Abstract—High performance power conversion equipment is currently gaining an increasing interest for aircraft applications. In particular, isolated bidirectional DC/DC converters are often proposed for modern aircraft distribution systems. A current fed isolated DC/DC converter, named Active Clamp Active Bridge topology, is identified as the most promising for the proposed application, interfacing a 270V DC network with a 28V DC network. A comparison between the selected topology and the well-known Dual Active Bridge topology has been carried out and an experimental prototype has been manufactured for the selected conversion architecture. Simulation and experimental results are provided in order to validate the trade off and the design of the proposed converter.

Keywords—DC/DC power conversion, Power Electronics, Dual Active Bridge, Active Clamp Active Bridge, More Electric Aircraft

I. INTRODUCTION

In recent years, the More Electric Aircraft (MEA) concept has gained an increased importance [1]. In fact, electrical systems are replacing hydraulic, mechanical, or pneumatic power sources in a wide range of aerospace applications [2]. This increase in electrical energy demand has led to a rapid technology development, particularly in power electronics [3]. Electrical systems are now considered for aircraft actuation systems, wing ice protection systems, environmental control systems, and fuel pumping. These novel electrical systems aim to increase future aircraft efficiency, thus reducing the environmental impact of such systems and their maintenance cost. However, due to the electrical system complexity, future aircraft will face similar issues to the one found in ground based microgrids [1]–[3]. In such scenario, several structures for future aerospace microgrids have been proposed. Both AC and DC grids are investigated showing a trend in increasing the voltage level in such grids [4], [5]. Focusing on High Voltage Direct Current (HVDC) aerospace microgrids, Fig. 1 shows a typical grid structure which comprises a Low Voltage Direct Current (LVDC) network and a HVDC network. On the LVDC network typically Low Voltage (LV) emergency batteries and other energy storage elements, such as Electro-Mechanical Batteries (EMB) are connected together with the Electrical Power Unit (EPU) which supplies avionics circuitry [6]. On the HVDC network, High Voltage (HV) generators [7], [8] provide the required voltage and power, together with HV batteries [6] and other EPU [9], which supplies electric actuators, de-icing systems and other MEA equipment [10]–[12]. Moreover, Starter Generator (SG) systems [13], [14], together with their Electrical Control Unit (ECU) are able to provide the required power to start the aircraft engines, and generate additional power during flight, on the HVDC or LVDC network.

As shown in Fig.1, an isolated, bidirectional DC/DC converter has to be included in the system in order to provide an active interface between the two DC networks [15], [16]. The converter must be designed to provide high power density and high efficiency, to be integrated in the aircraft structure with minimum impact on volume, weight and heat management. At the same time, it must provide reliable and flexible operation, also capable to respond to fault events or other abnormal operating conditions with a controlled and predictable behaviour, while minimising the impact on power generation. In particular, the reliability of the DC-DC converter is crucial during “emergency flight” operation, as it will supply the essential equipment to guarantee flight safety. The converter will also operate in a stringent and harsh environment, including limited cooling capabilities; pressure drops, mechanical vibrations and possible EMC or lightning interference, and their operation should have minimum effect on the environment. In order to achieve these challenging requirements, it is clear that cutting-edge technological innovations and innovative design approaches will be required. Several isolated DC/DC converter topologies are proposed in literature. Flyback, Forward, and Forward-Flyback converters are investigated in [17], [18]. These topologies take advantage of resonant networks in order to achieve Zero Voltage Switching (ZVS) or Zero Current Switching (ZCS) commutations, providing highly efficient power conversion. However, additional passive components are
needed and control becomes challenging when a wide operating area is required by the application. On the other hand, Dual-Active-Bridge (DAB) [20] and Three-Phase Dual-Active-Bridge [21] present lower loss switching loss due to their natural capability of achieving ZVS commutation when the devices are turned on, in a wide operating range and without drastically increase the control system complexity. In unidirectional applications, Synchronous rectification also represents an high efficiency alternative to DAB [22]. However, it is not applicable to this work since bidirectional power flow is required. In alternative to DAB a current fed solution has been proposed in previous work, named Active Bridge Active Clamp (ABAC) converter [23]. In fact, the ABAC converter represents a promising solution for MEA applications, due to its inherent current control capabilities. In the following sections, the ABAC converter is introduced and compared with the DAB in terms of design, operating capabilities and weight and volume considerations. Simulations results are provided in several operating conditions in order to assess the benefits of ABAC converter. Experimental results are also provided on a 10kW prototype.

II. THE DUAL ACTIVE BRIDGE CONVERTER

The DAB converter [24], [25], shown in Fig. 2 is a well-known topology, which provides efficient DC/DC conversion as well as galvanic insulation.

![Fig. 2. The Dual Active Bridge (DAB) topology.](image)

As shown in Fig. 2, the DAB consists of two H-Bridge linked on their AC sides with a series inductor and an HF transformer. Conventional Single Phase Shift (SPS) modulation scheme [26] is adopted. With SPS the duty cycle of each bridge/arm is kept constant at 50% of the sampling period, while the phase shift \( \phi \) between the transformer primary and secondary voltages is used to control the power transfer. Theoretical operation waveforms for such topology is drawn in Fig. 3, where \( V_p \) and \( V_s \) are transformer primary current, \( i_{p} \) and \( i_{s} \) are the currents flowing before and after being filtered by the output capacitor \( C_{LV} \). It is noted that the output current \( i_{LV} \) has considerable large ripple which needs a significantly large capacitive filter. The power transfer inductance \( L_{ex} \) on the primary side is designed in order to achieve the required power rating for the converter [20],

\[
L_{ex} = \frac{NV_{HV}V_{LV}}{8f_{s}P_{rated}} \tag{1}
\]

where, \( N \) is transformer turn ratio, \( V_{HV} \) and \( V_{LV} \) are high DC bus voltage and low DC bus voltage respectively, \( f_{s} \) is switching frequency and \( P_{rated} \) is the rated power. In order to design the filter capacitors, \( C_{ex} \) and \( C_{LV} \) in Fig. 3 can be calculated through following equations [20],

\[
I_{p1} = \frac{1}{4f_{s}L_{ex}} (2V_{HV} \frac{\phi}{\pi} - V_{HV} + NV_{LV}) \tag{2}
\]

\[
I_{p2} = \frac{1}{4f_{s}L_{ex}} (2NV_{LV} \frac{\phi}{\pi} + V_{HV} - NV_{LV}) \tag{3}
\]

Referring to Fig. 3, the time intervals \( \Delta_1 \) and \( \Delta_2 \) can be calculated using the following equations:

\[
\Delta_1 = I_{p2} - I_{p1} = \frac{I_{p2} - I_{p1}}{V_{HV} - NV_{LV}} L_{ex} \tag{4}
\]

\[
\Delta_2 = \frac{I_{p2} - V_{HV} + NV_{LV}}{NV_{LV}} L_{ex} \tag{5}
\]

Therefore, the peak-to-peak output voltage ripple \( r_{v} \) can be obtained based on following equation:

\[
r_{v} = \frac{NV_{LV}}{V_{LV}} \int_{\Delta_1}^{\Delta_2} (I_{p1} - I_{LV}) dt = Q_{1} + Q_{2} C_{LV} V_{LV} \tag{6}
\]

where,

\[
Q_{1} = \frac{(NL_{p1} - I_{LV}) + (NL_{p2} - I_{LV})}{2} \Delta_1 \tag{7}
\]

\[
Q_{2} = \frac{(NL_{p2} - I_{LV})}{2} \Delta_2 \tag{8}
\]

The output capacitor \( C_{LV} \) can then be derived from (6) as:

\[
C_{LV} = \frac{Q_{1} + Q_{2}}{r_{v} V_{LV}} \tag{9}
\]

A similar approach can be used to calculate the value of \( C_{HV} \).

![Fig. 3. Main waveforms for DAB with SPS modulation](image)

In order to analyse the ZVS operating area for the DAB converter the voltage transfer ratio \( M \) is defined:

\[
M = \frac{V_{LV}}{V_{HV}} \tag{10}
\]

In buck operations, i.e. when the power is transferred from primary to secondary converter side, the secondary phase shift \( \phi \)
is positive and ZVS is achieved when \(p_1 > 0\) and \(p_2 > 0\). Thus, from (2), (3) and (10) the following conditions can be obtained:

\[
\frac{(1 - NM)\pi}{2} < \phi < \frac{(NM - 1)\pi}{2NM} \tag{11}
\]

Similarly in boost operations, i.e. when the power is transferred from secondary to primary converter side, the secondary phase shift \(\phi\) is negative and ZVS is achieved when \(p_1 < 0\) and \(p_2 < 0\), obtaining the following constraint:

\[
\frac{(NM - 1)\pi}{2NM} < \phi < \frac{(1 - NM)\pi}{2NM} \tag{12}
\]

In Fig. 4, the boundaries (11) and (12) are plotted for different values of \(N\) and the nominal operating conditions (\(V_{HV} = 270\) V and \(V_{LV} = 28\) V) are highlighted. Results show that when \(N = 10\), ZVS on switches turn-on can be achieved in almost the entire phase shift range for both primary and secondary switches.

The control scheme adopted with DAB converter is shown in Fig. 5 for both Buck mode, where the power flows from the HV network to a LV load, and Boost mode, where the power flows from the LV network to a HV load. Clearly only voltage control is feasible with DAB converter without adding additional inductors in the circuit. In terms of modulation, conventional Single Phase Shift (SPS) modulation is implemented. In this case the duty cycle of the HV and LV bridge, \(d_{HV}\) and \(d_{LV}\) respectively, are kept fixed at 50% of the sampling interval \(T_s\). Implementation of SPS is straightforward but presents numerous disadvantages, such as large back-and-forth power [27], hard switching [20], [24], dead band effect [26] under light load non-nominal voltages operation condition. In order to overcome these problems, advanced modulations has been proposed in literature, such as Extended Phase Shift (EPS) [28], Dual Phase Shift (DPS) [29], Triple Phase Shift (TPS) [30], Triangular current Manipulation (TRM) and Trapezoidal current Manipulation (TZM) [31].

III. THE ACTIVE BRIDGE ACTIVE CLAMP CONVERTER

The ABAC converter, shown in Fig. 6, features a similar operating behaviour to the DAB. The main difference is that it provides a current-fed LV stage, taking advantage of external output inductors. The power transfer inductor \(L_{ex}\) transfers power from primary to the clamping circuits, while the two output inductors, \(L_1\) and \(L_2\), serves as buffers to transfer energy from the clamping circuits to the LV output. When using SPS modulation, the switches are all operating at 50% duty, and the two clamping circuits are complementarily switched in order to produce a square waveform on the transformer primary. The design of the input capacitor \(C_{HV}\) is identical to the one in DAB.
On the other hand, ZVS constraint for secondary switches turn-on is defined referring to Fig. 7, where the hard turn-off of the switches already took place and, after the dead-time, the complimentary switch on each clamp circuits is turning on. It can be noted that in the case buck operations T5 and T8 are soft switched during turn-on only if the secondary current $i_s$ is bigger than the respective inductor current. Vice-versa, in the case of boost operations, T6 and T7 are soft switched during turn-on only if the secondary current $i_s$ is bigger than the respective inductor current, after inverting its sign.

$$\frac{(NM - 1)\pi}{2NM} < \phi$$  \hspace{1cm} (16)$$
$$\frac{(NM - 1)\pi}{2NM} > \phi$$  \hspace{1cm} (17)$$

Fig. 7. Soft switching conditions for ABAC converter with SPS modulation

From Fig. 8 is possible to note that soft-switching conditions have to be satisfied when $i_s/L_1 = i_s/L_2 = \pm I_{\text{Lmin}}$ and $i_s = I_{\text{L1}}$.

Fig. 8. Main waveforms for ABAC converter with SPS modulation

According to this analysis soft-switching conditions can be rewritten as in (18) and (19) for buck and boost operations, respectively.

$$I_{\text{Lmin}} < I_{s1}$$  \hspace{1cm} (18)$$
$$I_{\text{Lmin}} < I_{s2}$$  \hspace{1cm} (19)$$

The expressions of $I_{\text{Lmin}}$ and $I_s$ are obtained as follows:

$$I_{s} = \frac{N}{4f_L I_{\text{ex}}(2V_{\text{HV}} \phi - V_{\text{HV}} + 2NV_{\text{LV}})}$$  \hspace{1cm} (21)$$
$$I_{s} = \frac{N}{4f_L I_{\text{ex}}(V_{\text{HV}} - 2V_{\text{HV}} \phi + 2NV_{\text{LV}})}$$  \hspace{1cm} (22)$$

Equations (18) to (22) lead to complex expressions that can be simplified if a negligible current ripple $r_1$ is considered for the output inductors, resulting in the soft-switching region expressed by equations (23) and (24) for buck and boost operations, respectively.

$$\sqrt{\frac{1 - NM}{2}} \pi < \phi < \sqrt{\frac{(NM - 1)\pi}{2NM}}$$  \hspace{1cm} (23)$$
$$\sqrt{\frac{(NM - 1)\pi}{2NM}} < \phi < \sqrt{\frac{NM - 1}{2}}$$  \hspace{1cm} (24)$$

According to the condition in (23) and (24), where $r_1 = 0$ is assumed, the ZVS region for ABAC converter is shown in Fig. 9 for various values of N.

Fig. 9. Soft switching region for ABAC when $r_1=0$.

The results in Fig. 9 represent an ideal case where infinite output inductance is considered and, in terms of soft switching capabilities, represents the worst scenario. When a finite value of output inductance is considered, the increased value of $r_1$ widen the soft switching are by adding an offset term to the boundary conditions, as shown in Fig. 10.

Fig. 10. Soft switching region for ABAC for $N=5$ and different values of output inductance.

In this general case the soft-switching region is expressed by equations (25) and (26) for buck and boost operations, respectively, with $L_o = L_1 = L_2$ in Fig. 6.
calculated from equation (13) as follows:

\[
\frac{1 - NM - \frac{L_{ex}}{2NL_{ex}}}{2NM} < \phi < \frac{(NM - 1)\pi}{2NM} 
\]

\[
\frac{(NM - 1)\pi}{2NM} < \phi < \frac{NM - 1 - \frac{L_{ex}}{2NL_{ex}}}{2NM}
\]

From these equations is clear that lower values of \(L_0\) may be desirable. However it should also be considered that lower values of \(L_0\) results in a higher current ripple and, thus, higher AC losses in the output inductors.

The control scheme adopted with ABAC converter is shown in Fig. 11 for both Buck mode and Boost mode. Conversely to DAB converter, with ABAC converter the implementation of a current control loop is straightforward by regulating the LV current on the output inductors. In terms of modulation, conventional Single Phase Shift (SPS) modulation is implemented.

As for DAB converter, the aforementioned limitations of SPS can be avoided by controlling \(d_V\). In fact in ABAC converter the clamp voltages are directly related to \(V_{LV}\) and \(V_S\). By controlling \(d_V\) in order to keep the amplitude of \(V_p\) and \(V_s\) well matched, it is possible to ensure soft switching in a wide input and output voltage operating range [32]. However this will affect the maximum power transfer. The advanced modulation aforementioned for DAB cannot be directly applied to ABAC converter since they will affect the output current interleaving [33]. As an alternative, if a split secondary structure is considered, as shown in Fig. 12 it is possible to restore the applicability of these modulation by independently controlling the phase shift of the two secondaries [34], [35].

\[
I_{ex}^{\text{rms}} = \frac{2F_{p,m}}{N_2^2} = \frac{V_{mV}V_{LV}}{2N_{f}P_{r}}
\]

The structure of Fig. 12 is completely equivalent to a DAB converters in terms of modulation and it is considered as a term of comparison between the two topologies.

IV. SIMULATION RESULTS

Simulations are carried out considering the parameters of Table I in both Buck (when power is transferred from an HV source to a LV load) and Boost (when power is transferred from an LV source to a HV load) modes.

TABLE I. DESIGN PARAMETERS FOR THE TOPOLOGIES UNDER EVALUATION

| Description        | Symbol | DAB (Fig. 6) | ABAC (Fig. 12) |
|--------------------|--------|--------------|----------------|
| Switching Frequency| \(f_{sw}\) | 100kHz | 100kHz |
| Sampling Time      | \(T_s\) | 10\(\mu\)s | 10\(\mu\)s |
| Rated Output Power | \(P_{out}\) | 8.4kW | 8.4kW |
| Rated Input Voltage| \(V_{mV}\) | 270V | 270V |
| Rated Output Voltage| \(V_{LV}\) | 28V | 28V |
| Power Transfer inductance| \(L_{ex}\) | 7.56\(\mu\)H | 608.4\(\mu\)H |
| HV Filter Capacitance| \(C_{HV}\) | 20\(\mu\)F | 20\(\mu\)F |
| LV Filter Capacitance| \(C_{LV}\) | 3\(\mu\)F | 20\(\mu\)F |
| Clamp Capacitance  | \(C_{clamp}\) | / | 30\(\mu\)F |
| Output Inductance  | \(L_{o1}, L_{o2}\) | / | 3\(\mu\)H |
| Output Inductors DC resistance| \(R_{bus}\) | / | 2.56m\(\Omega\) |
| HF Transformer primary resistance| \(R_p\) | 20\(\Omega\) | 10\(\Omega\) |
| HF Transformer secondary resistance| \(R_s\) | 0.5m\(\Omega\) | 1m\(\Omega\) |

Fig. 13 show steady state operations in Buck mode, when the DAB and ABAC converters are providing 8.4kW on a LV resistive load. Results show similar input/output performance of the two topologies under investigation. As it can be noted, the DAB presents a considerable current ripple on the LV output, which has to be filtered through the LV capacitor, while ABAC present a continuous LV output current with minimal ripple. Fig. 14 show steady state operations in Boost mode, when the DAB and ABAC converters are providing 6kW on a LV resistive load with the same HV current ripple considerations already discussed for buck mode. Fig. 15 shows the behaviour of the two converters when a short circuit is applied on the LV bus, with a short circuit resistance of 78.4m\(\Omega\), at time 0.1s. In this case the DAB converter presents high current ripple on the LV output with a peak value of 950A which is equal to approximately three times the converter rated current. For this reason additional inductors may be required to actively control the current during faults. On the other hand ABAC converter inherently allows current control and present a continuous LV output current, saturated at control at 400A which presents minimal overshoot. In fact the ABAC converter inherently allow current control, i.e. to modify the control output in order to obtain the desired current, which is limited to 400A. When this value is reached, the control output is no longer modified. An efficiency comparison is also shown in Fig. 15. The power electronics devices has been selected accordingly with the required voltage and current rating and paralleled in order to increase the converter current rating or efficiency. In particular, the IPT02N10N3 Silicon Mosfet, rated 100V, 300A has been
selected for the converter LV side, while the C2M0025120D Silicon Carbide Mosfet, rated 1.2kV, 90A has been selected for the high voltage side. In order to perform a fair comparison the same number of devices in parallel has been selected for the two converters (one on the HV side and four on the LV side) with the only difference that for the ABAC converter these devices are equally divided between the two secondary circuits. Regarding the passive components, capacitors losses are considered negligible while only the DC losses are considered for the magnetics components. In fact, the AC losses can be minimized by performing an accurate planar magnetics design. The efficiency is estimated through PLECS simulations. The results, shown in Fig. 15, highlight the high efficiency for these two topologies when the soft switching constraints are satisfied. When the soft switching is lost as for example in the 300V/21V curve efficiency drops drastically due to increased switching losses. Advanced modulations can be implemented to increase the soft switching region. It is important to highlight that this efficiency analysis is carried out with the solely scope of comparing the two converters and do not represent an accurate measurement of efficiency, since additional losses in the magnetics are not considered and switches thermal models are obtained using datasheet parameters.

V. WEIGHT AND VOLUME CONSIDERATIONS

A first estimation of weight and volume for the three topologies has been calculated by considering off the shelf components and their datasheet information, shown in Table II, combined with the design parameters. The weight and volume estimation does not include heatsink weight and the additional weight and volume of PCB boards, wiring and other components. The design is carried out at three different power rating, 5kW, 10kW and 20kW respectively, in order to identify trends in the components volume and weight. The results in Fig. 16, show that ABAC converter has more weight and volume associated with magnetic components while the DAB converter has an higher capacitor weight and volume.

| Components         | Description                              | Weight [kg] | Volume [l] |
|--------------------|------------------------------------------|-------------|------------|
| LV Capacitor       | Ceramic Capacitor, C Series, 1 µF, 100V  | 0.0003      | 0.0003     |
| LV Capacitor       | Cornell Dubiler, Film Capacitor, 10µF, 50V | 0.0007      | 0.0013     |
| HV Capacitor       | Film Capacitor, B32774 Series, 10µF, 450V | 0.05        | 0.0116     |
| 5kW Transformer    | STANDEX Series P350 series               | 0.4         | 0.0935     |
| 10kW Transformer   | STANDEX Series P560 series               | 0.7         | 0.2513     |
| 20kW Transformer   | STANDEX Series P900 series               | 1.2         | 0.6533     |
| 5kW Inductor       | Micrometals E100-40 15 12µH@13rms        | 0.0215      | 0.0096     |
| 5kW Inductor       | Micrometals E310-2 1200µH@44rms         | 0.027       | 0.0096     |
| 10kW Inductor      | Micrometals E165-40 7.16µH@16rmas       | 0.0961      | 0.0628     |
| 10kW Inductor      | Micrometals E160-2 600µH@88rms          | 0.1065      | 0.0628     |
| 20kW Inductor      | Micrometals E220-40 3.7µH@72rms         | 0.2147      | 0.1296     |
| 20kW Inductor      | Micrometals E220-2 900µH@76rms          | 0.2223      | 0.1296     |
| Output Inductor    | Coilcraft SR10/0.3 3, 50A, 93.6A        | 0.0364      | 0.0132     |
| HV Device          | CREE C2M0025120D, 904, 1.2 kV            | 0.0011      | 0.0035     |
| LV Device          | INFINEON IPTD200/10SN, 300A, 100 V       | 0.0014      | 0.0003     |

The increased capacitor weight and volume in DAB is mainly related to their demanding current capabilities which requires a combination of film and ceramic capacitors and discourage the use of electrolytic capacitors. On the other hand even if magnetics in the ABAC converter has to handle high currents, their inductance value is relatively low and few turns are required, thus reducing their weight and volume and allowing further optimization of these components. This trend gets more and more evident when increasing the rated power, making the ABAC converter particularly suitable for high power ratings.

Fig. 13. Simulation results: Waveforms in Buck mode operation for DAB and ABAC converter when transferring 8.4kW on a resistive LV load.

![Figure 13: Simulation results: Waveforms in Buck mode operation for DAB and ABAC converter when transferring 8.4kW on a resistive LV load.](image-url)
**Fig. 14.** Simulation results: Waveforms in Boost mode operation for DAB and ABAC converter when transferring 6kW on a resistive HV load.

**Fig. 15.** Simulation results: Waveforms during Buck mode operations when transferring 10kW on a resistive LV load and imposing a short circuit on the load at time 0.1s with a short circuit resistance of 78.4 mΩ; Efficiency estimation in various operating conditions for both DAB and ABAC converter.
VI. EXPERIMENTAL RESULTS

A 8.4kW experimental prototype based on the ABAC converter has been manufactured, as shown in Fig. 17. Experimental results are shown in Fig. 18 and Fig. 19 when the converter is operating in buck and boost mode, respectively.

The results are obtained considering the design parameters of Table I and a power transferred to the load of 3kW and 5kW, in buck and boost mode respectively. Voltages and currents at primary and one secondary of the HF transformer are presented, as well as voltage and current on both converter external interfaces (supply and load). The load voltage is well regulated at 28V, while providing the load current without any noticeable oscillation. On the other hand, looking at the current flowing through one of the four output inductors, it is possible to notice that the triangular ripple, which is cancelled at the load point, matches the value previously obtained in simulations. Experimental results validate the results of simulations except for the inductive voltage drops on the secondary transformer voltage, which are caused from parasitic inductance on the PCB board design.
VII. CONCLUSIONS

In this work a comparison between DAB converter and ABAC converter for aerospace applications has been carried out. The two topologies are compared in terms efficiency, weight and volume for the specific application of a 270V/28V 10kW bidirectional DC/DC converter. The design procedure has been described and soft switching analysis has been performed for both topologies. Accordingly to these analysis the two converter are evaluated through simulations in term to input and output waveforms, control, and efficiency. Additionally a first estimation of weight and volume for the two topologies when the converter power rating vary from 5kW to 20kW is carried out. Finally, experimental results validate the results obtained by simulations. From the obtained results it can be concluded that the ABAC converter represents a promising alternative to the DAB converter, able to reduce the converter weight and volume at high power ratings and inherently control the LV current whilst maintaining similar efficiency of a DAB converter.

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