Postfault Control of Scalar (V/f) Controlled Asymmetrical Six-Phase Induction Machines

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ABSTRACT With the accelerated development in semiconductor power devices along with the dictated rigorous reliability standards in some industrial sectors, the application of medium-voltage high-power multiphase induction machine with multiple three-phase windings is now considered as a leading technology in high-power safety-critical applications. This paper proposes a parameter-independent postfault control scheme for asymmetrical six-phase induction machine based on simple scalar V/f control, which can successfully ensure the most common postfault scenarios used in this respect, namely, equal stator copper loss and minimum copper loss modes. Moreover, the proposed controller can effectively be used in either open-loop or closed-loop speed control modes. The proposed controller is experimentally validated using a 1.5kW prototype induction machine. The effect of the neutral arrangement on the dynamic performance is also explored.

INDEX TERMS Asymmetrical six-phase, induction machine, scalar control, fault-tolerant, open phase, neutral configuration

I. INTRODUCTION

In high-power adjustable speed drives, the six-phase induction machines have usually been laboring in many industrial applications such as ship propulsion, and oil and gas applications, where high-reliability standards are mandatory [1]. Although the field-oriented control proved itself optimum to obtain a high torque response performance in the electric drive arena, it entails accurate machine parameter identification and complex controller structure. On the other hand, it seems fair to admit that the simple V/f scalar control is still widely being used especially if a simple control approach is required [2] and/or constant load is experienced such as fans, pumps, blowers, mixers, extruders, mills, hoist drives, excavators, test stands, ships’ drives, conveyor belts, etc. Although this control method is a rather simple while independent of machine parameters, the torque response under low-speed operation will likely be slow. Nevertheless, the scalar V/f control technique is still widely adopted in different multiphase-based industrial applications [3]-[5].

Several control techniques concerning the multiphase motors in general and the six-phase machines, in particular, have been proposed in the literature. Most of them were, in essence, excerpted from the conventional three-phase controllers but extended to regulate the higher phase order machines. In the standard Field Oriented Control (FOC), the torque/flux regulation is carried out using the current components of the fundamental \( \alpha \beta \) subspace [6]-[7]. The reference secondary current components of the secondary \( xy \) subspace are commonly set to zero under normal operation to avoid excessive circulating currents [7].

The extended multiphase version of some other controllers commonly proposed for three-phase machines has also been introduced such as Direct Torque Control (DTC) [8], [9] and Sliding Mode Control (SMC) [10]. Recently, the Finite-Control Set Model Predictive Control (FCS-MPC) was shown promise to provide a fast-dynamic response with relative flexibility to include some operational constraints [11]. It is worth mentioning that many other innovative controllers, which have been recently proposed in the literature, are still limited to three-phase machines [12]-[14].

When safety-critical applications come into play, a stable postfault drive operation would be a crucial need. The postfault control of six-phase machines has, therefore, been widely investigated in the literature [7], [15]-[17]. However, most employed controllers were mainly based on FOC along
with the Vector Space Decomposition (VSD) modeling approach. The employment of a postfault controller based on V/f scalar control was, therefore, limited to the three-phase [18] case. Insufficient work has been done to extend the concept to the five-phase case only [19]-[21].

Leaving aside how the reference torque producing αβ voltage/current components are derived in different postfault strategies, the secondary current components, and hence the remaining healthy phase currents, are commonly controlled to achieve an optimized postfault operation. The most commonly employed optimization criteria are the Equal Copper Loss (EL) and Minimum copper Loss (ML) control criteria. The former maximizes the torque/current ratio and is also denoted as Maximum Torque (MT) criterion [7], while the latter minimizes the machine losses [22]. The Full-Range Minimum Loss (FRML) postfault control strategy has recently been proposed in [23] to combine the operational merits of these two possible control modes over the whole permissible loading range. In all these postfault techniques, the machine is controlled to theoretically maintain a ripple-free torque profile under open phase conditions. This is carried out by ensuring a balanced αβ current components in a fashion similar to the healthy case. This assumption, of course, neglects the effect of the induced ripple torque components due to non-fundamental subspaces, which may be an acceptable assumption for distributed winding layouts. The reference current components of the xy secondary subspace and the zero-sequence subspace, in case of single neutral configuration (1N), are therefore excerpted from those of the fundamental αβ current component using predefined optimal constant gains [7]. Since reference current components of different subspaces are generally unbalanced, and in order to account for different machine asymmetries, Proportional-Resonant (PR) controllers, or equivalent dual-PI regulators in synchronous and anti-synchronous references frames, are typically used to ensure a proper current tracking to the reference sequence current components [7].

In this paper, a postfault controller based on conventional scalar V/f control is proposed for an asymmetrical six-phase induction machine. The proposed controller offers the following merits over standard FOC-based postfault controllers:

- The proposed controller is parameter-independent.
- It offers a lower number of current controllers than standard FOC-based controllers; hence, the tuning process will be simplified.
- The machine can either be operated in open-loop [24] or closed-loop speed control modes.

The machine VSD based model is first presented in Section II. The proposed postfault controller is then introduced in Section III. This controller is investigated for the two possible neutral arrangements: isolated neutrals (2N) or single neutral (1N) configurations; the latter was recently shown better in terms of fault-tolerant capability [7]. The experimental validation is then carried out in Section IV using a 1.5kW prototype six-phase induction machine. The effect of the neutral arrangement on the dynamic performance is also explored. Finally, the main conclusions are presented in Section V.

II. MACHINE VSD MODEL

The asymmetrical six-phase machine comprises two three-phase windings with a spatial phase shift of 30°. These two winding sets can either be configured with isolated or connected neutrals, as shown in Fig. 1. The machine is preferably modeled using the VSD approach [1], where the machine phase variables are decomposed into three decoupled subspaces, namely αβ, xy, and zero planes using Clarke’s transformation given by (1).

\[
\begin{pmatrix}
1 & -0.5 & -0.5 & 0.866 & -0.866 & 0 \\
0 & 0.866 & -0.866 & 0.5 & 0.5 & -1 \\
1 & 1 & -0.5 & -0.5 & -0.866 & 0.866 & 0 \\
0 & 0 & -0.866 & 0.866 & 0.5 & 0 & -1 \\
0 & 0 & 0 & 0.5 & 0.5 & 0 & 0 \\
\end{pmatrix}
\]

(1)

In induction machines, distributed windings are typically employed. Hence, primitive harmonic-free models for the xy and zero subspaces are usually assumed, while the machine dynamic performance mainly depends on the αβ subspace.

Based on the harmonic-free model assumption [25], the voltage equations for different subspaces are, therefore, given as

\[
v_{as} = R_s i_{as} + p \lambda_{as}
\]

\[
v_{bs} = R_s i_{bs} + p \lambda_{bs}
\]

\[
0 = R_s i_{ar} + p \lambda_{ar} + \omega_r \lambda_{br}
\]

\[
0 = R_s i_{br} + p \lambda_{br} - \omega_r \lambda_{ar}
\]

\[
v_{xs} = R_x i_{xs} + p \lambda_{xs}
\]

\[
v_{ys} = R_y i_{ys} + p \lambda_{ys}
\]

\[
v_{0+s} = R_0 i_{0+s} + p \lambda_{0+s}
\]

\[
v_{0-s} = R_0 i_{0-s} + p \lambda_{0-s}
\]

The flux linkage equations are given by

\[
\lambda_{as} = L_{s} i_{as} + L_{m} i_{ar}
\]

\[
\lambda_{bs} = L_{s} i_{bs} + L_{m} i_{br}
\]

\[
\lambda_{ar} = L_{m} i_{as} + L_{r} i_{br}
\]

\[
\lambda_{br} = L_{m} i_{bs} + L_{r} i_{ar}
\]

\[
\lambda_{xs} = L_{xy} i_{xs}, \quad \lambda_{ys} = L_{xy} i_{ys}
\]

\[
\lambda_{0+s} = L_{0s} i_{0+s}, \quad \lambda_{0-s} = L_{0s} i_{0-s}
\]

The machine parameters for different subspaces can easily be identified as given by [25]. The machine torque is decided
from the fundamental $\alpha$-$\beta$ subspace and is given as
\[
T_e = 3pL_m(i_{r\alpha}^s - i_{as}^s) \tag{8}
\]
where $p$ is the machine pole-pair number.

III. Postfault Operation

According to the available literature, the six-phase machine is usually controlled under postfault operation to generate rated airgap flux, while backward flux component is eliminated. The maximum achievable torque will, however, depend on the employed postfault scenario, as detailed in [7]. Theoretically, a pre-fault rated output, $P_o = 1pu$, can be guaranteed if the remaining healthy phase currents are controlled to follow the reference currents given in Table I under different postfault scenarios and different neutral configurations [7]. However, some phases will experience a notable current overload, which may yield excessive hot spots in the stator winding. In the literature, there are different presumable postfault scenarios that can be adopted based on application, as detailed in [1]. The most common scenario is to keep the currents in the remaining un-faulted phases at their pre-fault values, which yields a notable derating, while the machine stator loss, which is the main source of winding overheating, will be less than the rated value. In some other practical applications, such as electric traction, machines are usually featuring high overloading [26], and an efficient cooling system is usually embedded. Therefore, current overloading within a certain maximum period may be an option to increase the maximum achievable machine torque.

Under optimal current control, the $xy$ and zero-sequence current components are generally expressed in terms of the reference torque producing $\alpha\beta$ components as given by [7]
\[
\begin{bmatrix}
x_x \\
y_y
\end{bmatrix} = [K_{xy}] \begin{bmatrix}
i_{\alpha}^s \\
i_{\beta}^s
\end{bmatrix} \text{ and } \begin{bmatrix}
i_{0+}^s \\
i_{0-}^s
\end{bmatrix} = [K_0] \begin{bmatrix}
i_{\alpha}^s \\
i_{\beta}^s
\end{bmatrix} \tag{9}
\]

In the proposed scalar controller, the conventional V/f control is used to derive the $\alpha\beta$ voltage components of the fundamental subspace, as depicted by the controller block diagram shown in Fig. 2. Under healthy conditions, this will, in fact, be the same as a conventional three-phase case. This controller can, therefore, be used under either open or closed loop speed control modes. For the former, the reference frequency $\omega^*_{st}$ is set based on the desired no-load speed; as the machine is mechanically loaded, the machine speed drops [24]. Under closed loop speed control, the machine speed error is used to generate the rotor slip frequency, $\omega_{sl}$, which is, in turn, used to calculate the reference stator angular frequency $\omega^*_{st}$.

Under healthy conditions, the $\alpha\beta$ voltage components of the winding phase voltages will be equal to the reference $\alpha\beta$ components of the inverter leg voltages. Assuming a balanced stator winding, the corresponding $\alpha\beta$ current components should, therefore, be balanced. During open-phase(s) conditions, if the reference $\alpha\beta$ voltage components are kept balanced, the corresponding $\alpha\beta$ current components can still remain balanced as long as the other non-fundamental subspaces are properly controlled [7].

In standard IRFOC [7], the $\alpha\beta$ current components are derived from the reference $dq$ current component after applying Park’s transformation. The other reference sequence current components are, therefore, derived based on (9). The measured currents are transformed to their sequence components using (1). Then, the voltage components of each subspace are obtained using a pair of Proportional Resonant (PR) controllers; dual-PI regulators in synchronous and anti-synchronous reference frames can also be employed [7]. Thus, the FOC-based controller will entail three pairs of dual-PR controllers to ensure same controller structure under both healthy and postfault conditions while fully controlling all sequence current components.

In the proposed scalar control, since the reference $\alpha\beta$ voltage components are derived based on simple V/f scalar control, the corresponding $\alpha\beta$ current components are not directly controlled. They may, therefore, experience a degree of unbalance even though the $\alpha\beta$ components of the inverter leg voltages are balanced, which may be due to different winding asymmetries. This unbalance will be translated into inadequate reference non-fundamental current components derived based on (9). Hence, a preceding step before calculating the reference non-fundamental sequence current component is to extract the positive sequence current components, $i_{a\beta}^+$, from the measured $\alpha\beta$ components, $i_{a\beta}$. This is carried out using the “$i_{a\beta}^+$ Extraction” block shown in Fig. 2. The $i_{a\beta}^+$ current components are then used to derive the corresponding secondary sequence current components based on (9). Since these reference secondary components are generally unbalanced, the corresponding voltage components of the $xy$ and zero planes are decided using (PR) controllers similar to FOC-based controller and as detailed in [7]. The sequence voltage components are then transformed back to their phase values using inverse Clarke’s transformation. If the $xy$ and zero sequence controllers are disabled, the machine will be operating under open-loop current control mode. While, by enabling the optimal current controller, the secondary voltage components of different PR regulators outputs will force the remaining phase currents to follow the

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Image 67x549 to 74x587

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Image 85x626 to 91x651

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Image 101x586 to 105x611

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Image 112x633 to 118x658

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Image 117x592 to 121x617

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Image 122x593 to 124x616

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Image 128x583 to 134x607

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Image 134x591 to 136x07

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Image 155x562 to 166x611

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Image 205x626 to 211x651

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Image 232x633 to 238x658

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Image 250x581 to 259x605
optimal reference currents given by Table I. In this case, the total number of the required PR current controllers will be only four, or two dual-PR controllers, which is less by two than the standard IRFOC. The digital implementation of the PR controller with anti-windup is shown in Fig. 3, while the positive sequence extraction block used to extract the positive current components, $i_{αβ}^+$, is given in Fig. 4.

IV. EXPERIMENTAL RESULTS

A. EXPERIMENTAL SETUP

The proposed postfault scalar controller is verified using the six-phase prototype machine given in Table II. Two three-phase voltage source inverters connected to the same dc-link are used to supply the machine stator. The dc-link is constructed using a programmable dc source, while conventional PWM modulation at 5kHz switching frequency is employed. Hence, different low order space harmonics will cause a notable distortion in the current waveform, which will indeed be higher than the FOC-based control case. Moreover, the current waveform distortion under 1N arrangement is relatively higher than the 2N configuration. This is mainly due to the induced third harmonic air gap flux component due to zero-sequence excitation, which is, in nature, a pulsating flux component [25]. This component may also saturate the core and cause this notable distortion in the αβ current components. The 2N may, therefore, be preferable to avoid this notable current distortion.

**TABLE I**

<table>
<thead>
<tr>
<th>Variable</th>
<th>Healthy</th>
<th>Isolated Neutrals (2N)</th>
<th>Connected Neutrals (1N)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$i_{a1}$</td>
<td>$\angle 120^\circ$</td>
<td>$\angle 173^\circ-90^\circ$</td>
<td>$\angle 1.44^\circ-130.9^\circ$</td>
</tr>
<tr>
<td>$i_{a2}$</td>
<td>$\angle 120^\circ$</td>
<td>$\angle 173^\circ-90^\circ$</td>
<td>$\angle 1.44^\circ-130.9^\circ$</td>
</tr>
<tr>
<td>$i_{b1}$</td>
<td>$\angle 120^\circ$</td>
<td>$\angle 173^\circ-90^\circ$</td>
<td>$\angle 1.44^\circ-130.9^\circ$</td>
</tr>
<tr>
<td>$i_{b2}$</td>
<td>$\angle 120^\circ$</td>
<td>$\angle 173^\circ-90^\circ$</td>
<td>$\angle 1.44^\circ-130.9^\circ$</td>
</tr>
<tr>
<td>$i_{c1}$</td>
<td>$\angle 120^\circ$</td>
<td>$\angle 173^\circ-90^\circ$</td>
<td>$\angle 1.44^\circ-130.9^\circ$</td>
</tr>
<tr>
<td>$i_{c2}$</td>
<td>$\angle 120^\circ$</td>
<td>$\angle 173^\circ-90^\circ$</td>
<td>$\angle 1.44^\circ-130.9^\circ$</td>
</tr>
<tr>
<td>$i_{d1}$</td>
<td>$\angle 120^\circ$</td>
<td>$\angle 173^\circ-90^\circ$</td>
<td>$\angle 1.44^\circ-130.9^\circ$</td>
</tr>
<tr>
<td>$i_{d2}$</td>
<td>$\angle 120^\circ$</td>
<td>$\angle 173^\circ-90^\circ$</td>
<td>$\angle 1.44^\circ-130.9^\circ$</td>
</tr>
</tbody>
</table>

**FIG. 5**

2kHz. The PI and PR regulators are tuned via trial and error, and the controller gains are given in Table III.

**B. CURRENT WAVEFORMS UNDER DIFFERENT POSTFAULT MODES**

The proposed controller is investigated under different postfault scenarios as well as neutral point configurations. The healthy case is used as a benchmark reference case. Under fault conditions, phase $a_1$ is physically disconnected to emulate a single open line circuit fault. The machine is operated under closed loop speed mode, with the switch S set to position 2, while the reference speed is set to 1000 rpm. The constant gain matrices $[K_{xy}]$ and $[K_0]$ are selected from Table I based on the selected operating mode. Under healthy case, 2N configuration is used while the reference $xy$ and zero sequence current components are set to zero.

Fig. 5 depicts the measured current waveforms in per unit under rated load. It is clear that the proposed controller successfully ensures the desired reference optimal currents under all cases. The corresponding sequence current components are given in Fig. 6. The obtained current waveforms agree with a large extent to their optimal reference currents given in Table I.

Unlike FOC-based controllers, the $αβ$ current components are not directly controlled when scalar V/f control is employed. Hence, different low order space harmonics will cause a notable distortion in the current waveform, which will indeed be higher than the FOC-based control case. Moreover, the current waveform distortion under 1N arrangement is relatively higher than the 2N configuration. This is mainly due to the induced third harmonic air gap flux component due to zero-sequence excitation, which is, in nature, a pulsating flux component [25]. This component may also saturate the core and cause this notable distortion in the $αβ$ current components. The 2N may, therefore, be preferable to avoid this notable current distortion.

Switch S:
1: Closed loop speed control
2: Open loop speed control

FIGURE 3. Digital implementation of PR controller ($K_\alpha$, is the anti-windup gain).

FIGURE 4. Positive and negative sequence decomposition of the $\alpha\beta$ current components.

TABLE II
Prototype IM SPECIFICATIONS

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated phase voltage (V)</td>
<td>110</td>
</tr>
<tr>
<td>Rated frequency (Hz)</td>
<td>50</td>
</tr>
<tr>
<td>Rated Power (kW)</td>
<td>1.5</td>
</tr>
<tr>
<td>Rated speed (rpm)</td>
<td>1420</td>
</tr>
<tr>
<td>Rated phase current (A)</td>
<td>3.4</td>
</tr>
<tr>
<td>Pole number</td>
<td>4</td>
</tr>
</tbody>
</table>

TABLE III
Controller Parameters

<table>
<thead>
<tr>
<th>Controller Type</th>
<th>Parameter</th>
</tr>
</thead>
<tbody>
<tr>
<td>xy- and zero current controllers</td>
<td>$K_\mu = 5$, $K_1 = 50$, $K_t = 50$</td>
</tr>
<tr>
<td>Speed controller</td>
<td>$K_\mu = 0.1$, $K_1 = 1$</td>
</tr>
</tbody>
</table>

C. DYNAMIC RESPONSE

In this subsection, the machine dynamic response under either open-loop or closed loop speed controller is investigated. The proposed controller is first investigated during machine starting under free-running open-loop speed control. The machine is started at a specific constant voltage and frequency while mechanically unloaded. After the machine reaches its final no-load steady-state speed, a step load is then applied. This test can give a clear clue to the effect of the low order space harmonics of different subspaces on the machine dynamic response [27]. Under direct starting, the inrush starting currents may saturate the current measurement boards. Hence, this test is carried out under a lower phase voltage to avoid this problem.

Fig. 7(a) shows the starting speed profiles of the machine speed under different cases. While Fig. 7(b) shows the effect of step loading on the machine speed profile under the same cases. Under open-phase conditions with 2N arrangement, the machine dynamic response is very close to the healthy case, as clear from Fig. 7(a), which proves the effectiveness of the proposed controller. On the other hand, the machine under 1N connection exhibits a longer starting time due to the effect of the subsynchronous speed point caused by the induced third harmonic component [25]. This represents the second notable effect of a single neutral arrangement.
However, this sluggish response can be avoided by using a ramp starting or under closed-loop speed control.

Fig. 8 shows the RMS line currents for the five considered cases. Under healthy case, the line currents will be equal. However, a small difference in current magnitude between the two three-phase winding sets is noted, which is mainly due to different winding asymmetries. Under open-phase case, the MT mode corresponds to approximately equal line currents with magnitudes of 1.73 times and 1.44 times the healthy line current magnitude for the 2N and 1N configurations, respectively. Under ML mode, the line currents are generally different. The current magnitudes of some of these phases are within the healthy current magnitude; however, the current magnitudes of phases \(i_{a2}\) and \(i_{b2}\) are 1.8 times the healthy current magnitude case for the 2N configuration. Under 1N connection, the current \(i_{a2}\) is the phase current which has the maximum current magnitude (1.86 times the healthy current) for same rotor speed. Finally, the current distortion due to the induced third harmonic component under this connection will cause a relatively small deviation from the optimal reference currents, as clear from Fig. 8.

The machine response under closed-loop speed control is shown in Fig. 9. The machine is initially started at 500rpm while mechanically loaded with an initial load torque of 2 Nm. At 1s, the machine reference speed is step increased to 1000 rpm. Rated load is then applied after a while. The machine speed profiles under step-speed change and mechanical loading are shown in Fig. 9, which seem similar to the healthy case for all postfault operating modes. The corresponding profiles of the RMS line currents are given in Fig. 10. Clearly, by controlling the machine speed, the sluggish response shown in Fig. 7 and obtained under free-running open-loop speed control due to the effect of non-fundamental subspaces is entirely avoided, while the machine dynamic response is quite similar to the pre-fault (healthy) case.

**FIGURE 5.** Steady-state current waveforms under different postfault strategies. (a) Healthy. (b) MT2N. (c) ML2N. (d) MT1N. (e) ML1N.
Figure 6. Sequence current components under different postfault strategies. (a) Healthy. (b) MT2N. (c) ML2N. (d) MT1N. (e) ML1N.

Figure 7. Machine speed profiles under different cases and free-running open-loop speed control. (a) Starting period. (b) Step loading.
FIGURE 8. RMS line currents under free-running open-loop speed control. (a) Healthy. (b) MT2N. (c) ML2N. (d) MT1N. (e) ML1N.

FIGURE 9. Machine speed profile under different cases and closed-loop speed control. (a) Step speed from 500 rpm to 1000 rpm. (b) Step loading.
V. CONCLUSION

This paper proposed a postfault controller for scalar controlled asymmetrical six-phase induction machines. The proposed controller has the same structure under either healthy or fault conditions. It was proved that the machine could optimally be controlled during open phase fault condition under either open-loop or closed-loop speed control modes. The proposed scalar V/f-based controller does not require accurate information about the machine parameters while provides a robust and straightforward control in applications that do not require a high dynamic response. It also comprises a smaller number of current regulators than FOC-based controllers. Although the 1N connection offers a higher steady-state fault tolerant capability than the 2N connection under fault conditions, the former exhibits a sluggish starting response under open loop speed control due to the effect of the third harmonic components caused by zero sequence excitation. This effect has been explored using experimental investigation since it was commonly neglected in the available machine models. Under closed-loop speed control, the machine dynamic response can, however, be assumed identical for all cases. As far as the current waveform quality is concerned, the 1N connection generally corresponds to a higher distortion in the current waveforms, which is mainly due to the third harmonic induction caused by zero sequence excitation. This effect can, however, be compensated, if needed, by employing a multiple resonant current controller structure, which will, of course, entail a higher number of current regulators.

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FIGURE 10. RMS line currents under closed-loop speed control. (a) Healthy. (b) MT2N. (c) ML2N. (d) MT1N. (e) ML1N.
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