Research and Realization of High-Power Medium-Voltage Active Rectifier Concepts for Future Hybrid-Electric Aircraft Generation

Andrew Trentin, Giacomo Sala, Member, IEEE, Luca Tarisciotti, Member, IEEE, Alessandro Galassini, Member, IEEE, Michele Degano, Member, IEEE, Dmitry Golovanov, Member, IEEE, David Gerada, Member, IEEE, Zeyuan Xu, Member, IEEE, Antonio La Rocca, Carol N. Eastwick, Stephen J. Pickering, Patrick Wheeler, Senior Member, IEEE, Jon C. Clare, Senior Member, IEEE, and Chris Gerada, Senior Member, IEEE

Abstract—In this article, we describe the research and development of a 3 kV active rectifier for a 4 MW aerospace generator drive system demonstrator. The converter is fed by a multiphase high-speed/high-frequency permanent magnet generator. The main aim of the work is to demonstrate for the first time the feasibility of an MW-class generator system meeting future hybrid-electric propulsion requirements. A concept with multiple and isolated three-phase systems feeding different power buses is proposed to meet the availability requirements. Multiple converters (one for each three-phase system) are connected in series and/or in parallel to achieve the rated power and dc-link voltage. This article describes the key design concepts and the development and testing of the converter to meet the challenging application requirements. Reduced power tests are carried out on a full-scale 4 MW converter prototype, validating the proposed design. The work represents a step forward in terms of voltage, power, and output frequency with respect to the state-of-the-art.

Index Terms—Aerospace generation drives, high-power generation systems, high-power high-voltage converter design, hybrid-electric aircraft, multiphase drives, variable speed drives.

I. INTRODUCTION

OVER the past two decades, all the major stakeholders in the aviation industry have been working to define aircraft technology roadmaps for the reduction of net carbon emissions [1]. In Europe, the vision for sustainable aviation set by the Flightpath 2050 program targets a reduction of CO₂ emissions per passenger kilometer of 75% and NOₓ of 90%, compared with the 2000 figures [2], [3]. Within the environmentally responsible aviation framework, NASA has set equivalent goals and timescales to reduce both fuel burn by 50% as well as NOₓ emissions by 75% [4].

A number of programs have led to the development of concepts for a new generation of future air vehicles that are ever-more efficient, environmental friendly, and quiet [5], [6]. These developments are all aiming at reducing the fuel consumption and have a strong direct socioeconomic and environmental impact [7]. The programs mentioned above are tackling the problem from varying angles, for example, trying to improve combustion technologies and/or to look for new disruptive technologies leading toward electric and hybrid-electric solutions. Full electric propulsion systems are not yet able to replace the conventional fuel engines [8]. However, thanks to the on-going development of power electronics, hybrid-electric solutions already have the potential to revolutionize aircraft propulsion and generation systems as well as the aircraft design. A hybrid-electric propulsion system combines the advantages of fuel-based and battery-powered systems and offers new design freedoms [9].

Among all the passenger aircraft, 50–150 seat regional aircraft are considered as suitable candidates for this first short-to-medium term transition, and key companies have put together joint efforts in the attempt to boost the development of this technology [1]. A clear example is the E-Fan X demonstrator, which is a collaboration between Rolls-Royce and Airbus for the
Increasing the speed of generators increases the power density of the electrical machine. However, producing large power at high voltage and speed introduces significant challenges from many points of view, such as increased insulation and power rating of each component, the introduction of rotor dynamics and mechanical stresses due to the generator and turbine interactions, increased fundamental frequency with consequent increase of ac losses in the generator, and increased loss density requiring more intensive cooling systems for both the converter and electrical machine. The achievable power and power density and reliability/availability of the drive are limited by state-of-the-art technologies.

Many of these limitations can be mitigated, especially when high power density and fault tolerance are required through the use of multiple three-phase or multiphase sets of windings [21]. In the literature, a wide range of multiphase electrical machine designs have been proposed to improve system power density in applications, such as high-power turbocompressors [22], electric ship propulsion [23]–[25], and other safety critical applications [14]. The key advantages of a multiphase solution can be summarized as a split of the machine power among a higher number of phases, flexibility in terms of architecture layout, and more importantly for aerospace applications, increased availability [26].

Similarly, to enable the use of lower voltage rated devices, power electronics can be designed using multilevel converter topologies. Although series (cascade) connected switching devices may be considered as an alternative, their use reduces the maximum switching frequency and the efficiency of the converter [27]–[29]. Therefore, multilevel converter topologies are often applied to MV and HV three-phase drives [30]. The main disadvantage of multilevel converters is the increased number of components and, thus, lower power density with respect to simpler topologies. However, this issue can be addressed by an optimized converter design for the specific application with neutral point clamped (NPC) and cascaded H-bridge (CHB) converters being particularly considered for drive applications [31]. The NPC converter is a three-level converter with a structure that requires the active balancing of its two dc-link capacitors, and it is difficult to scale to a higher number of levels [32], [33].

By taking advantage of a multiphase electrical machine architecture, it is also possible to combine the advantages of the NPC and CHB converter topologies to achieve a multiphase and multilevel drive architecture. In fact, in a multi three-phase drive, the flexibility of the architecture allows the assembly of multiple three-phase converters in series and parallel layouts.

This work presents and demonstrates the challenges in developing a 4 MW (5 MVA) 3 kV high power-density drive, suitable for the future on-board generating systems of hybrid-electric aircraft. The article presents an architecture of a high-power medium-voltage drive, featuring a high-speed multiphase machine with a permanent magnet rotor and multiple isolated three-phase windings in the stator. In this article, the converter structure is designed with a cascade and parallel architecture of NPC submodules, and the design choices are described in detail. This demonstrates that, with the current state-of-the-art component technology, a high power-density drive for future hybrid-electric platforms can be achieved with innovative approaches to the architecture.

The research contributions of this work can be summarized as follows.
1) For the electrical machine design adopted, a feasibility analysis for different candidate converter structures and devices is performed.

2) Based on the selected topology, a control scheme that combines several control operating modes and voltage balancing systems is designed to address the specific requirements of the converter and application.

3) A full-scale prototype has been constructed and tested, thus validating the control approach and design as well as providing efficiency measurements for the full generation system.

II. HIGH-LEVEL CONVERTER DESIGN

The three main parameters that need to be considered when designing a medium-voltage aircraft dc generation system are as follows.

1) The electrical machine’s nominal speed and, therefore, its fundamental electrical frequency. In the case considered, it is set to 1 kHz.

2) The dc-link nominal voltage. For this development, the requirement is to have two independent dc output ports, i.e., two isolated dc links each with a dc voltage of 3 kV.

3) The nominal power of the system, in this case, 4 MW (5 MVA) was targeted.

It is important to highlight that the design decisions are also affected by the generator design. Although the article focuses on the power electronic conversion system, information about the generator is given in Appendix A.

A. Device and Operating Switching Frequency Selection

As already demonstrated in previous work [45], the converter switching frequency should be at least ten times higher than the fundamental frequency. Thus, considering the generator fundamental frequency of 1 kHz, the minimum switching frequency for the power conversion system is equal to 10 kHz. However, a further increase in the switching frequency benefits the converter operation by reducing the phase current ripple and the output voltage low-frequency harmonic content as well as increasing the current controller bandwidth that is obtainable. The benefits of increasing the switching frequency are even more pronounced when dealing with a high-power and high-efficiency generator, which presents a small value of leakage inductance. On the other hand, increasing the switching frequency also increases the switching losses limiting the selection of power devices. For these reasons, the switching frequency has been set at a tradeoff value of 15 kHz, where both SiC MOSFETs and traditional Si insulated-gate bipolar transistor (IGBTs), rated at 1.2 kV, are viable options. The higher voltage rating IGBTs are unlikely to perform adequately at a switching frequency of 15 kHz. For example, Si IGBTs rated at 3.3 kV are intended for applications where the maximum switching frequency is in the range of 2–3 kHz [34], [35].

Similarly, 1.7 kV SiC MOSFETs are still rarely used due to procurement issues and SiC devices for voltages higher than 1.7 kV are not yet commercially available.

B. Converter Architecture

To work within the limitations of the power devices, various converter architectures can be considered. In this article, only two- and three-level NPC I-type converters are considered. The reason for this choice is mainly due to the manufacturer availability and lead time for the related power modules.

The three-level NPC T-type converter, which presents a better efficiency than other topologies, has been discarded as it does not allow the use of lower voltage rated devices [36]. Also, the three-level active NPC has been excluded because of the high number of devices per leg. Similarly, more complex layouts with series-connected devices have not been considered since the available switching frequency is reduced, owing to the parasitic components in the commutation circuit, which could lead to possible overvoltage stress of the device die [37].

In conclusion, 1.2 kV devices are considered in this work. For these devices, the typical dc-link voltage for an active frontend is 700–750 V (using a two-level converter). Therefore, to meet the specification of a 3 kV dc voltage by using only two- and three-level converters as building blocks, a number \(N_v\) of these converters must be connected in series.

If two-level converters are connected in series, four modules are needed to sustain a dc voltage of 3 kV. While, only two converters in series are sufficient if three-level I-type power modules are considered. In fact, the three-level I-type topology features an inherent voltage-sharing capability that makes the converter able to sustain a dc link of 1.5 kV, halving the required number of series converters.

C. Selection of the Optimal Number of Phases

The final consideration is the nominal power of the conversion system. In this specific case, the converter is required to convert a power of 4 MW that is generated from an electrical machine, which has a power factor (PF) of 0.8, resulting in an apparent power of 5 MVA. From these data, it is straightforward to calculate the nominal power of every single converter, once the number of series \(N_P\) and parallel \(N_v\) power modules is defined. The total number of converter cells is \(N_c = N_v N_p\), and the total number of phases of the multi-three-phase electrical machine is \(N_{ph} = 3 N_c\).

In order to establish the output voltage of each cell, the maximum modulation index is set to 95% for the two-level solution, whereas for the three-level, it is set to 80% to take into account some margin for the internal voltage balancing. At this point, the rms phase current is evaluated from the power of each cell (5 MVA/\(N_c\)). Once the rms phase current is calculated, it is possible to estimate the rated current of the power module (the typical value at a temperature of 25 °C) by doubling the phase current. However, considering the relatively high switching frequency, the high power, and the complex converter layout, the rated current of the Si IGBTs power modules is considered three times the phase current for a safety margin. If SiC MOSFETs are used, this can be reduced to two times the phase current because of the lower switching losses [38].
TABLE I
SINGLE-CELL POWER, PHASE CURRENT, AND VOLTAGE (RMS VALUES) FOR DIFFERENT CONFIGURATIONS USING A TWO-LEVEL CONVERTER

| $N_s$ | $N_p$ | $N_{ph}$ | Cell Power [VA] | Phase Voltage [Vrms] | Phase Current [Arms] | SiC current rating [A] | Si current rating [A] |
|-------|-------|----------|-----------------|----------------------|----------------------|------------------------|----------------------|
| 4     | 2     | 8        | 24              | 62500                | 291                  | 716                    | 1432                 | 2149                 |
| 4     | 4     | 16       | 48              | 31250                | 291                  | 358                    | 716                  | 1047                 |
| 4     | 6     | 24       | 72              | 20800                | 291                  | 239                    | 477                  | 716                  |
| 4     | 8     | 32       | 96              | 15600                | 291                  | 179                    | 358                  | 537                  |

D. Summary

As a final remark, it must be considered that a higher number of cells results in a more complex electrical machine and converter design, in terms of sensors, gate drives, and control. Ideally, the best solution is the one that minimizes the number of cells while respecting the device’s capabilities.

Looking at Table I, the configurations featuring two-level power modules based on SiC MOSFETs will require a high number of converters (72 or 96), owing to the limitation of the currently commercially available devices, with current ratings up to 450 A. For this reason, the two-level SiC MOSFET modules are excluded. Conversely, the market availability of two-level power modules with Si IGBTs is high. Considering off the shelf components and assuming one power module is used per phase, there are multiple choices in terms of rated current: 450, 600, 900, 1200, 1400, 1500, and 1800 A, respectively. The solution with the lowest number of converters is the one with four in series and four in parallel, which leads to a total of 16 modules supplying a 48-phase electrical machine. The rated current of the power module is 1200 A. As a reference choice, the Infineon FF1800R12IE5P could be considered. This solution is given in Table I, highlighted in bold.

In the case of three-level I-type power modules, there is a limited number of available choices. However, it is possible to find power modules with rated currents from 400 to 1800 A that use Si IGBT, while presently, there are no power modules that have SiC MOSFETs for this converter structure, at least for this current range. As for the previous case, the solution with the lowest number of converters (i.e., eight modules rated at 1200 A) is the preferable choice. This leads to the Vincotech 70-W624N3A1K2SC-L400FP power module selection. This solution is given in Table II, highlighted in bold.

The final choice for the converter architecture was made by also considering its effect on the reliability of the system. The main drawback of using power cells connected in series is that if a single cell stops working (e.g., due to a fault detection), the full converter series needs to be disabled. Contrarily, in the case of parallel cells, each cell can still operate even when one or more cells are turned off. It results that reducing the number of converters in series and increasing the number of converters in parallel is a suitable choice for the reliability of the full system, which is able to deliver more power under a fault detection in a single cell. In conclusion, the three-level converter solution, highlighted in bold in Table II, has been selected due to the reduced number of phases ($N_{ph} = 24$) and the effectiveness of its redundancy, resulting from the limited number of cells in series ($N_s = 2$).

III. SELECTED CONVERTER DESCRIPTION

The selected topology for this prototype is a modular multilevel and multiphase converter. Two independent dc-link channels, namely BRAVO and ECHO, are designed each with series–parallel connection of four three-level I-type NPC cells, as shown in Fig. 2. The pairs of NPC converters are connected in series to provide a 3 kV dc link. Each three-phase NPC is connected to one of the eight star-connected three-phase machine windings and, consequently, processes 0.5 MW of power. Each channel (BRAVO or ECHO), therefore, interfaces to four star-connected three-phase ac generator windings. Each NPC,
as shown in Fig. 3(a), comprises three three-level IGBT power modules from Vincotech (70-W624N3A1K2SC-L400FP) and two custom-made dc-link capacitors, as shown in Fig. 3(b), each of 600 μF.

When multiple active rectifiers are integrated into a single application, there are both design and control challenges. The control must implement balancing algorithms to maintain the voltages of all the capacitors close to the same value. Also, in the converter design, the modularity of the layout has to be carefully integrated into the overall structure, as shown in Figs. 4 and 5. The power modules and the capacitors have been arranged to minimize the mass and the volume of the overall converter. The four series NPC clusters have been mechanically grouped into four layers with the arrangement, as shown in Fig. 4, and stacked one on top of the other, as in Fig. 5, where the full converter prototype is shown.

The mass of the prototype converter is 240 kg, and its volume is 411 L, so the power density is ~20.8 kVA/kg and ~12 kVA/L. The power density is calculated using the mass of all components and the volume within the cuboid converter structure envelope, not including coolant mass. Considering that the converter is developed with commercially available Si IGBT technology, this is an outstanding power density with respect to the reference values presented in the most recent developments [5], [6], [36]. To control the converter, two independent control systems are used (namely BRAVO and ECHO). These two systems are identical in their structure and they are synchronized through a serial communication with a bandwidth of 10 MHz. In this way, at every sampling instant, the two systems can exchange information about their status and all the relevant information to implement the field-oriented control. Fig. 6 shows the structure of the control hardware. Each controller employs one TMS320C6713 digital signal processor (DSP) and two ProASIC3A3P400 field-programmable gate arrays. Both the control systems (BRAVO and ECHO) generate 48 gate-drive signals while monitoring the status of the gate drives receiving a discrete signal for each output phase.

Additionally, each controller has 20 independent 14-bit analog-to-digital converters that are synchronously triggered to acquire all the electrical measurements (specifically one current sensor per phase and two voltage sensors per NPC). Furthermore, each control system is capable of monitoring 24 multiplexed temperature measurements. Finally, all the reference signals for each converter are sent via a PC using a USB interface.

### IV. Control Design

The overall control concept for the generator is based on a field-oriented scheme [39], assuming the direct axis of the phase current aligned with the permanent magnet rotor flux.

Considering the modular (parallel and series) converter structure under investigation, the control system presents several challenges. First, for each dc channel, the converter has two voltage sources connected in parallel, represented by the series-connected NPC branches. Thus, it must be capable of controlling the instantaneous voltage generated by the branches to the same...
value in order to avoid undesired circulating currents through the converter. Second, within each NPC branch, the converter must generate equal dc voltages on the series-connected NPCs.

Considering these challenges, the two dc-link channels (BRAVO or ECHO) have been with the control scheme, as shown in Fig. 7. Each channel can be operated in different modes, such as speed control, torque control, and dc-link voltage control. Considering the latter, the outer control loop aims to track the reference set point for the total dc voltage of the respective channel. Similarly, when in speed mode, the outermost control loops control the speed of the shaft. The output of the outermost loop (either speed or dc-link voltage) defines the reference value of the global quadrature current component, \( I_{q,\text{ref}} \). Since the application considered is in aerospace, it is vital to reduce the converter volume and weight. This means that the capacitor value must be minimized, making it challenging to passively balance the dc voltage between the series-connected NPCs and within the single NPC. This is achieved by an active dc voltage balancing algorithm, which modifies \( I_{q,\text{ref}} \) in order to maintain the dc voltages of the two series-connected NPCs equal. For example, \( V_{DC1} \) and \( V_{DC2} \) in Fig. 7 should be equal. Consequently, the two NPCs of each series connection have different current references, namely \( I_{q1}^* \) and \( I_{q2}^* \), respectively. Finally, a current control is implemented for each NPC cell. Furthermore, the capacitor voltage balance within the single NPC is integrated into the modulation algorithm, which generates the required switching signals and, at the same time, maintains the voltage on the two inner capacitors of each NPC equal (\( C_1 \) and \( C_2 \) in Fig. 3). This algorithm employs a zero-sequence voltage injection depending on the direction of the phase currents and the reference voltage of each phase. This algorithm aims to maintain the voltage of the two inner capacitors of each NPC equal, for example, \( C_1 \) and \( C_2 \) in Fig. 3 [40]. Fig. 8 shows the structure of the outer loops. If the system is operating in motoring mode, the speed control can be implemented as in Fig. 8(a).

A proportional–integral (PI) controller generates a total torque reference, which is then divided by 4 to define a torque reference for each NPC. After being limited, the torque reference is then converted to a quadrature current setpoint by simply implementing (1), where \( k_t \) [N·m/A] is the machine torque constant. This reference current \( I_{q,\text{ref}} \) is the input to the current control loop, represented by the schematic diagram of Fig. 9.

\[
I_{q,\text{ref}} = \frac{T}{k_t}. \tag{1}
\]

When the system is operating in generating mode, a dc voltage control is implemented, as illustrated in Fig. 8(b). In this case, the PI controller generates a dc current reference, which is then saturated and converted into a torque reference \( T_{\text{ref}} \) by using

\[
T_{\text{ref}} = I_{\text{DC,ref}} \frac{V_{\text{DC}}}{\omega_m}. \tag{2}
\]

Once the torque reference has been defined, the control algorithms follow the same structure, already described for the speed control.
Fig. 9(a) shows the control structure implemented for the dc voltage balancing control of the series-connected NPCs. Based on the difference between the capacitor voltages $V_{DC1}$ and $V_{DC2}$, a PI controller generates an additional current reference $I_{q,bal}$.

This is then summed/subtracted to/from the quadrature current setpoint of each NPC ($I_{q1}^*$ and $I_{q2}^*$) to compensate for the voltage unbalance

$$I_{q1}^* + I_{q2}^* = (I_{q,ref} + I_{q,bal}) + (I_{q,ref} - I_{q,bal}) = 2I_{q,ref}.$$  

(3)

It is important to highlight that this balancing control does not affect the dynamics of the outer loop while maintaining the dc voltages balanced. A similar technique has already been applied in other series-connected multilevel converters, such as CHB converters [41].

Finally, a standard $d-q$ current control with feedforward compensation terms is implemented for each NPC converter, as shown in Fig. 9(b). In terms of modulation, a naturally sampled pulsewidth modulation strategy is used for each NPC converter in this application primarily for the benefits it can provide in terms of ease of implementation and computational burden on the control hardware. Additionally, a capacitor voltage balancing strategy is implemented within the modulation algorithm [40].

V. POWER LOSSES AND EFFICIENCY OF THE POWER CONVERTER

A PLECS model has been implemented for the computation of the converter losses. The simulation calculates the conduction and switching losses as a function of the instantaneous current, voltage, and junction temperature of each of the switching devices. Assuming the nominal operation for the generation unit is at 15 000 r/min with a dc-link voltage of 3 kV, a total power of 4 MW, a PF of 0.8, and a switching frequency of 15 kHz, the simulated total losses of the converter are $\sim$58.5 kW (27.5 kW conduction and 31 kW switching) leading to a converter efficiency of 98.5%.

The distribution of the losses within a power module is displayed in Fig. 10. Although the devices with the highest losses are the IGBTs 2, 3 and the diodes 1, 4, the different thermal impedances mean that diodes 1 and 4 reach the highest junction temperature ($T_j$). Their temperature reaches 130 °C, assuming a cold plate surface temperature of 100 °C. If the same converter is used to drive a machine (as a motor), at the same power, the efficiency will be slightly lower 98.4% with the total losses around 64 kW.

In this case, the distribution losses within a power module will be different; in this case, IGBTs 1 and 4 will be the devices with the highest loss and highest junction temperature. Considering the same cold plate surface temperature of 100 °C, these IGBTs reach a junction temperature of 125 °C.

VI. EXPERIMENTAL RESULTS

To validate the design of the converter, because of the limited dc-link voltage available (1.5 kV instead of 3 kV), at the current test facility, the system has been tested using the arrangements, as shown in Fig. 11. The machine is not coupled to a load drive and both channels of the converter (BRAVO and ECHO) are connected to the same dc power supply. The power supply is capable of maintaining 1.5 kV, instead of the nominal value of 3 kV, and tests are performed using a recirculating power operation of the system [42]–[44]. In Fig. 11, power is recirculated by having the BRAVO converter in speed mode and ECHO in torque (or current) mode. In particular, BRAVO is motoring and ECHO is generating. With this mode of operation, the dc power supply only provides the system losses.

In Section VII, several experimental measurements are presented; in Figs. 12–14, all data are recorded by the DSPs at a sampling rate of 66.6 $\mu$s.

Fig. 12 shows the reference and measured speeds at the shaft from a 1 MW test, where ECHO is controlling 300 A (peak value). The oscillation is equal to 0.026% of the nominal speed value. This fluctuation does not affect the control operation and
Fig. 13. Bravo converter: capacitor voltages (375 V), dc links’ NPCs (750 V), and main dc-link voltages (1500 V).

Fig. 14. Reference (Ref) and measured (Mea) $i_q$ and $i_d$ currents of NPC4 from both BRAVO and ECHO converters.

it can be neglected in practical operation. Due to the dc power supply voltage limit, the maximum speed of the machine is set to 7500 r/min. Fig. 13 illustrates the single capacitor voltages (375 V), series capacitor voltages (750 V), and dc-link voltages (1500 V) for the BRAVO channel. This is equivalent for ECHO. During the tests, the main dc link was controlled by the external dc power supply. This is responsible for the dc-link oscillation, which is also reflected onto the dc-link voltage of the NPCs. The control for series voltage balance is working as designed and keeps the dc voltage of all the NPCs at the same level within a small tolerance. Fig. 13 also highlights the hysteresis voltage control for the internal capacitors of each NPC. The voltage oscillation is higher than expected at rated operation because the lower voltage for this test (half of the nominal value) results in much lower energy stored in the capacitors.

In Fig. 14, the transformed direct and quadrature currents, of the fourth NPC of both BRAVO and ECHO, are shown. While BRAVO is controlling approximately $+300$ A along the $q$-axis, ECHO is keeping the speed of the shaft at 7500 r/min by maintaining approximately $-300$ A along its $q$-axis. The low-frequency ripple of the $I_q$ currents is a consequence of the series dc voltage balancing algorithm, which is implemented to equalize the dc voltages of the series-connected NPCs. Fig. 15 shows the measurements of the four-phase currents while controlling 1 MW of recirculating power.

Two phases ($B$ and $C$) of each channel, BRAVO and ECHO, are shown. In the figure, a noticeable phase shift between the currents of the two systems is present. In fact, since one system is in monitoring mode, while the other is in generation mode, a magnetic phase difference of $30^\circ$ is present due to the winding distribution of the generator (see Appendix A).

The experimental results achieved with the prototype converter and its related control algorithm agree with the expected design values. During the high-power tests, carried out in recirculating mode, the resulting drive system control was able to follow the reference. The obtained results can be considered as a first validation of the proposed high-power and high-voltage multilevel and multiphase drive architecture.

During the same tests, with reference to Fig. 11, it was possible to assess the efficiency of the drive with the two channels (BRAVO and ECHO) in a “series” configuration. The power flows have been measured by acquiring the dc-link voltage and the two direct current (IDC) currents of BRAVO and ECHO, respectively, and feeding them to a power analyzer. Finally, the evaluation of the total losses has been carried out by measuring the output current of the power supply. The last measurement has also been useful to confirm the accuracy of the acquired data, as the sum of the three currents (or powers) flowing to the dc-link connection point must be zero.

Fig. 16 shows the comparison of the efficiency resulting from the simulations and experimental tests during recirculating power operation. Using a complete system model, which
is including the machine, the validation has been performed at 7500 r/min, a dc-link voltage of 1.5 kV, and a reference torque for the ECHO channel corresponding to the \( q \)-axis current \((I_{q_{\text{ref/ECHO}}})\) varying from 100 to 500 A pk with the steps of 100 A pk. Analyzing the highest power test, the total losses of the series drive system are about 85 kW compared with the 84.3 kW obtained from the simulations. The split of the losses according to the simulation results is 53.9 kW (64.7% of the total) in the machine. The remaining losses are split in the ratio 15.8 kW for BRAVO and 14.6 kW for ECHO, i.e., the motoring and generating sides, respectively. The losses in BRAVO are higher because, as mention in Section V, the efficiency is lower when motoring.

In order to show the remaining converter mode, an additional test is reported. The generator is connected to a prime mover operating at its nominal speed of 3000 r/min. The bidirectional dc power supply connected to the converter is controlled in current mode and operates as an electronic load. The voltage reference for the converter is 1.5 kV, and the current reference for the dc power supply is set to its maximum value of 160 A, resulting in a total power of approximately 256 kW.

Fig. 17 shows the BRAVO voltages: the two dc-link voltages of the two layers, all voltages of the NPCs, all internal voltage of the NPC, and the current of the dc link.

VII. CONCLUSION

In this article, the research and development of a prototype 4 MW (5 MVA) medium-voltage active rectifier (with two dc links at 3 kV) to interface to a multiwinding three-phase high-speed generator had been described. A generator with eight independent three-phase windings had been considered. The converter system adopted had been described and the development of the architecture concept to meet the topological requirements of the electrical machine and maximize the reliability of the system had been explained. The resulting structure used NPC building blocks in cascade to achieve high-voltage operation with relatively low-voltage and high-speed devices. In this way, the high-frequency requirement imposed by the high-speed generator can be met. An experimental rig had been setup to perform a scaled-down test up to 1 MW, adopting a recirculating power loop. The experimental results were in line with the expected performance and they validated the initial design and modeling. The outcomes of this work represented a step forward in terms of power density, voltage, power rating, and frequency demand with respect to the state-of-the-art and demonstrated the feasibility of a high-power generator interface suitable for future high-speed, multi-megawatt, and medium-voltage series hybrid-electric aerospace systems.

APPENDIX A

GENERATOR PARAMETERS

The generator features a permanent magnet machine whose parameters are listed in Table III.

The machine has two sets of multiphase windings, one for each channel of the converter (BRAVO and ECHO). The two sets of windings are shifted by 30 electrical degrees. This phase shift is considered as an offset for the measurement of the rotor position in the control algorithm that manages the ECHO channel. Also, due to manufacturing aspects, the terminals of the winding belonging to the ECHO channel are placed on the opposite end side of the electrical machine (i.e., shifted by additional 180 electrical degrees). This results in an equivalent shift of 150 electrical degrees between the two windings. This also explains the results, as shown in Fig. 15, where the opposite reference for the \( q \)-axis currents of the two channels has a phase shift of 30 electrical degrees (180–150).

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Andrew Trentin received the "Laurea" master's and Ph.D. degrees in electrical engineering from the University of Bologna, Bologna, Italy, in 2001 and 2005, respectively.

Since 2005, he has been a Research Fellow with Power Electronics, Machines, and Control Research Group, The University of Nottingham, Nottingham, U.K., and promoted to Senior Research Fellow in 2012. His research interests are power electronics and electrical drives for different applications and in direct ac/ac matrix converters.

Giacomo Sala (Member, IEEE) received the B.Sc. degree in power engineering, the M.Sc.(Hons.) degree in electrical engineering, and the Ph.D. degree in electrical machines and drives from the University of Bologna, Bologna, Italy, in 2012, 2014, and 2018, respectively.

He has been a Research Associate/Fellow until 2019 with Power Electronics, Machines, and Control Group, Department of Electrical and Electronic Engineering, The University of Nottingham, Nottingham, U.K. In 2019, he joined the Department of Electrical, Electronic, and Information Engineering “G. Marconi,” University of Bologna, where he is currently an Assistant Professor. His research interests include design, modeling, and control of multiphase electrical machines, fault-tolerant controls, and fault diagnosis of multiphase drives.

Luca Tarisciotti (Member, IEEE) received the master’s degree in electronic engineering from the University of Rome “Tor Vergata,” Rome, Italy, and the Ph.D. degree in electrical and electronic engineering from the PEMC Group, The University of Nottingham, Nottingham, U.K., in 2009 and 2015, respectively.

In 2015, he became a Research Fellow with the University of Nottingham, U.K., until 2018. He is currently an Assistant Professor with the University Andres Bello, Santiago, Chile. His research interests include matrix converters, dc/dc converters, multilevel converters, advanced modulation schemes, and advanced power electronics converter control.

Alessandro Galassini (Member, IEEE) received the master's degree in mechatronics engineering from the University of Modena and Reggio Emilia, Reggio Emilia, Italy, and the Ph.D. degree in power sharing for multi-phase electrical machines from the University of Nottingham, Nottingham, U.K., in 2012 and 2017, respectively.

He is currently a Research Fellow with the Power Electronics, Machines, and Control Group, The University of Nottingham. His research interests include control of electrical drives for future transportation systems.

Michele Degano (Member, IEEE) received the master's degree in electrical engineering from the University of Trieste, Trieste, Italy, and the Ph.D. degree in industrial engineering from the University of Padua, Padua, Italy, in 2011 and 2015, respectively.

Between 2014 and 2016, he was a Postdoctoral Researcher with The University of Nottingham, U.K., where he joined the Power Electronics, Machines, and Control Research Group and appointed as an Assistant Professor in advanced electrical machines in 2016. He was promoted an Associate Professor in 2020. He is currently the PEMC Director of Industrial Liaison leading research projects for the development of hybrid electric aerospace platforms and electric transports. His main research focuses on electrical machines and drives for industrial, automotive, railway, and aerospace applications, ranging from small-to-large power.

Peter H. Connor received the M.Eng. and Ph.D. degrees in mechanical engineering from the Department of Mechanical, Materials, and Manufacturing Engineering, The University of Nottingham, Nottingham, U.K., in 2009 and 2014, respectively.

He is a Senior Research Fellow with the Power Electronics, Machines, and Control Research Group, Faculty of Engineering, The University of Nottingham. His research interests are mechanical design and thermal management of electrical machines for industrial power generation and high-speed, high power-density traction, and aerospace applications.

Dmitry Golovanov (Member, IEEE) received the Ph.D. degree in superconducting electrical machines from Moscow Aviation Institute, Moscow, Russia, in 2011. He has an experience of working in the industry as a Researcher in VNIIEEM Corporation JSC, Russia, in the field of design of electrical machines and in Samsung SDI, South Korea, in the field of Li-ion batteries. He is currently a Research Fellow with The University of Nottingham, Nottingham, U.K. His main research interests include high power density electric machines for aerospace and automotive industry application, and superconducting electrical machines.

David Gerada (Member, IEEE) received the Ph.D. degree in high-speed electrical machines from The University of Nottingham, Nottingham, U.K., in 2012.

From 2007 to 2016, he was with R&D Department, Cummins, Stamford, U.K., first as an Electromagnetic Design Engineer from 2007 to 2012, and then as a Senior Electromagnetic Design Engineer and Innovation Leader from 2012 to 2016. At Cummins, he pioneered the design and development of high-speed electric machines, transforming a challenging technology into a reliable one suitable for the transportation market while establishing industry-wide used metrics for such machinery. In 2016, he joined The University of Nottingham, where he is currently a Principal Research Fellow, responsible for developing state-of-the-art electrical machines for future transportation that push existing technology boundaries while propelling the new technologies to higher technology readiness levels.

Dr. Gerada is a Chartered Engineer in the U.K. and a member of the Institution of Engineering and Technology.
Zeyuan Xu (Member, IEEE) received the Ph.D. degree in mechanical engineering from the University of Manchester, Manchester, U.K., in 2002. He was subsequently a Research Fellow with the University of Manchester Institute of Science and Technology, Brunel University, and the University of Nottingham. He is currently a Senior Research Fellow in the thermomechanical design of high-speed electrical machines within PEMC Group, The University of Nottingham, Nottingham, U.K. His main research interests include turbulent thermofluid flow, heat transfer enhancement, thermal management of advanced electrical machines and power electronics, electrical machine structure analysis, rotor dynamics analysis, and mechanical design.

Antonino La Rocca received the Ph.D. degree in mechanical engineering from the Department of Mechanical, Materials, and Manufacturing Engineering, The University of Nottingham, Nottingham, U.K., in 2016. He is a Research Fellow with Fluids and Thermal Engineering Research Group and Power Electronics, Machines, and Control Research Group, Faculty of Engineering, The University of Nottingham. His research field is the thermo-mechanical modeling and design of high-speed and high-power dense electrical machines and power electronics designs for advanced generation and propulsion systems by the use of lumped parameters thermal networks and computational fluid dynamics and finite-element analysis.

Carol N. Eastwick received the B.Eng. degree in mechanical engineering from Imperial College London, London, U.K., in 1990, the master’s degree in electrical engineering, and the Ph.D. degree in mechanical engineering from The University of Nottingham, Nottingham, U.K., in 1991 and 1995, respectively. She is a Professor with the University of Nottingham and the Head of the Gas Turbine Transmissions Research Centre. She worked on the modeling and experimental investigations of thermofluids associated with rotating machinery for over 20 years. Her research interest focuses on two-phase fluid flows associated with solid combustion and cooling/thermal management.

Stephen J. Pickering received the B.Sc. and Ph.D. degrees in mechanical engineering from the University of Nottingham, Nottingham, U.K., in 1979 and 1984, respectively. Following several years in the industry, he was appointed as a Lecturer in 1988 with the University of Nottingham, where he currently holds the Hives Chair in mechanical engineering. He has extensive research interests in the area of thermofluids and has undertaken research into the thermal management of electric machines and power electronics’ systems for over 25 years.

Patrick Wheeler (Senior Member, IEEE) received the B.Eng.(Hons.) degree and the Ph.D. degree in electrical engineering for his work on matrix converters both from the University of Bristol, Bristol, U.K., in 1990 and 1994, respectively. In 1993, he moved to the University of Nottingham and worked as a Research Assistant with the Department of Electrical and Electronic Engineering. In 1996, he became a Lecturer with Power Electronics, Machines, and Control Research Group, The University of Nottingham, Nottingham, U.K. Since January 2008, he has been a Full Professor with the same research group. He has authored or coauthored 750 academic publications in leading international conferences and journals. Prof. Wheeler was the Head of the Department of Electrical and Electronic Engineering, The University of Nottingham, from 2015 to 2018. He is currently the Head of the Power Electronics, Machines, and Control Research Group, Global Director of the University of Nottingham’s Institute of Aerospace Technology, and the Li Dak Sum Chair Professor of electrical and aerospace engineering. He is a member of the IEEE PELS AdCom and was an IEEE PELS Distinguished Lecturer from 2013 to 2017. He was involved in the writing of the rules for TTXGP, the first electric superbike Grandprix, in 2009. Since then, he has been involved in the regulations for electric superbike racing as well as founding the University of Nottingham’s Electric Superbike team, which has finished on the podium in three out of the last four years as well as being the European Champions two years in a row. He has also been involved in solar cars, initially as the international observer for the inaugural Chilean Solar Challenge in 2016.

Jon C. Clare (Senior Member, IEEE) was born in Bristol, U.K., in 1957. He received the B.Sc. and Ph.D. degrees in electrical engineering from the University of Bristol, Bristol, U.K., in 1979 and 1990, respectively. From 1984 to 1990, he was a Research Assistant and a Lecturer with the University of Bristol, where he was involved in teaching and research on power electronic systems. Since 1990, he has been with the Faculty of Engineering, The University of Nottingham, Nottingham, U.K. He is currently a Professor of Power Electronics and is the Head of the Department of Electrical and Electronic Engineering. He is a Member of Power Electronics, Machines, and Control Research Group, Nottingham. His research interests are in power-electronic converters and their applications and control. Dr. Clare is a Fellow of the IET and the recipient of a Royal Society Wolfson Research Merit Award.

Chris Gerada (Senior Member, IEEE) received the Ph.D. degree in numerical modeling of electrical machines from The University of Nottingham, Nottingham, U.K., in 2005. He is an Associate Pro-Vice-Chancellor for Industrial Strategy and Impact and Professor of electrical machines. He subsequently worked as a Researcher with The University of Nottingham on high-performance electrical drives and on the design and modeling of electromagnetic actuators for aerospace applications. In 2008, he was appointed as a Lecturer in Electrical Machines; in 2011, as an Associate Professor; and in 2013, as a Professor with The University of Nottingham. His principal research interest lies in electromagnetic energy conversion in electrical machines and drives, focusing mainly on transport electrification. He has secured over £20M of funding through major industrial, European, and U.K. grants and authored more than 350 refereed publications.

Dr. Gerada was a recipient of the Research Chair from the Royal Academy of Engineering in 2013. He served as an Associate Editor for the IEEE TRANSACTIONS ON INDUSTRY APPLICATIONS and is the Past Chair of the IEEE IES Electrical Machines Committee.