Open-phase fault tolerant driving operation of dual-inverter-based traction drive

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Abstract
A method for achieving open-phase fault tolerant operation of a dual-inverter-based traction drive with charging capability is presented. This method utilises a contactor to connect the switching nodes of the two half bridges which are normally used for AC charging in this dual inverter configuration. These half bridges are then modulated to provide a path for zero-sequence current and enhance the available voltage vector space of the drive by injecting a controllable common-mode voltage between the two inverters. A control method and five-level modulator based on zero-sequence voltage control and phase shifted carriers are introduced to accommodate this novel remedial method. Using this approach, the open-phase faulted dual inverter drive is able to operate with twice the speed range of the conventional open-phase fault tolerant approach, which is to provide a path for zero-sequence current by connecting the negative DC rails of the two inverters with a contactor. Simulation and experimental results with a prototype dual inverter drive are presented to validate the theory of the new control and modulation method. The experimental results showcase the proposed algorithms performance by dynamically responding to an open-phase fault and starting from standstill, thereby proving its validity for an electric vehicle application.

1 | INTRODUCTION

Global concerns regarding greenhouse gas emissions have contributed towards the electrification of transportation. Key components of electric vehicles (EVs) are the traction motor, inverter and energy storage device. The most commonly used energy storage device is batteries, which require charging infrastructure either in the form of onboard charger or fast charging technologies [1].

Both of these charging methods require additional power electronic and magnetic components to perform the power conversion required for battery charging. An alternative charging strategy which does not require these additional components is onboard integrated charging. This method reutilises the traction inverter as the charging power electronic converter and the motor windings as the required magnetic components [2]. Compared to nonintegrated onboard chargers, this approach results in faster charging capability (since the traction inverter and motor are dimensioned for significantly greater power levels than onboard chargers) [3] and reduced vehicle weight since a dedicated charging unit is not needed.

In reference [4] an onboard integrated charging topology based on the dual two-level inverter and open-end winding (OeW) permanent magnet synchronous machine (PMSM) was introduced. This system is capable of achieving fast charging from a DC source, as well as conventional motor driving operation based on the modulation techniques described by [5]. A modification of this topology was proposed in [6] wherein two auxiliary half-bridges were introduced to enable charging from a single-phase AC source. Figure 1 shows two embodiments of this topology that can be used in open-phase fault tolerant operation (including ports for AC and DC charging).

The ability to continue operation under faulted conditions is essential for electric drives used for EV propulsion, since this allows ‘limp home’ functionality which allows users to safely drive to a location where the vehicle can be repaired [7, 8]. Faults can occur in either the power electronic switches or motor of the EV drive system.

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Voltage-source inverter electrical faults can be categorised into insulated-gate bipolar transistor (IGBT) open and short-circuit faults [9]. Open circuit faults will ultimately result in a DC current offset for the faulted and healthy phases [10]. Multiple strategies exist for detecting such a condition, such as Park’s vector method [11] or normalised DC current methods [12]. Inverter short-circuit faults can be detected by desaturation protection circuitry in the gate driver [13–15]. After detection, open or short-circuited inverter phases could be isolated using methods proposed in [7]. The application of such fault-isolation methods would cause the motor to be in an open-phase fault mode, where the phase with faulty power electronics would be disconnected.

Common electric motor faults include air gap eccentricity, bearing failure and stator winding failures [16]. Stator winding failures can be characterised by insulation failure, which manifests as short circuits either at the turn-to-turn, coil-to-coil or phase-to-phase levels [17]. Alternatively, open-circuit faults can occur within a motor phase due to mechanical failure in a terminal connector, or internal winding rupture [18].
The presence of a motor open-phase fault could be detected in the drive circuitry by the current measurement in the associated phase becoming zero.

A remedial strategy for open-phase fault tolerant operation of a four-leg interior PMSM drive was presented in [19, 20], where a fourth inverter leg was connected to the motor’s neutral point. In the faulted case, the modulation of this fourth leg was achieved either through a hysteresis current controller acting upon the neutral current, or a field oriented control strategy with space vector modulator, wherein the gating signals for the faulted phase leg were applied to the fourth leg. This study was extended by [21] where field weakening and sensorless control methods were introduced for the four-leg drive.

Open-phase fault tolerance has also been investigated for nonconventional motor topologies. Reference [22] investigated a phase-loss fault tolerant strategy for a flux-modulated permanent magnet compact in-wheel motor. Postfault operation was achieved by changing the winding configuration from star to delta and modifying the control strategy accordingly. Remedial operation of a dual three-phase motor drive using dual T-type three level inverters was investigated by [23]. Remedial operation was achieved for inverter open- and short-circuit faults by modifying the modulation strategies of the T-type converters.

The methods discussed in [19, 20] utilised feedforward compensation in the stationary reference frame to derive required reference voltage vectors for the open-phase faulted four-leg inverter drive. A similar approach was adopted in [24] for a four-leg matrix converter drive. A method has been proposed by [25] which uses a modified Park transformation for the open-phase faulted PMSM to allow control without the need of feedforward terms.

Open-phase fault-tolerant operation of an onboard integrated charger based on a dual inverter with a single DC voltage source was presented in [26]. This study introduced a method of space vector modulation optimised for postfault remedial operation of the system. The single DC supply of this topology meant that unlike the system of Figure 1, a common-mode path naturally exists, which can be leveraged in case of an open-phase fault. A similar study was conducted in [27] wherein a dual inverter based OeW PMSM drive with single DC voltage source was controlled after an open-phase fault mode as two independent H-bridges.

This study presents methods of configuring the topology shown in Figure 1 to continue operation despite an open-phase fault of the drive system. This is achieved through the addition of a contactor, \( S_1 \), either between the negative DC links of each inverter (upper image) or between the switch nodes of the auxiliary half bridges (lower image). The latter method is found to improve the available voltage range for the dual inverter drive in the case of a single phase open-circuit fault. Additionally, \( S_1 \) can be kept closed during driving operation for the proposed method if IGBTs \( Q_{dL}, Q_{dR}, Q_{dL}', \) and \( Q_{dR}' \) are gated off. Modulation of the two auxiliary half-bridges during an open-phase fault allows the common-mode voltage between the two inverters to be controlled. Using this strategy, the available voltage vector space is found to be the same as the nonfaulted dual inverter (in healthy three-phase operation).

It should be noted that this study only focuses on remedial operation aspect of the fault tolerant drive. The fault diagnosis and isolation procedure are not covered, but methods described in previous studies could be utilised.

2 | MODELLING OF OPEN-PHASE FAULT

2.1 | Current vector space

The absence of a zero-sequence path in dual inverter with isolated DC supplies gives rise to the following equation:

\[ i_a + i_b + i_c = 0 \] (1)

If one phase of the drive (for example, phase \( a \)) exhibits an open-circuit fault, Equation (1) will necessitate \( i_a = -i_c \). The Clarke transform can be utilised to transform the obtained stationary \((a\beta)\) reference frame:

\[
\begin{bmatrix}
  i_a \\
  i_b \\
  i_0
\end{bmatrix} = \frac{2}{3} \begin{bmatrix}
  1 & -1 & 0 \\
  -\frac{1}{2} & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\
  1 & 1 & 1
\end{bmatrix} \begin{bmatrix}
  i_a \\
  i_b \\
  i_c
\end{bmatrix} \] (2)

Setting \( i_a = 0, i_b = -i_c \) in Equation 2 will result in an \( a\beta \) trajectory which is only in the \( \beta \) axis. This is evident from Figure 2 which shows the trajectory for the open-phase faulted motor with no neutral path as a vertical line. The available vector space of a healthy motor with all three phases operational is plotted in this figure as a unit circle. The trajectory of the faulted motor is reminiscent of that of a single-phase motor, which does not have self-starting capability. This is unacceptable for an EV, which must naturally be capable of restarting after the vehicle comes to a halt. The addition of a common-mode path relaxes the constraint imposed by Equation (1).

The control method introduced by [20] for control of three-phase wye-connected interior permanent magnet machine with a four-leg inverter is adapted for the traction inverter with integrated charging studied in [6]. If the motor phase which exhibits the open-circuit fault is defined as phase \( a \), setting \( i_0 = -i_a \) causes the value of \( i_c \) computed by the inverse Clarke transform Equation (2) to be zero. This constraint will cause a 60 degree phase shift to occur between \( i_b \) and \( i_c \).

The presence of this 60 degree phase means that the motor phase currents have to be limited to a peak value of \( \frac{1}{\sqrt{3}} \) pu to ensure that the neutral current (given by \( i_b + i_c \)) will not exceed 1 pu.
The current vector space available in the $a\beta$ frame when this limitation is imposed and the open-phase faulted motor has $S_1$ closed is shown in Figure 2. This space is a circle with a radius of $1/3$ pu, compared with the unit circle that is obtained for the healthy system. This indicates that the system with an open-circuit fault will have one-third the torque production capability of the healthy drive. It should be noted that the neutral current present in this case will also create a zero sequence current which is orthogonal to the $a\beta$ plane in Figure 2. This current does not contribute to torque or magnetising flux in an electric motor. As such, the $a\beta$ projection of the motor current space vector shown in Figure 2 is used for control purposes.

### 2.2 Voltage vector space

This study considers two methods of creating this neutral path in the event of an open-phase fault. The conventional method, pictured in the upper image of Figure 1 uses a contactor between the negative DC link of each inverter. Closing this contactor after an open-phase fault has occurred provides the neutral path and sets the common-mode voltage between the two half-bridges ($v_{g1g2}$) to be zero. Operation after this reconfiguration would then be akin to the methods proposed by [26, 27].

The proposed method (shown in the lower image of Figure 1) is to connect a contactor between the switching nodes of the two auxiliary half-bridges which are normally used for single-phase charging operation. If the contactor $S_1$ is closed in this position, $v_{g1g2}$ can be obtained from the following equation, where $S_{Ua}, S_{Lb}$ are the switching functions of the auxiliary half-bridges formed by $Q_{Ua}, Q_{Lb}$ and $V_{bU}, V_{bL}$ are the two battery voltages.

$$v_{g1g2} = S_{Ua}V_{bU} - S_{Lb}V_{bL}$$ (3)

If $V_{bU} = V_{bL} = v_{dc}$ the following set of values can be obtained for $v_{g1g2}$ based on the values of $S_{Ua}$ and $S_{Lb}$: The motor phase $b$ and $c$ voltages can be obtained by computing:

$$v_{lb} = v_{b1} - v_{b2} + v_{g1g2}$$ (4)

$$v_{lc} = v_{c1} - v_{c2} + v_{g1g2}$$ (5)

Applying the Clarke transform on the values of $v_{lb}$ and $v_{lc}$ obtained for both systems (considering $v_{mc} = 0$) results in the set of voltage vectors in the $a\beta$ plane pictured in Figure 3. The available voltage vectors from the proposed method are plotted as circles, while the vectors from the conventional method are plotted as crosses. It should be noted that the proposed method also produces voltage in the zero-axis, orthogonal to the $a\beta$ plane. Since this voltage does not contribute to torque or flux production in the motor, it is not considered.

It is interesting to note that the voltage vector space available for the proposed method is the same as the healthy dual inverter drive first studied in [5]. This implies that utilisation of the auxiliary half bridges during an open-phase fault allows the same voltage range of a healthy drive to be obtained.
The influence of modulating the auxiliary half bridges in the proposed method can be understood by considering the derivation of a single voltage vector. Consider the voltage vector for the conventional method at \([v_a, v_b] = [0, 4/3]\) (the vertical arrow in Figure 3). This vector can be achieved with \([S_{stU}, S_{stL}, S_{clU}, S_{clL}] = [0, 1, 1, 0]\) which gives rise to \(v_{bs} = -1\) pu, \(v_{cs} = 1\) pu (remembering that \(v_{g1g2} = 0\) in the conventional method).

If the proposed method is used, the value of \(v_{g1g2}\) becomes controllable according to Table 1. Setting \(v_{g1g2} = -1\) pu will cause an influence to \(v_a\) of \(2+2/3\) pu (obtained by evaluating the Clarke transform shown in Equation (2)), which can be seen as the horizontal arrow of Figure 3. The summation of the vertical and horizontal voltage vectors of Figure 3 generates the diagonal vector, pointing from the origin to \([v_a, v_b]\) = \([2/3, 4/3]\).

Alternatively, modulating the auxiliary half bridges to set \(v_{g1g2} = +1\) pu will generate an offset of \(-2/3\) pu to \(v_a\), and ultimately give rise to an overall voltage vector from the origin to \([v_a, v_b]\) = \([-2/3, 4/3]\). Thus, it can be seen that control of \(v_{g1g2}\) during the open-phase fault allows the voltage vectors produced by the conventional method to be shifted horizontally on the \(\alpha\) plane.

The modulation ranges achievable with space vector modulation are shown as the circles inscribed within the hexagons of this Figure 3. The maximum peak phase voltage which could be produced by the proposed method is given by:

\[
V_{m1} = \frac{V_{blU} + V_{bl}}{\sqrt{3}}
\]  

In contrast, the modulation range achievable for the conventional method results in the following peak phase voltage:

\[
V_{m2} = \frac{V_{blU} + V_{bl}}{2\sqrt{3}}
\]

The following section explores how the limits in the available current and voltage vector space relate to the operating torque and speed range of the faulted drive system.

### Table 1: Common-mode voltages for the proposed method, with \(S_1\) closed

| \(S_{stU}\) | \(S_{stL}\) | \(v_{g1g2}\) |
|-----------|-----------|----------|
| 0         | 0         | 0        |
| 0         | 1         | \(v_{dc}\) |
| 1         | 0         | \(-v_{dc}\) |
| 1         | 1         | 0        |

#### 2.3 Operating limits

While the available current and voltage vector space are of interest in the analysis of an EV drive, the torque versus speed characteristics are of most practical importance. The electromagnetic torque produced by a nonsalient PMSM is obtained by:

\[
T_{em} = \frac{3}{2} \psi_m i_q
\]
where $\psi_m$ is the PMSMs flux linkage, $p$ is the number of rotor pole pairs and $I_q$ is the motor $q$-axis current.

The base electrical frequency (the maximum electrical frequency without requiring field weakening considering maximum torque production) can be given by:

$$\omega_b = \frac{V_m}{\sqrt{\left(\psi_m^2 + (L_qI_m)^2\right)}}$$  \hspace{1cm} (9)

where $L_q$ is the motor $q$-axis inductance and $V_m$ is the maximum available voltage for a given converter configuration, which is provided in Equations (6) and (7). $I_m$ is motor current vector limit which was shown for the healthy and faulty cases in Figure 2.

Once the operating electric frequency of the PMSM is greater than the base frequency ($\omega > \omega_b$), the required field weakening $d$-axis current reference can be computed from:

$$i_d^* = \frac{(\omega - \omega_b)\psi_m}{L_d\omega}$$  \hspace{1cm} (10)

where $L_d$ is the motor $d$-axis inductance. Together with the voltage and current limitations of the drive, Equations (8–10) can be utilised to derive the torque versus speed characteristics of a given motor under healthy conditions as well as open-phase faults with the two remedial methods. These calculated curves for a TM4 HSM60 surface permanent magnet traction motor (parameters in Table 2) are shown in Figure 4.

The maximum torque of the healthy drive system is three times greater than either faulted case, due to the restriction placed on $I_m$ during the faulted case to ensure that the neutral current does not exceed the maximum current of the drive system. The maximum operating speed of the healthy drive is substantially higher than either faulted system, since the larger achievable value of $I_m$ allows a greater magnitude of injected $i_d^*$ during field weakening.

| TABLE 2 Parameters of TM4 HSM60 motor under study |
|----------------------|-----------------|----------|
| Parameter            | Description     | Value    |
| $P$                  | Pole pairs      | 5        |
| $L_d$                | $d$-axis inductance | 0.837 mH |
| $L_q$                | $q$-axis inductance | 0.837 mH |
| $L_o$                | Leakage inductance | 0.471 mH |
| $\psi_m$             | Magnet flux linkage | 0.127 Wb |
| $R_s$                | Stator resistance | 45 mΩ    |
| $I_m$                | Stator current limit | 200 A    |
| $P_{nom}$            | Nominal power   | 65 kW    |
| $J_{rot}$            | Rotor inertia   | 0.09 kgm² |
| $V_{b1}$             | Voltage of battery 1 | 400 V    |
| $V_{b2}$             | Voltage of battery 2 | 400 V    |
| $f_{sw}$             | Switching frequency | 10 kHz  |

Comparison of the torque-speed curves in the two faulted cases shows that the maximum achievable speed using the proposed method is doubled compared to that of the conventional method after an open-phase fault. This is due to Equations 6 and 7 which show a ratio of two between $V_{m1}$ and $V_{m2}$.

The following sections will investigate the control and modulation strategy which will be used with the proposed remedial method of the dual inverter drive.

3 | CONTROL METHODOLOGY

3.1 | Current controller

The current control is performed in the synchronous reference frame [28]. Figure 5 shows the control diagram. Traditional speed and current controllers with feedforward decoupling in the $dq$ frame are used.

A carrier based modulation scheme is used in this control algorithm. As a result, the reference voltages generated by the controllers must be in the stationary $abc$ frame. These $abc$ voltage references can be obtained from a set of required $\alpha\beta\gamma$ references via the inverse Clarke transform:

$$\begin{bmatrix}
\psi_a^* \\
\psi_b^* \\
\psi_c^*
\end{bmatrix} = \begin{bmatrix}
1 & 0 & 1 \\
\frac{1}{2} & \frac{\sqrt{3}}{2} & 1 \\
\frac{1}{2} & -\frac{\sqrt{3}}{2} & 1
\end{bmatrix} \begin{bmatrix}
\psi_a \\
\psi_b \\
\psi_c
\end{bmatrix}$$  \hspace{1cm} (11)

In an open-phase faulted system, phase $a$ of the dual inverter cannot apply any voltage to the motor. It should be noted, however, that the phase $a$ voltage at the motor terminals would be the open-circuit induced voltage, $e_a$. A modification must be made to the $\alpha\beta\gamma$ voltage references supplied to the inverse Clarke transform in order to ensure that the phase $a$ reference voltage will be equal to $e_a$. This is achieved by modifying the zero-sequence reference voltage, $\psi_0^*$:

$$\psi_0^* = \bar{e}_a - \psi_a^*$$  \hspace{1cm} (12)

where $\bar{e}_a$ is the estimated phase $a$ open-circuit induced voltage, given by [20]:

$$\bar{e}_a = -\omega \left(\psi_m + (L_d - L_o)i_d\right) \sin \theta + (L_q - L_o)i_q \cos \theta$$  \hspace{1cm} (13)

Figure 5 shows that this estimated voltage is used as a feedforward compensation term. This is akin to the approach used by [20], with the exception that the feedforward term is injected into the stationary reference frame, rather than the $dq$ frame. Injection in the stationary reference frame greatly simplifies the required computations for the feedforward term, and is possible because this study uses a specialised carrier-
based modulation scheme rather than a standard space vector modulator as in [20]. Details of this modulator will be provided in the following section.

3.2 Modulation

A modulator based on the principle of phase-shifted carriers [29] is used to obtain the gating signals of the dual inverter and auxiliary half bridges for the proposed method (Figure 6). The phase shifts of the carrier waveforms are determined in the same method as a five-level converter, resulting in a 90 degree phase shift between each carrier waveform. Figure 7 shows an example of the carrier waveforms used.

The reference voltage used to determine the control signals for the auxiliary half bridges (m0*) is equal to the zero sequence voltage applied by the dual inverter. This is given by \((v_b^* + v_c^*)/3\) since the dual inverter cannot supply phase \(a\) voltage due to the open-phase fault. The reference voltage used for phases \(b\) and \(c\) of the dual inverter (mb**and mc**) are obtained by subtracting this zero sequence reference voltage from the values of \(v_b^*\) and \(v_c^*\), respectively. Figure 7 also shows examples of these references for a given \(v_b^*\) and \(v_c^*\).

The resulting motor phase \(b\) and \(c\) output voltages are shown in Figure 8 for a case where \(V_{blU} = V_{blL} = 400\) V. A five-level motor phase voltage waveform is observed, as is expected from the phase-shifted carrier modulator given in Figure 6. The highest voltage level is equal to \(V_{blU} + V_{blL} = 800\) V.

3.3 Simulation results

In this section, simulation results are presented for the system with parameters in Table 2 in the healthy and faulted cases (utilising the conventional and proposed remedial methods). In each case, the motor starts at an initial speed of 150 rad/s and is accelerated until it reaches the maximum achievable speed. No load torque was imposed on the motor for this simulation.

Figure 9 shows a comparison of the simulated motor speed, torque and \(i_d^*\). As predicted by the analytical comparison in Figure 4, the peak torque of the healthy drive system is three times higher than the faulted case.
**FIGURE 6** Modulator used in the proposed open-phase fault tolerant system

**FIGURE 7** Modulation waveforms for test case
The maximum torque ripple obtained for the healthy case is 4 Nm. This is lower than the maximum torque ripple obtained for either fault tolerant approach, which is 6 Nm for the conventional method of fault tolerance, or 13 Nm for the proposed method. However, the proposed method of fault tolerance is capable of achieving double the speed range of the conventional fault tolerant method.

Figure 10 shows the motor current and voltage waveforms in the healthy case at a simulation time where the motor was at the edge of the constant torque operating region (0.17 s).
Waveforms of the faulted case are shown for a simulation times corresponding to the edge of its constant torque region in Figure 11 (0.78 s). As expected, the motor phase voltage in the faulted case utilises a higher voltage level of 800 V, which is equal to $V_{bL} + V_{dL}$, giving rise to a five-level voltage waveform. In contrast, the phase voltages in the healthy case have a maximum voltage of 533 V or $2/3*V_{bL} + V_{dL}$.

4 | EXPERIMENTAL VERIFICATION

The analysis and simulations of the proposed control and modulation methods presented previously are validated in this section on an experimental test bench (Figure 12). The control and modulation algorithms described in Figures 5 and 6 were implemented digitally on a Texas Instruments TMSF28335 digital signal processor. Two 750 V, 800 A rated Infineon Hybrid kit Drive modules were used to construct the dual inverter. The TM4 HSM60 traction motor used in this study is also pictured in Figure 12. Parameters of this motor were provided earlier in Table 2.

To test the faulted system, a breaker S was placed in series with switching node A of the dual inverter. Opening this breaker would emulate the open-phase fault. Switching nodes HB and HB′ of the front end normally used for AC charging were connected together, as shown in Figure 13. This represents switch $S_L$ of Figure 1 being held closed during driving operation. During the healthy (three-phase) operation, the IGBTs connected to nodes HB and HB′ are gated off. Two bidirectional supplies were used to emulate the isolated batteries which would be connected to the DC links of the dual inverter. The voltages of these supplies were set to 100 V to ensure that the multilevel phase voltage waveforms expected from the proposed modulation method could be observed at lower motor speeds. This constraint existed because the dynamometer used for these experiments was not capable of operating at speeds significantly greater than 1000 rpm. The HSM60 traction motor was connected on the same shaft as a Kollmorgen servomotor, which in turn had a three-phase resistive load bank ($R_L$ in Figure 13) connected to its stator windings to serve as an electrical load.

4.1 | Operation in faulted mode

The experiments of this section are intended to investigate the performance of the dual inverter drive in the open-phase fault condition, after remedial operation has been initiated. Breaker S in Figure 13 was kept open during the experiments of this section. For the experiments of this section, the resistive load bank $R_L$ was set to a value of 14.3 Ω.

Figure 14 shows the motor phase $b$ and $c$ currents, phase $b$ voltage and motor speed reading when the drive received a speed step command of 1000 rpm, after initially cruising at 200 rpm. At the initial speed, the phase $b$ voltage only has a maximum voltage of $+/-100$ V. As the speed increases, this voltage reaches the $+/-200$ V level after a time of approximately 4 seconds. The proposed modulation methods ability to deliver greater voltage magnitudes to the open phase faulted traction motor at higher speeds is visible from the fact that higher voltage levels are requested as the speed of the motor (and hence requested modulation indices) increases. The
estimated load torque of the Kollmorgen motor during this test is shown as the bottom waveform of Figure 14.

A zoomed in view of the motor phase b and c currents and voltages at a speed of 1000 rpm is shown in Figure 15. The voltage waveform clearly shows the presence of the +/−200 V levels which are generated by the modulation of phase a of the dual inverter. The motor phase current waveforms exhibit some fifth and seventh harmonic components which are due to space harmonics in the TM4 traction motor.

Simulated waveforms obtained with the same operating conditions are shown in Figure 16. The motor phase b and c voltage waveforms exhibit the same shape as in the experimental results, indicating that the experimental implementation of the modulator is valid. The simulated phase current waveforms do not exhibit fifth and seventh harmonic content, as motor space harmonics are not considered in the simulation model utilised.

Experimental results of the motor phase b and c currents, along with the neutral current of the converter (the current flowing through the wire connecting the switching nodes A and A' of Figure 13) are shown in Figure 17. The peak current obtained in the neutral connection is 40 A. This is greater than the peak neutral current for the simulated case in Figure 16 (30 A), due to the presence of fifth and seventh harmonic content in the phase current waveforms.

4.2 Dynamic fault response

The experimental results of this section are intended to showcase the performance of the proposed remedial method
in responding to an open-phase fault and reducing the torque ripple of the faulted motor. Figure 18 shows the oscilloscope capture of an experiment where the motor was operating in three-phase healthy mode initially at a speed of 400 rpm, and was then subjected to an open-circuit fault in phase $a$. This fault was emulated by opening the breaker $S$ in Figure 13.
In this section, the load resistance $R_L$ was decreased to 3.3 Ω in order to increase the load torque provided by the Kollmorgen motor. The estimated load torque profile provided by the Kollmorgen motor during this test is shown as the bottom waveform of Figure 18.

Immediately after the fault occurred, the control and modulation strategy utilised by the dual inverter drive was still that of a healthy three-phase machine. A short time scale oscilloscope capture of this operation mode is shown in Figure 19. Clearly the phase shift between the phase $b$ and $c$ current waveforms during this period is 180 degrees, rather than the 60 degrees required by the remedial strategy. As a result, a torque ripple of 12 Nm was obtained, with significant low frequency oscillations present. The average torque

**Figure 15** Experimental waveforms of phase $b$ and $c$ current (CH2,C4) and voltage (CH1,CH3) at a steady state speed of 1000 rpm

**Figure 16** Simulated waveforms of phase $b$ and $c$ and neutral current, along with the phase $b$ and $c$ voltages at a steady state speed of 1000 rpm
FIGURE 17 Experimental waveforms of phase \(b\) (CH4) and \(c\) (CH2), along with current in neutral connector (CH3) at a steady state speed of 1000 rpm.

FIGURE 18 Experimental waveforms of phase \(a\) (CH1), \(b\) (CH2), \(c\) (CH3) currents and shaft torque (CH4) for the case where an open circuit fault occurred after the motor was running in three-phase healthy mode. The estimated speed based on phase current frequency is shown as the second from bottom waveform, while the estimated load torque is the bottom waveform.
produced by the motor decreases during this period, which causes a reduction in rotational speed. This is clearly visible from the rotational speed, which is plotted as the second from bottom waveform in Figure 18.

After a short period wherein the open phase fault was present but no remedial action was taken, the fault-tolerant method proposed by this study was initiated. A short time scale oscilloscope captures once steady state operation is regained is shown in Figure 20. The phase shift between phase $b$ and $c$ current waveforms is now 60 degrees, as required. The torque ripple has accordingly been reduced to only 4 Nm as a result. The average torque produced by the motor increases back to

![Figure 19](image1.png)

**Figure 19** Experimental waveforms of phase $a$ (CH1), $b$ (CH2), $c$ (CH3) currents and shaft torque (CH4) after an open-circuit fault occurred in phase $a$, but no remedial action was taken by the drive system.

![Figure 20](image2.png)

**Figure 20** Experimental waveforms of phase $a$ (CH1), $b$ (CH2), $c$ (CH3) currents and shaft torque (CH4) after an open-circuit fault occurred in phase $a$ and remedial action was taken by the drive system.
the prefault value, which causes the motor to accelerate back towards its prefault speed, as can be seen from the rotational speed waveform of Figure 18.

4.3 | Startup in faulted mode

Another operation mode of interest is the ability of the faulted system to restart after coming to rest. This performance is illustrated by Figure 21, which shows motor phase $a$, $b$ and $c$ currents along with measured torque when the motor is accelerated from standstill. The speed increases from zero to approximately 200 rpm which is reached at the end of the experimental record.

5 | CONCLUSION

This study has introduced an open-phase fault tolerant driving algorithm of a dual inverter-based traction drive with onboard integrated charging capability. This method can be used in cases where an open-phase fault due to power electronics or motor faults is detected and isolated. By using an external contactor to connect the switching nodes of the two half bridges normally used for AC charging, the maximum voltage vector achievable is doubled compared to conventional methods of achieving open-phase fault tolerance. This ultimately results in a doubled maximum speed for the open-phase faulted motor.

A novel control and carrier-based modulation method was developed to allow the use of this method. These algorithms were finally implemented on a digital signal processor and validated on an experimental setup, demonstrating their efficacy in allowing a larger phase voltage magnitude to be obtained compared to conventional fault-tolerant methods. The proposed algorithm was shown experimentally to decrease harmonic torque ripple produced by an open-phase faulted PMSM by a factor of 3.

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