrutinized so far and that are, by and large, restricted to the standard topologies, such as for example a three-level NPC inverter. Some of the recently introduced topologies for three-phase systems, that have already found practical applications, are the five-level H-bridge NPC inverter, the three-level active NPC inverter, the five-level active NPC inverter and the modular multilevel converter (MMC) [119]. Other topologies, currently under development for three-phase systems, are the transistor-clamped converter, CHB supplied from unequal dc sources, the hybrid NPC-CHB converter, the hybrid FC-CHB converter, and so on [119].

Current multiphase multilevel inverter topologies with dual-inverter supply of an OeW drive are by and large restricted to the use of two two-level inverters with equal dc link voltages or a single dc voltage source. On the other hand, corresponding solutions for three-phase drives are much more versatile and include operation of two two-level converters with unequal dc-link voltages, use of NPC inverter at one or both sides of the OeW, use of the two converters in 'bulk' and 'conditioning mode', respectively, and various even more complex schemes that produce higher numbers of levels at the output [119].

Similarly, progress in three-phase matrix converters includes topologies that have not been so far analyzed in conjunction with multiphase systems. For example, recent advances in three-to-three-phase matrix converters have included development of multilevel matrix converters, which are typically based on cascaded matrix converter, indirect matrix converter-NPC, or FC matrix converter topologies [119].

It is expected that, in future, the emphasis in multiphase power electronic converter control will shift towards afore listed topologies. This will also require adaptation of the corresponding three-phase PWM control methods to the multiphase systems, so that the same quality of operation as in a three-phase case can be achieved. The complexity of topologies is likely to dictate use of carrier-based PWM methods.

Other areas of PWM control, where limited or no progress has been reported, but which have been recently explored in more depth for three-phase systems and are hence likely to attract attention in conjunction with multiphase systems, are predominantly related to high power loads where low switching frequency is used. The examples include the previously addressed SHE, as well as the selective harmonic mitigation [119]. In this context, it is also hoped that synchronized PWM schemes, which have already been developed for multiphase systems, will be experimentally verified and their viability thus proven. It is worth noting that these three PWM methods are essentially steady-state control methods and, as such, invariably restrict the achievable quality of the dynamic response of the controlled plant.

In very recent times some innovative ways of utilizing the additional degrees of freedom have been proposed and these are reviewed as well, with the emphasis placed on the solutions that enable realization of integrated on-board battery chargers for EVs. It is believed that further interesting developments will follow in relation to the integrated on-board battery charger solutions based on multiphase machines and power electronics. This may include, for example, topologies that are simultaneously viable for both single-phase and three-phase charging.

Last but not least, considering that two entirely new ways of using the additional degrees of freedom in multiphase systems have come into existence in the last five years or so, it is hoped that future may result in further inventions of a similar nature.

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