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***Abstract* —The paper investigates the difference of two PWM control methods for dual bridge matrix converters (DBMC). The first method is well known and allows zero current switching of the rectifier side switching. The second method is proposed by this paper which allows zero voltage switching in the inverter IGBTs. This paper analyses the power losses and lone term reliability of IGBT modules in the DBMC used as a motor drive for both methods. It was shown that the switching losses of the IGBT for DBMC can be shifted between the rectifier and inverter by switching the two PWM control method. As a result, the DBMC shows some advantages over other topologies in terms of the sizing of the IGBT, balancing of the thermal system and easier to operate under higher switching frequency conditions.**

Keywords – power cycle, dual bridge matrix converter, Mean Time to Failure (MTTF), zero voltage switching, zero current switching

I. INTRODUCTION

The dual-bridge matrix converter (DBMC) is a relative new developed concept [1][2][3]. Compared to the conventional matrix converter [4], it possesses the same high quality performance, including near sinusoidal input/output waveforms, adjustable input power factor, and a compact system design due to the absence of large energy storage components. Moreover, it has a number other advantages over the conventional matrix converter, including safer commutation, reduced number of switches, simple clamp circuit, easier operation with multi-motor configurations [5], etc. With all above advantages, the DBMC is suitable to be used for variable frequency AC/AC conversion applications, including distributed electrical power systems or many on-board electrical power generation systems. The freedom of selecting both input and output frequencies can help operate the system at higher frequency to reduce the volume, weight and cost.

One disadvantage of the matrix converter is that it has much higher numbers of power switches than other topologies. Since power semiconductor is one of the most costly components in a converter, it is beneficial to investigate the long term reliability and the sizing of the semiconductors used in the matrix converter. The end of life period of complex multi-chip modules is often defined by the thermo-mechanical failure mechanisms [6][7][8]. One of most important failure schemes is called power cycling capability. The power cycling capability defines wear out mechanisms of the bond-wire on the silicone chip. It is found that the lifetime of the bondwire for the chip might be very short if the temperature variation on the IGBT is higher. Thus, it is always favorable to have a minimized junction temperature variation on the IGBT for an converter to guarantee a longer lifetime of the module.

This paper proposes a zero voltage switching PWM method (ZVPWM) for DBMC. With this method, the inverter side of the converter commutates at zero voltage and thus has no switching losses. Comparing to the tradition zero current switching method (ZCPWM) at rectifier side IGBTs, the proposed method can help balance the junction temperature of both rectifier and inverter side IGBT and reduce the maximum temperature variation of the IGBT chips. As a result, a longer overall lifetime of the IGBT modules can be achieved.

The paper is organized into the following steps, firstly, the paper analyzes the proposed PWM method and its difference between the traditional PWM method. After that, the power losses of both IGBT and Diode are calculated and compared. Then, the junction temperature variations of the IGBTs under both conditions are studied. A general discussion on how to maximize the lifetime of the DBMC will also be provided in the final paper.

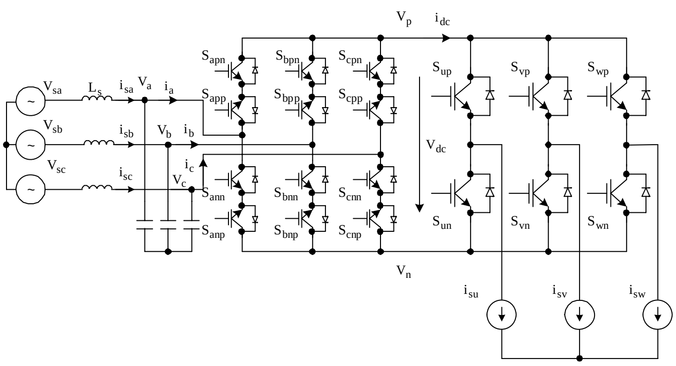


Figure 1. 18 switch dual bridge matrix converter topology

II. CIRCUIT AND SYSTEM CONFIGURATION

In order to simplify the analysis, it is assume that there is no input filter in the input of the DBMC, thus, the following equations can be obtained from figure.1.

$$L_s = 0\,;\; C_f = 0\,;\; V_{sx} = V_x\,, i_{sx} = i_x \qquad (1)$$

where

L_s/C_f is the filter inductance/capacitance

x, input phase name, can be phase A, B or C

V_{sx}, i_{sx}: phase x source voltage and current

V_x, i_x: phase x converter side voltage and current

It is assumed that the input source voltage and output load are both balanced. The input voltage $\vec{V}_s$ and output current $\vec{i}_o$ are represented as:

$$\vec{V}_s = \sqrt{\frac{2}{3}}(V_{sa} + \alpha \cdot V_{sb} + \alpha^2 \cdot V_{sc}) = V_m e^{j\omega_i t}$$
$$\vec{i}_o = \sqrt{\frac{2}{3}}(i_u + \alpha \cdot i_v + \alpha^2 \cdot i_w) = i_{om} e^{j(\omega_o t + \varphi_i)} \qquad (2)$$

where, $\alpha = e^{i2\pi/3}$, V_m and i_{om} are the per phase input voltage and output current amplitude respectively. ω_i and ω_o are the angular frequency of the input and output of the converter respectively. φ_i is the initial angle of the output current.

Another attractive character of the matrix converter is that a unity power factor can be achieved from the input side. Under this condition, the maximum voltage transfer ratio can be achieved. In this paper, it is assumed that the matrix converter is controlled under unity input power factor. Then, the following assumption can be made:

$$\vec{i}_s = \sqrt{\frac{2}{3}}(i_a + \alpha \cdot i_b + \alpha^2 \cdot i_c) = i_m e^{j\theta_a}$$
$$\vec{V}_o = V_{om} e^{j(\omega_o t + \varphi_o)} = kV_m e^{j\theta_o} \qquad (3)$$

where, $0 \le k \le 0.866$ is the voltage transfer ratio of the converter, φ_v is the angle of the converter output voltage.

A three phase matrix converter is studied to drive an induction motor. The following information is defined:

Input line voltage:	480Vac line to line
Output frequency:	1Hz ~ 60Hz
Switching frequency:	4 kHz,
Output rated voltage:	415Vac line to line voltage
Rated current:	65Arms
IGBT module:	FS100R12KE3
Motor power factor:	0.85

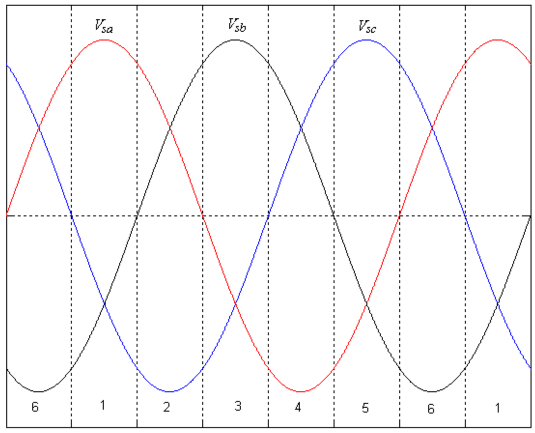
III. Duty Ratio And Power Losses Calculation

A. *Space Vector Control of the Rectifier*

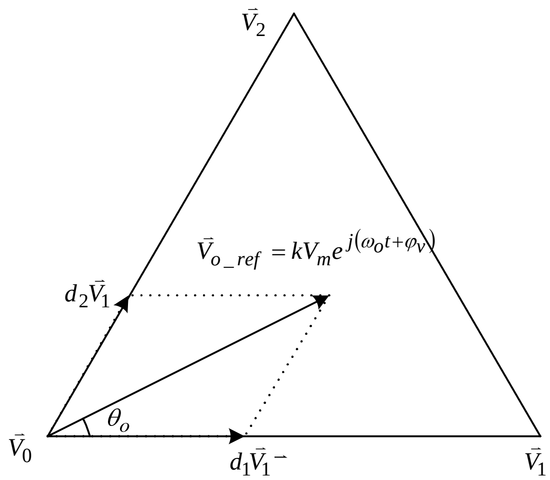
A similar duty ratio calculation method as discussed in [9] separates the matrix converter into six intervals for line side zero current switching can be performed same as Figure.2 (a). In each PWM cycle, two non-zero vectors of the line side converter will be generated for different intervals. Table.I further explains the non zero vectors of the rectifier in each interval and their corresponding duty ratios.

TABLE I. Non-Zero Vector of Line Side Switching Under Each Interval at Each PWM Cycle

interval	Non-Zero V_{r1}				Non Zero V_{r2}			
	d_{r1}	V_{sx}	V_{sy}	V_{dc}	d_{r2}	V_{sx}	V_{sy}	V_{dc}
1	$\lvert\cos\theta_b\rvert$	V_{sa}	V_{sb}	V_{sab}	$\lvert\cos\theta_c\rvert$	V_{sa}	V_{sc}	V_{sac}
2	$\lvert\cos\theta_b\rvert$	V_{sb}	V_{sc}	V_{sbc}	$\lvert\cos\theta_a\rvert$	V_{sa}	V_{sc}	V_{sac}
3	$\lvert\cos\theta_c\rvert$	V_{sb}	V_{sc}	V_{sbc}	$\lvert\cos\theta_a\rvert$	V_{sb}	V_{sa}	V_{sba}
4	$\lvert\cos\theta_c\rvert$	V_{sc}	V_{sa}	V_{sca}	$\lvert\cos\theta_b\rvert$	V_{sb}	V_{sa}	V_{sba}
5	$\lvert\cos\theta_a\rvert$	V_{sc}	V_{sa}	V_{sca}	$\lvert\cos\theta_b\rvert$	V_{sc}	V_{sb}	V_{scb}
6	$\lvert\cos\theta_a\rvert$	V_{sa}	V_{sb}	V_{sab}	$\lvert\cos\theta_c\rvert$	V_{sc}	V_{sb}	V_{scb}



(a)



(b)

Figure 2. line side intervals definition and load side space vector PWM (a) six intervals for the rectifier, (b) space vector of the inverter

B. Space Vector Control of the Inverter

For inverter side, assuming that $0<\theta_o<\pi/3$ and the system is operating in interval 2, the output voltage vector can be approximated by its three adjacent voltage vectors $\vec{V}_1$, $\vec{V}_2$, and the zero voltage vector $\vec{V}_0$ as shown in Figure 2 (b).

The duty ratios of these vectors are, respectively,

$$d_1 = \frac{2k}{\sqrt{3}}\sin(\frac{\pi}{3}-\theta_o)$$

$$d_2 = \frac{2k}{\sqrt{3}}\sin\theta_o;$$

$$d_0 = 1 - d_1 - d_2 \qquad (4)$$

C. Duty Ratio Calculation of the DBMC

Combining (3) and (4), in each PWM cycle, the DBMC can have four non-zero vectors with the following duty ratio.

- Line side V_{r1}, inverter side V_1, duty ratio: $d_{11} = d_1 d_{r1}$
- Line side V_{r2}, inverter side V_1, duty ratio: $d_{21} = d_1 d_{r2}$
- Line side V_{r1}, inverter side V_2, duty ratio: $d_{12} = d_2 d_{r1}$
- Line side V_{r2}, inverter side V_2, duty ratio: $d_{22} = d_2 d_{r2}$

The duty ratio of the zero vectors can be calculated as

$$d_0 = 1 - d_{11} - d_{12} - d_{21} - d_{22} \qquad (5)$$

IV. LINE SIDE ZERO CURRENT SWITCHING AND LOAD SIDE ZERO VOLTAGE SWITCHING

It is well known that both the line and load side converter can generate zero vectors for DBMC. When either side generates zero vectors to the system, the input current vector as well as the output voltage command of the DBMC becomes zero.

When the load side converter generates zero vectors, the DC bus current becomes zero. Zero current switching of the line side converter can be guaranteed if the line side converter switches under this condition [2][3] . Figure.3 (a) shows a typical schematic of the converter when zero vector is generated in the load side inverter.

In contrary, Figure.3 (b) shows the schematics where the input converter generates zero vectors. Under this condition, the DC bus voltage becomes zero and a zero voltage switching can be generated if the inverter commutates under this condition.

The switching losses may be either generated in the input side converter or load side inverter. For example, if it is decided that the line side IGBT can only switch when the load side vector is zero, zero current commutation of the line side converter can be guarantee. Reference [2][3] discussed this method in detail.

Similarly, zero voltage switching of the load side converter can be applied. For instance, if it is decided that the load side IGBT can only switch when the line side vector is zero, zero voltage commutation of the load side IGBT can be guaranteed.

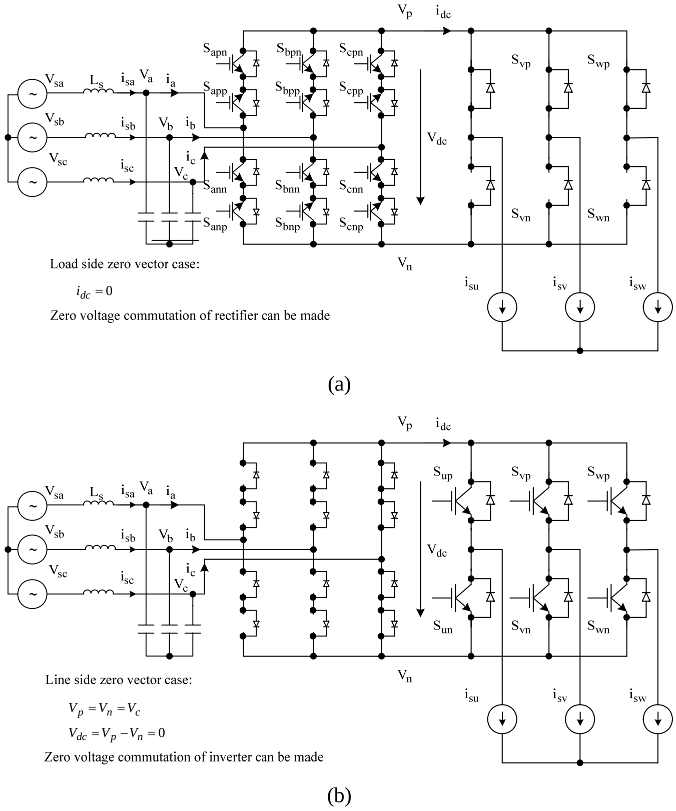


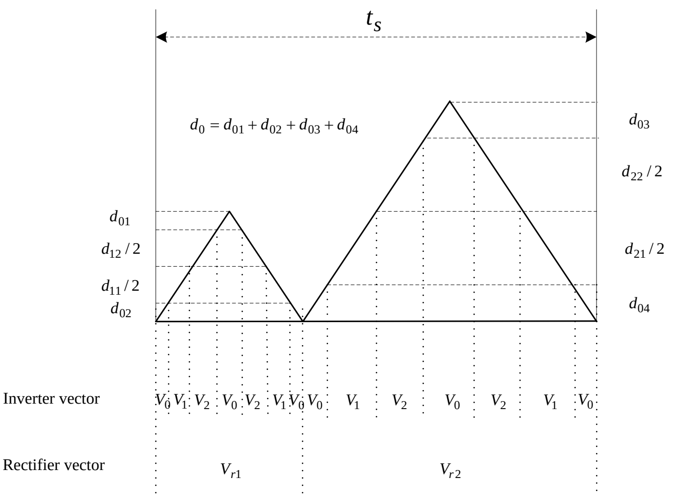
Figure 3. zero current and zero voltage switching (a) line side zero current switching (c) line side zero voltage switching

Appropriately organizing the four non-zero vectors in one PWM cycle results in either inverter side or rectifier side zero switching losses.

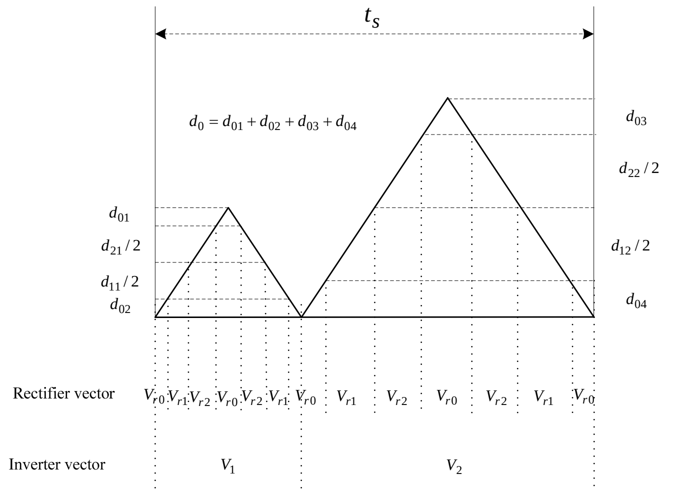
Figure.4 (a) shows one example of rectifier side zero current switching method. Under this condition, the rectifier side vector switches only two time between vector V_{r1} and V_{r2}. The space vector PWM of the inverter repeated for each non-zero vectors of the rectifier. From this figure, it can be found that the rectifier side switching only switching at zero vector of the inverter. As a result, zero current switching of the rectifier side IGBT can be guaranteed.

Figure.4 (b) shows one example of load side zero voltage switching method. In contrary, the inverter side vector switches only two time between vector V_1 and V_2 in each PWM cycle. The space vector PWM of the rectifier repeats for each non-zero vectors of the inverter. From this figure, the inverter IGBT switches only commutate at zero vector of the rectifier. As a result, zero voltage switching of the rectifier side IGBT can be guaranteed.

The difference between the two switching methods is that the switching losses of one side of the converter can be zero. By doing this, one can adjust the power losses between the rectifier and inverter switches. As a result, a maximum long term reliability of the converter can be achieved.



(a)



(b)

Figure 4. Zero voltage switching and zero current switching of the DBMC, (a) line side zero current switching, (b): load side zero voltage switching

V. Power Losses and Junction Temperature Profile Comparison of the Two Methods

One of most critical operation condition for a traditional inverter using as a motor drive is the low speed operating condition. Under this condition, the power losses of the IGBT follow the same frequency of the out frequency. As a result, the instantaneous junction temperature is much higher than high speed operating condition.

The two PWM control method can move the switching losses from the inverter side to that of the rectifier side. As a result, the low speed performance of the

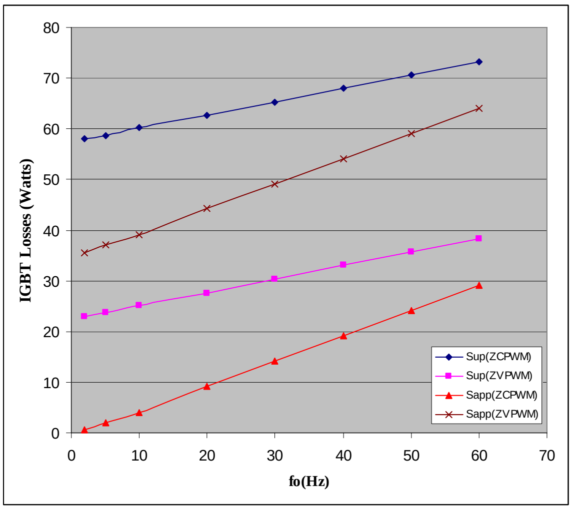
A. Comaprison of Rectifier / Inverter Side IGBT losses

Figure.5 (a) shows the power losses difference of the inverter side IGBT S_{up} and S_{app} between the two PWM control methods under rated output current. From this figure, it can be found that the power losses of Sup are much lower under ZVPWM conditions. On the other hand, S_{app} has much higher losses under ZVPWM.

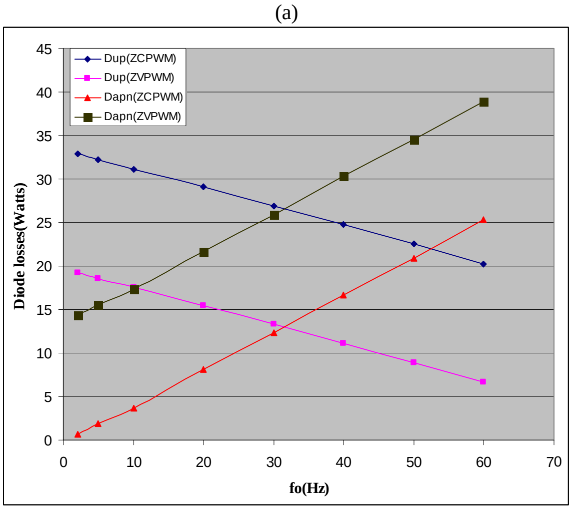
Figure.5 (b) shows the power losses of the D_{up} and D_{apn} between the two PWM control methods. The similar conclusions can be derived.

From Figure.5 (a) and (b), it can be found that the inverter side switches S_{up} and D_{up} shows much higher power losses under ZCPWM. On the hand, the rectifier side switches shows much higher power losses under ZVPWM.

From Figure.5 (a), it can also found that the power losses between Sup and Sapp are closer under ZVPWM than under ZCPWM.



(a)



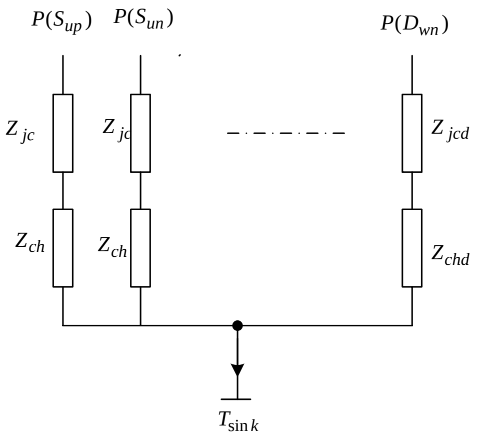
(b)

Figure 5. Comparison of average power losses at different output frequency and rated output current (65Arms). (a) IGBT losses comparison, (b) Diode losses comparison

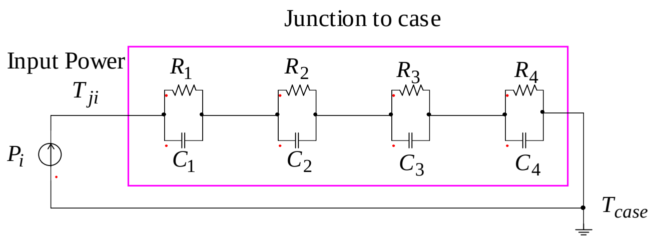
B. Junction Temperature Variation of Inverter IGBT

One of most critical operation condition for a traditional inverter using as a motor drive is the low speed operating condition. This is due to the following two reasons, first, because the power losses of the IGBT follow the same frequency of the out frequency, the maximum junction temperature of the IGBT are higher than high frequency condition [6]. Second, the temperature variation of the junction are much higher than that of the high output frequency condition, the long term reliability of the IGBTs are largely reduced [7][8].

The junction temperature profile of the inverter side IGBT are also studied under various output frequencies for the two PWM method. During the simulation, it is assumed that the case temperature of each IGBT is 80°C. The junction two case layer thermal impedances are shown in Figure.6 (b) and Table.II.



(a)



(b)

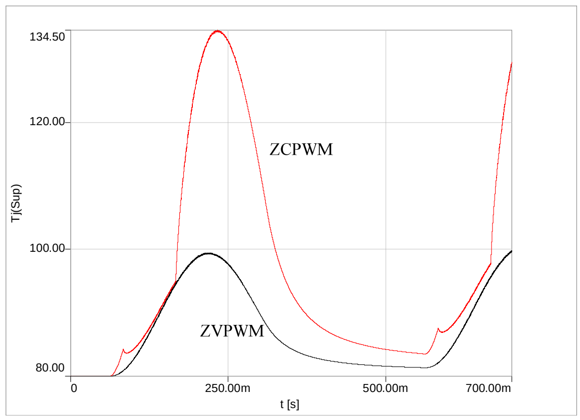
Figure 6. Simplified thermal system of the maxtrix converter system. (a) overall thermal network model, (b) thermal impedance between the junction and case layer of one chip.

TABLE II. THERMAL IMPEDANCE OF THE TESTED IGBT/DIODE

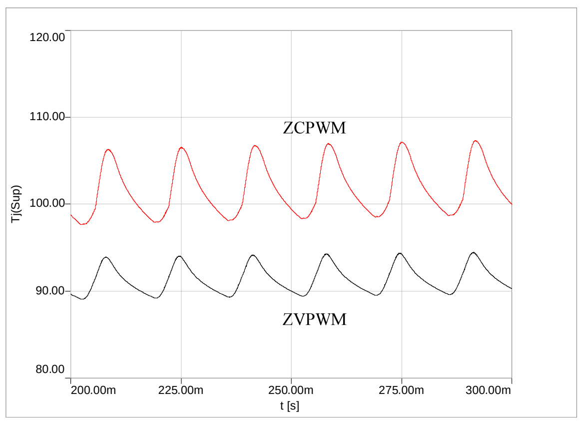
Segments	1	2	3	4	Zch
Ri: (K/W)	0.00493	0.01501	0.13088	0.10919	0.13
τi: (s)	1.187e-5	0.002364	0.02601	0.06499	0.7
Rd: (K/W)	0.00908	0.02726	0.24202	0.20164	0.15
τd: (s)	1.187e-5	0.002364	0.02601	0.06499	0.7

From Figure.7 (a) and (b) shows temperature profile of the inverter side IGBT Sup under 60Hz and 2Hz respectively and rated output current. The following conclusions can be made

- The temperature variation of 60Hz condition for both PWM control method are very small.
- When the motor operates at 2Hz condition, the temperature variation under each fundamental cycle are much higher under ZVPWM than that of the ZCPWM method. Thus, a much longer lifetime can be expected on ZVPWM due to the IGBT bondwire lift-off mechanism. Thus, ZVPWM is more favorable for the DBMC to operate under low speed or DC condition
- The maximum junction temperature of ZVPWM is much lower than that of the ZCPWM. As a result, the output current capability of the ZVPWM is much higher than that of the ZCPWM at low speed conditions.
- Due to zero switching losses at inverter side for the ZVPWM, the maximum junction temperature at 2Hz condition is only slightly higher than that at 60Hz condition. Under ZCPWM, the maximum junction temperature of S_{up} at 2Hz is much higher than that of the 60Hz condition.
- In the final paper, the long term lifetime expectation of the bond-wire for each IGBT will be provided.
- In the final paper, the performance of the rectifier side IGBT will also be evaluated and discussed.



(a)



(b)

Figure 7. Junction temperature profile of Sup at various output frequency under rated output current 65Arms (a) 2Hz, (b) 60Hz

VI. CONCLUSION

This paper discussed two PWM control methods for DBMC. The first method allows a zero current switching on the rectifier side switches. The second methods were proposed by this paper which allows zero voltage switching on the load side inverter.

The following conclusion are made

- The ZVPWM method can minimize the losses of the inverter side IGBT and Diode
- The ZCPWM method minimizes the losses of the rectifier side IGBT/Diode.
- Under low speed condition, the inverter IGBT are the most stressed and ZVPWM is more favorable to reduce the maximum junction temperature as well as the temperature variation of the junction for each cycle.
- Switching between the ZCPWM and ZVPWM can balance the power losses of the IGBTs and help maximizing the silicone usage and long term reliability.

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