Comparative evaluation of capacitor-coupled and transformer-coupled dual active bridge converters

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Abstract—Dual Active Bridge (DAB) converters are emerging to become the preferred high power DC-DC conversion topology to satisfy the requirements of modularity, high voltage transfer ratio, high efficiency and bidirectional power transfer capability. However transformer design for DAB converters remains a challenging problem, especially for high voltage conversion ratio and higher switching frequencies. Capacitor-coupled transformer-less DC-DC converters capable of arbitrary high voltage transfer ratio with reasonable efficiency have been recently introduced. The voltage gain in this case is achieved by series combination of capacitor coupled active bridge modules. This paper introduces the capacitor-coupled dual active bridge converter and compares it with the traditional transformer-coupled dual active bridge converter. A brief overview of the analytical models of both transformer-coupled and capacitor-coupled DAB converters is presented and the important design factors of both topologies are identified, followed by a comparative evaluation of a published benchmark design with details of all the power circuit components.

I. INTRODUCTION

All the global economic trends show a continuous drop in the projected price of solar power (especially utility-scale solar) [1] while production-scale photovoltaic panel efficiency is steadily improving [2]. This accompanied by various other trends such as the increase in large-scale computing power represented by cloud computing [3], falling prices of electrical energy storage and large scale introduction of electric vehicles. The net effect is that there is a significant demand for an high power dc power conversion systems. Dc-dc converters have an increasing role to play in this environment. Modularity, high voltage transfer ratio, high efficiency, source isolation, bidirectional power transfer capability are some of the desirable characteristics of dc-dc converters for sensitive load applications like datacenters. Dual active bridge (DAB) converters are one of the most popular converter topologies that satisfy these requirements.

DAB converters (Fig 1) were first introduced by [4] [5] as an alternative to resonant and pseudo-resonant DC-DC topologies for high power applications. The DAB topology eliminated the large output filter inductor present in previous configurations and introduced a smaller inductance in series with the HF transformer. DAB converters consist of two voltage-source full-bridge converters connected by a series inductance and isolated by a high frequency (HF) transformer as shown in Fig 1. The high frequency transformer can be either single phase or three phase depending on the application. A three phase DAB has lower current ripple, slightly higher efficiency and better transformer utilization at the cost of more switching devices [7]. The series inductance can be minimized by operating at sufficient high frequency and utilizing the leakage inductance of the transformer. The presence of active bridges at both primary and secondary windings allows for bidirectional operation with zero voltage switching along the control range defined in [5].

Fig. 1. Schematic of transformer-coupled single-phase Dual Active Bridge converter

The basic operating characteristics of DAB were enumerated by [4]- [6] and the optimal design constraints with respect to efficiency, power density etc. were explored by [8]- [11]. The effect of modulation strategy on the converter efficiency was explored by [10], [11]- [13], where it was shown that the optimal modulation scheme was one with the lowest rms current in the transformer while still maintaining ZVS on the primary and secondary converters.

The most direct approach to increase the power transfer capability of the DAB converter is to operate at higher primary and secondary voltages with the attendant difficulty in transformer insulation and reduced switching frequency [15]. The second approach is a cascaded modular architecture where multiple full bridge converters both on the primary and secondary side can be connected in series or parallel combination to increase the power rating of the converter without increasing the VA rating of the switching devices or passive components [16]. The trade-off here is that control complexity necessitates the use of distributed controller. But

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in both cases, the transformer design is not a trivial problem, especially when the voltage conversion ratio is greater than 1:2 or 1:3 [17]. Transformers in bidirectional converters are also susceptible to saturation, which requires passive or active methods for flux balancing [18]. Even for low power application, higher switching frequencies can cause problems in magnetic components.

Transformer-less switching mode dc-dc converters with high conversion ratio are an old concept, proposed as early as [19] when it was quickly realized that the typical boost converter topology has significant disadvantages when the voltage conversion ratio is greater than 1:3 or 1:4. Many transformer-less converter topologies are modular and any arbitrary conversion ratio can be achieved by cascading appropriate number of stages. A capacitor-coupled transformer-less dc-dc converter topology has been proposed by [20] [21]. This topology consists of a number of branches of capacitor coupled high frequency rectifiers fed in parallel from a high frequency ac source and connected in series at the output. This topology is very similar to a transformer-coupled DAB converter, except that voltage gain here is achieved by series combination of capacitor coupled converter modules as shown in Fig 2. A two-stage implementation of this converter topology has been evaluated by [22]. This paper examines the design equations and compares the dual-active bridge variant of this topology with existing DAB transformer-coupled converter topology.

A brief review and analysis of transformer-coupled and capacitor-coupled DAB converters is presented in Sections II using a unified single state equivalent circuit. The discussion focuses on unidirectional power transfer, although the approach can be extended easily to accommodate such variations. The two important design considerations that define the converter nominal operating conditions are enumerated in Section III. Section IV lists the design equations that are used to present the qualitative comparison of both the topologies.

The final concluding section presents a summary of the paper.

II. ANALYTICAL MODELING

A. Single stage equivalent circuit

The transformer-coupled (TCDAB) and capacitor coupled (CCDAB) converters have an identical full bridge structure at the input terminals of the ac side. In case of TCDAB, the voltage gain from input to output is achieved via the transformer turns ratio. The output stage is structurally similar to the input stage, typically with higher voltage and reduced current, both of which are decided by the turns ratio. The series inductor \( L_s \) which is referred to the output side acts as the intermediary current stiff component to interface between two fixed voltage stiff components \( C_{in} \) and \( C_{out} \). The primary and secondary transformer winding voltages and currents are shown in Fig 3. The discussion here focuses on a step-up application, where the transformer nominal turns ratio in case of TCDAB is defined as \( n \geq \frac{V_{out}}{V_{in}} \).

The input stage of the capacitor coupled DAB is similar to transformer coupled DAB. However, the output voltage is realized through the series connection of several output states at the dc terminals. Each output stage consists of two series capacitors \( C_s \), series inductor \( L_s \), converter full-bridge \( S5-S8 \) and an output filter capacitor \( C_f \) as shown in Fig 2. The inductor current per stage is \( i_{Ln} \). The voltage transformation in case of a CCDAB is realized using multiple output stages. In the case of CCDAB, the number of stages for a given \( V_{in} \) and \( V_{out} \) takes the role of transformation ratio, similarly defined as \( n \geq \frac{V_{out}}{V_{in}} \).

It is convenient to use an equivalent single stage unity turns-ratio converter to represent both CCDAB and TCDAB so that the analysis can be developed in a unified manner. Fig 4 shows the single stage equivalent circuit that is suitable for analysis of both TCDAB as well as CCDAB. The transformer of the TCDAB is modeled simply by its leakage inductance, lumped together with the series inductance of the coupling circuit, while the magnetizing inductance has been assumed to be infinite. On the other hand, the capacitor of CCDAB is assumed to be of negligible reactance, and has been replaced by a short circuit, while the number of stages has been reduced.

![Fig. 2. Schematic of capacitor-coupled single-phase Dual Active Bridge converter](image)

![Fig. 3. Typical voltage and current waveforms of TCDAB on the primary (top frame) and the secondary (bottom frame)](image)
to one. The single stage equivalent circuit has $1/n$ output power of the full TCDAB or CCDAB. There is no change in input voltage, but the input current is reduced by $1/n$. On the output side, the output voltage is $V_{outn}$ while the inductor current is $i_{L,n}$. The results from the analysis of the single stage circuit can be readily ‘unwound’ to extend the results to the n-turn or n-stage converter as required.

### B. Modeling assumptions

The analytical model of the transformer-coupled DAB (TCDAB) is discussed in detail by [6] and [10]. The transformer-coupled (TCDAB) and capacitor coupled (CCDAB) converters can be analyzed by similar framework based on the following assumptions.

- The converter operation during various PWM intervals is assumed to be linear, ie ZVS interval in currents is excluded.
- Phase shift modulation is the preferred modulation scheme.
- The effect of modulation strategy is only to regulate the power flow between fixed voltage sources at the input and output.
- Second order effects like ESR of series inductor are ignored.

Typical waveform of the inductor current is illustrated in Fig 5 over a half cycle of operation. Both TCDAB and CCDAB control the power flow by adjusting the phase difference between the input and output voltage stages. The phase of the input voltage bridge is $\theta_i$ and the phase of (all) the output voltage bridge(s) is $\theta_o$. The phase shift between the input and output bridges is $\phi$. $T_{\phi}$ is the phase shift expressed in units of time. $t_D$ is the time interval where the output voltage of the input or output bridge is positive while the current direction is negative. This is the diode conduction interval.

$$\phi = \theta_i - \theta_o$$

$$T_{\phi} = \frac{\phi}{2\pi F_s}$$

C. Input and output voltage

The ac side voltages across the two ends of the inductor may be expressed as

$$v_i(t) = \begin{cases} +V_{in} & \text{if S1,S4 ON} \\ 0 & \text{if S1,S3 ON; or S2,S4 ON} \\ -V_{in} & \text{if S2,S3 ON} \end{cases}$$

$$v_{on}(t) = \begin{cases} +V_{outn} & \text{if S5,S8 ON} \\ 0 & \text{if S5,S7 ON; or S6,S8 ON} \\ -V_{outn} & \text{if S6,S7 ON} \end{cases}$$

The dc output voltage of the single stage converter may be transformed to that of the real converter for both CCDAB and TCDAB will be

$$V_{out} = nV_{outn}$$

D. Input and output current

The steady state inductor current waveform during the various intervals may be expressed as a piecewise linear segments as

$$i_L(t) = \begin{cases} I_{L0} + \frac{V_{in} + V_{outn}}{L} t & t \leq T_{\phi} \\ I_{L1} + \frac{V_{in}}{L} & T_{\phi} \leq t \leq 0.5T_s \\ -I_{L0} - \frac{V_{outn}}{L} & 0.5T_s \leq t \leq 0.5T_s + T_{\phi} \\ -I_{L1} - \frac{V_{outn}}{L} & 0.5T_s + T_{\phi} \leq t \leq T_s \end{cases}$$

where $I_{L0}$ is the current at the beginning of the switching half-cycle, $I_{L1}$ is the current at time $t = T_{\phi}$ that can be determined as

$$I_{L0} = \frac{\pi(V_{outn} - V_{in}) - 2|\theta_o|V_{outn}}{4\pi F_s L_s}$$

$$I_{L1} = \frac{\pi(V_{outn} - V_{in}) + 2|\theta_o|V_{in}}{4\pi F_s L_s}$$

by relating the boundary conditions and through symmetry. Furthermore, the diode conduction interval may be determined to be

$$t_D = \frac{|I_{L0}|L_s}{V_{in} + V_{outn}}$$
From the solution to the inductor current waveform, its RMS value and peak value may be readily determined. Furthermore, the ac side current at the input bridge terminals for the real CCDAB and TCDAB converter may be determined using

\[ i_i = n i_L \]  \hspace{1cm} (10)

The RMS value and the peak value of the input current may also be determined similarly through a multiplication by a factor of \( n \).

### E. Input and output power

Using the relationships for voltage and current, the average power throughput may be determined to be

\[ P_o = \frac{V_{in} V_{outn} \phi (\pi - |\phi|)}{2\pi^2 F_s L_s} \approx \frac{V_{in} V_{outn}}{2\pi F_s L_s} \phi \]  \hspace{1cm} (11)

Furthermore, the apparent power or kVA throughput of the converter can be defined as

\[ S = V_{in} I_{Lrms} \]  \hspace{1cm} (12)

From \( S \) and \( P \), we can define the reactive power transfer as

\[ Q = \sqrt{S^2 - P_o^2} \]  \hspace{1cm} (13)

### III. DESIGN CONSIDERATIONS

#### A. Effect of transfer ratio

The voltage transfer ratio of the equivalent single stage converter is defined as \( \nu = \frac{V_{outn}}{V_{in}} \). This is different from the transformer voltage transformation ratio \( n \). The transfer ratio is a design factor that has a direct effect on the reactive power transfer of the DAB converter. Fig 6 shows a plot of per-unit real power \( P \), reactive power \( Q \) and apparent power \( S \). \( V_{in} = 12V, P_{out} = 150W, n = 1, \phi = 0.05\pi \) rad.

![Fig. 6. Effect of transfer ratio \( V_{outn}/V_{in} \) on real power \( P \), reactive power \( Q \) and apparent power \( S \).](image)

The result of the preceding discussion is that there are two obvious design elements that lead to optimal realizations of DAB converters. The first feature is that \( V_{in} \approx V_{outn} \), or \( nV_{in} \approx V_{out} \), and the second feature is that \( \phi \approx 0 \). The consequence of these optimal design features is that they may be applied as simplifying approximations in that the analysis of the converter, leading to simple design expressions.

#### B. Effect of \( \Phi \)

As shown in (11), the output power of a DAB converter can be expressed as a function of the displacement phase difference between the input ac voltage \( v_i(t) \) and output ac voltage \( v_o(t) \). Fig 7 shows the effect of this phase difference \( \phi \) on the per-unit real, reactive and apparent power. The figure is per-unitized to the real power transferred at \( \phi/\pi = 0.05 \). For small angles of \( \phi \), it can be observed that \( P \) vs \( \phi \) is linear, and the amount reactive power throughput is less than 1 pu, and is reasonably flat. But if the phase difference is increased more than this small angle \( \phi/\pi > 0.15 \), the per-unit reactive power \( Q \) is more than 1 pu. This figure illustrates that the DAB converter would have reasonable performance when the the operating phase shift cannot be much greater than 0.1 \( \pi \) rad. In reality, this leads to a limit in the dynamic range that is available for providing output regulation in the presence of load changes.

![Fig. 7. Effect of phase shift \( \phi \) on real power \( P \), reactive power \( Q \) and apparent power \( S \). \( V_{in} = 150V, V_{out} = 12V, P_{out} = 150W, n = 14 \).](image)

The result of the preceding discussion is that there are two obvious design elements that lead to optimal realizations of DAB converters. The first feature is that \( V_{in} \approx V_{outn} \), or \( nV_{in} \approx V_{out} \), and the second feature is that \( \phi \approx 0 \). The consequence of these optimal design features is that they may be applied as simplifying approximations in that the analysis of the converter, leading to simple design expressions.

#### C. Grounding and filtering considerations

According to the National Electric Code (NEC), for two-wire dc systems operating between 50-300V one of the conductors must be grounded. For dc systems operating at higher voltages, a three-wire system must be used and the neutral conductor must be grounded. [23] discusses grounding option for both cases - the positive conductor of a two-wire system is grounded via a high resistance, and the neutral conductor of a three-wire system is low resistance grounded. TCDAB has
galvanic isolation where either the input stage or the output stage can be independently grounded at any terminal, making it suitable for systems than can converters that operate in the 50-300V. On the other hand, CCDAB may be grounded at any terminal only if grounding is required only at one of the input terminals or at one of the output terminals. If grounding is required at both input and output ports, CCDAB requires a three wire system at both ports and the neutral may be grounded at both ports. For a CCDAB with \( n \) stages, the system ground is located at the midpoint of the input dc port and the output dc port as shown in Fig 8. Reactance grounding may be achieved to ensure that the connection is solidly grounded for dc and high impedance grounded for all other frequencies. The value of the grounding reactance \( L_g \) is selected such that \( L_g \gg L_s \). This also provides a closed the leakage path for any ac ground currents, through the midpoint of the sx source \( V_{in} \) is similarly grounded with \( L_g \).

![Fig. 8. Reactance grounding configuration for capacitively coupled DAB converter](image)

The capacitive coupling discussed in this paper is implemented with series coupling capacitors which block dc voltages and allow high frequency ac. Obviously, this does not provide any shielding for the common mode noise from being propagated across the converter. This is in contrast to TCDAB where the input stage common mode noise is isolated from the output stage by the isolation boundary across the transformer. Any common mode noise isolation in the CCDAB realization has to be added through explicit design.

**D. Design example**

For a qualitative comparison of both converter topologies, a prototype transformer-coupled and capacitor-coupled DAB converter is designed for the specifications listed in Table I. These specifications are derived from a typical dc-dc converter application for datacenter loads [24].

<table>
<thead>
<tr>
<th>TABLE I</th>
<th>DUAL ACTIVE BRIDGE CONVERTER RATINGS</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>( V_{in} )</td>
</tr>
<tr>
<td></td>
<td>V</td>
</tr>
<tr>
<td>12</td>
<td>12.5</td>
</tr>
</tbody>
</table>

**IV. DESIGN EQUATIONS**

**A. Switches**

1) Input switches: The input switch current is same in case of both TCDAB and CCDAB if the transformer turns ratio of TCDAB is same as the number of stages of CCDAB. The rms current of the switches on the input side can be approximated as

\[
I_{swrms} = n \sqrt{\frac{I_{L0}^2 + |I_{L0}| I_{L1}| + I_{L1}^2 (0.5 T_s - T_\phi)}{3}} \tag{14}
\]

where \( I_L \) is the inductor current on the secondary side in case of TCDAB and inductor current per stage in case of CCDAB.

The average switch current can be similarly defined as

\[
I_{swavg} = n \frac{1}{2 T_s} \left( |I_{L0}|(T_s - 2t_D) + |I_{L1}|(T_s - 2T_\phi) \right) \tag{15}
\]

The diode current on the input side is given as

\[
I_{dirms} = n \sqrt{\frac{I_{L0}^2 t_D}{3 T_s}} \tag{16}
\]

\[
I_{diavg} = \frac{n I_{L0} t_D}{2 T_s} \tag{17}
\]

The voltage ratings of the input switches and diodes are maintained at \( V_{in} \).

2) Output switches: Similar to the input side, the output switch currents will also be derived.

\[
I_{sworms} = \frac{1}{3 T_s} \left( |I_{L0}| T_s + |I_{L1}|(T_s + 2t_D - 2T_\phi) \right) \tag{18}
\]

The one difference is in the maximum reverse blocking voltage on the output side switches in case of CCDAB will be lower than TCDAB. The diode current on the output side is given as

\[
I_{dirms} = \sqrt{\frac{I_{L1}^2 T_\phi - t_D}{3 T_s}} \tag{19}
\]

The voltage ratings of the output switches and diodes are maintained at \( V_{out} \) in the case of TCDAB, and \( V_{out}/n \) in the case of CCDAB.

**B. Voltage transformation**

1) Magnetic coupling: The rms current of the transformer winding on the input side will be approximately same as the switch current on the input side, if we assume that the phase difference is negligible. Similarly, the rms current of the transformer winding on the output side will be same as the switch current on the output side. The rms voltage of the primary winding is \( V_{in} \) and on the secondary winding is \( V_{out} \). The area product is a measure of the power handling capability of the transformer. We can define the area product in terms of the sum of KVA on the primary and secondary winding.

\[
\sum V_A = V_{in} n I_{krms} + V_{out} I_{krms} \tag{22}
\]

\[
A_p = \frac{\sum V_A}{B_{max} F_s \cdot \sqrt{\sum K_f K_a}} \tag{23}
\]
where $K_f$ is the waveform coefficient (4.0 for square wave), $B_{\text{max}}$ is the designed maximum flux density in the core, $J_{\text{max}}$ is the maximum designed current density in the copper winding, $F_s$ is the switching frequency and $K_n$ is the window utilization factor.

2) Capacitive coupling: Capacitors in other dc-dc converters are primarily used for energy storage and accordingly operate with steady dc voltage and small ac ripple voltage. The series capacitor in capacitor-coupled DAB converter operates with large ac currents. The dc voltage across each series capacitor is a function of its position along the string and the output voltage [20]. Ceramic or Film capacitors are ideal for this combination of dc-ac and low energy storage. The capacitance is selected such that the capacitive impedance at switching frequency is negligible. This translates into a design rule,

$$C_s > \frac{50}{R_{\text{out}, f_s}},$$

(24)

The maximum voltage rating of the capacitors are maintained at $V_{\text{out}}/2$, while the RMS current ratings are identical to $I_{\text{Lrms}}$.

C. Series inductor

The series inductor is assumed to be on the secondary side of the transformer for TCDAB, since it leads to smaller currents. The design value for the inductance for rated power transfer in case of TCDAB can be derived from the power equation (11), with approximation for small $\phi = \pi/16$, to be

$$L_s \leq \frac{nV_{\text{in}}V_{\text{out}}}{32F_sP_{\text{out}}},$$

(25)

In the case of CCDAB, one series inductor is present in the series path of each of the stages whose value may be determined to be

$$L_s \leq \frac{V_{\text{in}}V_{\text{out}}}{32F_sP_{\text{out}}},$$

(26)

D. Performance characteristics

The operating performance of both CCDAB and TCDAB can be compared at the same power output in terms of their power loss and hence their efficiency.

1) Input switch loss: Both the input and output switches are operated under ZVS conditions which implies that switching loss is negligible. The only loss in the switches is during conduction. If $R_{\text{dsom}}$ is the drain to source on-state resistance of the switching devices and $V_{f_1}$ is the anode to cathode forward voltage drop of the antiparallel diode, the average loss at the input switch for both CCDAB and TCDAB is given by

$$P_{\text{swi}} = I_{\text{s}}^2R_{\text{dsom}} + I_{\text{davg}}V_{f_1}$$

(27)

2) Output switch loss: Since the reverse blocking voltage in case of CCDAB is lower than TCDAB, the output side switches will be different for the two realizations. $R_{\text{dsom}}$ is the drain to source on-state resistance of the switching devices and $V_{f_2}$ is the anode to cathode forward voltage drop of the antiparallel diode on the output side of CCDAB. $R_{\text{dsom}}$ is the drain to source on-state resistance of the switching devices and $V_{f_2}$ is the anode to cathode forward voltage drop of the antiparallel diode on the output side of TCDAB. The average switch loss in case of CCDAB is

$$P_{\text{swi}} = I_{\text{s}}^2R_{\text{dsom}} + I_{\text{davg}}V_{f_2}$$

(28)

The average switch loss in case of TCDAB is

$$P_{\text{swi}} = I_{\text{s}}^2R_{\text{dsom}} + I_{\text{davg}}V_{f_2}$$

(29)

3) Transformer loss: The sources of loss in the transformer are core loss in the magnetic core and copper loss in the winding. If the winding is designed to have diameter less than the skin depth of copper at the switching frequency, the copper loss can be modeled in terms of DC resistance. The core loss is derived from the datasheet of the ferrite material for the core. If $V_c$ is the volume of core, $B_{\text{max}}$ is the maximum flux density in the core, $K_c, \alpha$ and $\beta$ are constants derived from the datasheet, the core loss is derived as

$$P_{\text{core}} = V_cK_cB_{\text{max}}^\beta$$

(30)

The copper loss is derived as

$$P_{\text{cupri}} = (nI_{\text{Lrms}})^2R_{\text{dcsec}}$$

(31)

$$P_{\text{cupi}} = I_{\text{Lrms}}^2R_{\text{dcsec}}$$

(32)

where $R_{\text{dcpri}}$ and $R_{\text{dcsec}}$ are the primary and secondary copper winding resistance at DC.

4) Series capacitor loss: The power loss in a capacitor depends on the equivalent series resistance of the capacitor. ESR is expressed in terms of the dissipation factor ($\tan \delta$) of the capacitor.

$$X_c = \frac{1}{2\pi F_sC_s}$$

(33)

$$R_c = X_c\tan \delta$$

(34)

$$P_c = I_{\text{Lrms}}^2R_{\text{c}}$$

(35)

5) Series inductor loss: The series inductance can be designed as part of the transformer leakage inductance. If the series inductance is designed as an individual component using off-the-shelf parts, the power loss can be derived from the series equivalent resistance ESR $R_{\text{Ls}}$.

$$P_{\text{Ls}} = I_{\text{Lrms}}^2R_{\text{Ls}}$$

(36)

V. COMPARATIVE EVALUATION

Complete detailed designs of the TCDAB and CCDAB to meet the specifications of the dc-dc converter listed in I were developed to compare their features and verify their performance against the analytical design results. A qualitative and quantitative evaluation of the TCDAB and CCDAB converter topology for similar input, output and rated power conditions are presented further. The active switch ratings are based on the maximum reverse blocking voltage at the input and output stage and the maximum rms current at the operating point. The high frequency inductor for both TCDAB and CCDAB are selected from off-the-shelf ferrite core inductors. For low
values of inductance as in CCDAB, toroidal cores feature the lowest loss. Multilayer ceramic capacitors with the required current and voltage ratings are used to realize the coupling capacitor. The high frequency transformer is custom designed using a ferrite core and foil and litz winding. Details computer simulation of both TCDAB and CCDAB converters were developed using PLECS platform with electrical representation of all the devices including various parasitic properties. Extensive simulations were used to establish the validity of the analytical design models. Fig 9-10 illustrates selected waveforms that indicate the validity of the solutions compared against the analytical predication. The computer simulation was also used to determine the various design quantities such as RMS currents, average currents, switching losses, etc. for the different design components. Main design details of the switching devices and the reactive components including the transformer for both TCDAB and CCDAB is listed in Table III, along with a side-by-side comparison with the numerical design predictions of the analytical design equations presented in the foregoing section.

### A. Volume

Table II illustrates the volume of major power circuit components for both TCDAB and CCDAB. The volume of each component is calculated using the exact mechanical specifications in the datasheets and does not account for the volumetric clearance needed for creepage clearance and layout. From the table, CCDAB is 4× better than TCDAB in terms of volumetric power density. The semiconductor switch volume does not include any gate drive circuits. Such aspects related to the physical realization will affect the final 'box' volume of the power converters and are the subject of continuing investigations.

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</thead>
<tbody>
<tr>
<td>Series Inductor</td>
<td>48</td>
<td>0.1</td>
<td>1022</td>
<td>0.5</td>
<td>N/A</td>
<td>2230</td>
</tr>
<tr>
<td>Transformer</td>
<td>15600</td>
<td>0</td>
<td>N/A</td>
<td>2</td>
<td>38</td>
<td>2</td>
</tr>
<tr>
<td>Series capacitor</td>
<td>N/A</td>
<td>0.2</td>
<td>533</td>
<td>3</td>
<td>3</td>
<td>3</td>
</tr>
<tr>
<td>Input switch</td>
<td>4</td>
<td>0.2</td>
<td>15700</td>
<td>4.3</td>
<td>3800</td>
<td>5.7</td>
</tr>
</tbody>
</table>

**Efficiency**: 97% **96%**

### B. Efficiency

The computer simulation model was also used to determine the loss components across various devices. When comparing active semiconductor devices, it can be observed from the previous section the maximum blocking voltage on the output and input stage depend on the output and input voltage respectively in case of TCDAB. In case of CCDAB, due to the output-parallel topology, all the active semiconductor devices are rated for the same input voltage. Under rated operating conditions, both the input and output bridges operate under ZVS, and hence have no switching losses. This, combined with negligible conduction loss means that the total semiconductor power loss can be dissipated via air convection and copper trace conduction cooling.

### VI. CONCLUSIONS

DAB converters are an attractive approach to realize high power high frequency dc-dc converter applications, although transformer design remains a persistent barrier to broad adoption. This paper introduces the use of coupling capacitors in DAB converters along with multiple stages to overcome this barrier. The paper has presented detailed design equations for developing designs for both the topologies based on a unified equivalent circuit. Design considerations that identify...
the voltage transformation ratio to be nominally equal to the turns ratio of the transformer in the case of TCDAB, and equal to the number of stages in case of CCDAB and operation at reasonably small phase shift between the bridges have been used to develop simplifying approximations for the analytical model. The detailed designs have been verified using computer simulations, while experimental validation of the converter operation have been demonstrated elsewhere in other literature.

A detailed comparative evaluation between the two realizations based on a design example indicates that the CCDAB converter has the potential to be more compact than TCDAB converter with similar levels of operating efficiency. Modern semiconductor materials like GaN and SiC allow for high switching frequencies, but any miniaturization of power circuit is limited by the physical properties of passive magnetic components. This paper shows that CCDAB is a viable approach to utilize the characteristics of modern power switching devices to their highest potential, by overcoming the barriers posed by the magnetic components and reducing the volume/weight of the converter while preserving the attractive features of TCDAB.

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