Investigation of Fault Modes of Voltage-Fed Inverter System for Induction Motor Drive

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Abstract—The reliability of power electronics systems is of paramount importance in industrial, commercial, aerospace, and military applications. The knowledge about the fault mode behavior of a converter system is extremely important from the standpoint of improved system design, protection, and fault tolerant control. This paper describes a systematic investigation into the various fault modes of a voltage-fed PWM inverter system for induction motor drives. After identifying all the fault modes, a preliminary mathematical analysis has been made for the key fault types, namely, input supply single line to ground fault, rectifier diode short circuit, inverter transistor base drive open, and inverter transistor short-circuit conditions. The predicted fault performances are then substantiated by simulation study. The study has been used to determine stresses in power circuit components and to evaluate satisfactory post-fault steady-state operating regions. The results are equally useful for better protection system design and easy fault diagnosis. They will be used to improve system reliability by using fault tolerant control in the next phase of our work.

I. INTRODUCTION

THE reliability of power electronics equipment is extremely important in general industrial applications. A common fear that power electronics equipment is not adequately reliable is preventing its widespread application, and thus, the potential of power electronics for industrial productivity and energy saving is not being fully explored. It has been predicted [1] that voltage-fed inverters will eventually replace all other types of converters in the near future. However, the reliability of voltage-fed inverters is not adequate. In a voltage-fed system, besides the electrolytic capacitor in the dc link, the power semiconductors and their control electronics remain a weak link. A commonly practiced method of improving reliability is to design the power circuit conservatively. A second method is to have parallel redundant operation of components or circuits. Evidently, both these techniques are expensive and can be justified only for high-reliability applications. As an alternative to the redundancy technique, fault tolerant control has been proposed in [3] and [6]. The idea is to modify the drive control algorithm during a fault in such a way that the faulty drive continues to run in a degraded mode. The authors used a brushless dc motor and a special purpose electrically isolated multiwinding induction motor for their studies. McMurray [5] did a simulation study on a current-fed inverter induction motor drive. However, systematic studies on converter faults in a voltage-fed inverter induction motor drive and their remedial strategies have not yet been reported in the literature.

In this paper, the fault mode behavior of a voltage-fed inverter induction motor drive has been investigated for the key converter fault types. A number of practical fault modes have been identified. From these, a few selected fault types have been mathematically analyzed, and then the predicted fault performances have been verified by simulation. The standard stationary reference frame machine $d-q$ model has been extended to incorporate machine saturation and unbalance (e.g., open-phase winding) due to these faults. The machine faults are not considered in the present study. Although the simulation study assumes an open-loop volts/hertz control of the drive, the results can easily be applied for other types of control. The results are useful for optimal protection system design, determination of post-fault operating conditions, and design of fault tolerant control.

II. CONVERTER FAULTS AND PROTECTION

A voltage-fed inverter induction motor drive system, as shown in Fig. 1, can develop various types of faults that can be classified as follows:

- Input supply single line to ground fault ($F_1$)
- Rectifier diode short-circuit fault ($F_2$)
- Earth fault on dc bus ($F_3$)
- DC link capacitor short-circuit fault ($F_4$)
- Transistor base drive open fault ($F_5$)
- Transistor short-circuit fault ($F_6$)
- Line to line short circuit at machine terminal ($F_7$)
- Single line to ground fault at machine terminal ($F_8$)

Fails may also occur inside this machine. Some common machine faults caused by winding insulation failure due to excessive voltage or current stress are practically eliminated when inverter power supply is used. This is because the line voltage surges are absorbed at the converter input, and inverter overcurrent protection limits the machine current. The problem of broken rotor bars mainly due to direct on-line start is practically eliminated by soft starting with an inverter. For these reasons, machine faults are not considered in our study. Again, the possibil-
Fig. 1. A voltage-fed PWM inverter system indicating the possible failure modes.

Fig. 2. A typical protection system for a voltage-fed inverter drive.

The possibility of multiple faults occurring at the same instant is very remote, and thus are not considered. The secondary fault as a consequence of primary fault has also been ruled out from our consideration because an efficient protection circuit will prevent any such possibility.

The protection system in a commercial drive is usually designed conservatively so as to prevent any damage in the converter system. Fig. 2 shows a typical protection system in a voltage-fed inverter drive. It includes protection against ground fault, dc link overvoltage/undervoltage, and inverter overcurrent. For one or more of such faults, the transistor base drives are inhibited and the magnetic contactor at the input is opened. The input circuit breaker (CB) trips for steady overcurrent to the converter. The input fuses blow for short-circuit fault of diode rectifier or dc link capacitor. The inverter input fuse protects the rectifier and filter capacitor against shoot-through fault in the inverter. The machine's overtemperature is protected by a breaker activated by thermal relay as shown. The protection system, as discussed above, often fails to critically judge the seriousness of a fault, and therefore may not be adequate. For example, consider the transistor base drive open fault (F1) shown in Fig. 1. If the machine is lightly loaded, the fault may not cause the phase current to exceed the safe limit. However, the stress on transistor (Q1) may become excessive because it now carries the entire phase current. Such observation confirms that a systematic fault mode investigation in a converter system is essential not only for fault tolerant control development, but also for design of optimal protection strategy.

III. ANALYSIS OF FAULT MODES

In this section, only the following key types of faults will be considered for our study:
- Input supply single line to ground fault (F1)
- Rectifier diode short-circuit fault (F2)
- Transistor base drive open fault (F3)
- Transistor short-circuit fault (F4)

It can be shown that the effect of a fault occurring across any other device or phase is symmetrical, and therefore only the device or phase under fault, as indicated, will be considered. It will be shown later that the input supply line to ground fault (F1) is equivalent to the input supply single phasing fault, i.e., for the fault F1 shown in Fig. 1, the diode rectifier operation is the same as if the phase a fuse is blown. The other fault type shown in Fig. 1 will demand an immediate shutdown of the drive; therefore, the question of fault modes does not arise. These faults are not considered in our study.

A. Input Supply Single Phasing Fault (F1)

When fault (F1) occurs, the phase a voltage to the diode rectifier goes to zero. In this condition, the rectifier diodes can conduct only if the supply line voltage exceeds the dc link voltage. Fig. 3 shows the supply phase and line voltages, dc link voltage, and the diode currents for this fault. Since phase a voltage is zero, it is not shown in the figure. During the possible conduction intervals of the healthy rectifier (i.e., D1,D2, D3,D4, D5,D6, D7,D8), conduction occurs only during the D2,D4 and D3,D6 intervals when the line voltage $v_{ab}$ tends to exceed the dc link voltage. Evidently, the conduction mode is equivalent to that of a single-phase rectifier with healthy legs D3,D6 and D2,D4. This concludes that the fault F1 is equivalent to operation with a blown fuse in phase a. It will be shown later that a similar single phasing mode can be induced for rectifier diode short-circuit fault.

Since the line inductance is usually small, the rectifier operates in discontinuous conduction mode as shown in Fig. 3. The dc link filter capacitor is normally designed such that the ripple voltage is typically less than 5% of the average voltage. Preliminary analysis and simulation study indicate that in a well-designed system, the average dc voltage decreases only marginally due to single phasing. For an open-loop volts/hertz controlled PWM inverter drive (assumed in our system), this means that the fundamental voltage across the machine will remain practically unaltered. Therefore, the average power drawn by the machine, and consequently the average dc link current, remains constant. In a single-phase operation mode, each diode carries 50% more average current than that of a three-phase mode. Due to a poorer form factor of current wave, the rms current sharing is larger than 50%.

Although the average voltage variation is small in single phasing, the ripple voltage in the dc link may increase
substantially. Fig. 3 indicates the dominance of 120-Hz ripple in the dc link voltage. Neglecting the high-frequency ripple components, the dc voltage can be written in the form

\[ v_d = V_d + \hat{v}_d \sin (\omega_d t + \phi), \]  

where \( V_d \) is the average voltage and \( v_d \) and \( \omega_d \) are the peak voltage and angular frequency of the ripple component, respectively. For a sinusoidally modulated PWM inverter, the fundamental output voltages can be given as

\[ V_{as} = v_m \sin (\omega t), \]

\[ V_{bs} = v_m \sin \left( \omega t - \frac{2\pi}{3} \right), \]

\[ V_{cs} = v_m \sin \left( \omega t + \frac{2\pi}{3} \right), \]

where \( v_m \) and \( \omega \) are the amplitude and the angular frequency of the voltage waves, respectively. The dc link ripple voltage will create amplitude modulation of the alternating component of the output voltage waves which can be expressed as

\[ \hat{v}_{as} = 0.5mv_d \sin \omega t + 0.5m\hat{v}_d \sin \omega t (\omega_d t + \phi), \]

\[ \hat{v}_{bs} = 0.5V_d \sin \left( \omega t - \frac{2\pi}{3} \right) + 0.5m\hat{v}_d \sin \left( \omega t - \frac{2\pi}{3} \right) \cdot \sin (\omega_d t + \phi), \]

\[ \hat{v}_{cs} = 0.5V_d \sin \left( \omega t + \frac{2\pi}{3} \right) + 0.5m\hat{v}_d \sin \left( \omega t + \frac{2\pi}{3} \right) \cdot \sin (\omega_d t + \phi), \]

where \( m \) is the modulation index. Eqs. (5), (6), and (7) can be written in the form

\[ \hat{v}_{as} = \frac{mV_d}{2} \sin \omega t + \frac{m\hat{v}_d}{4} \left[ \cos (\omega_d t + \phi) - \cos (\omega_d t + \omega + \phi) \right] \]

\[ \hat{v}_{bs} = \frac{mV_d}{2} \sin \left( \omega t - \frac{2\pi}{3} \right) + \frac{m\hat{v}_d}{4} \left[ \cos (\omega_d t + \phi + 2\pi/3) - \cos (\omega_d t + \omega - 2\pi/3) \right], \]

\[ \hat{v}_{cs} = \frac{mV_d}{2} \sin \left( \omega t + \frac{2\pi}{3} \right) + \frac{m\hat{v}_d}{4} \left[ \cos (\omega_d t + \phi + 2\pi/3) - \cos (\omega_d t + \omega - 2\pi/3) \right]. \]

From (8)–(10), it is evident that the frequency components \( \omega_d + \omega \) have positive phase sequence, whereas \( \omega_d - \omega \) component has negative phase sequence. If the flux generated by harmonic voltages is neglected, a pulsating torque will be generated due to interaction between the fundamental frequency flux and the harmonic frequency currents. Since the phase sequence of the harmonic frequencies \( \omega_d + \omega \) and \( \omega_d - \omega \) are opposite, a pulsating torque of frequency \( \omega_d \) (120 Hz in this case) will be contributed by both of them. This is later verified by simulation.

B. Rectifier Diode Short-Circuit Fault (P_f)

A short-circuit fault in a rectifier diode causes excessive current stress on the line fuses. In general, this type of fault may cause one or more input fuses to blow. If the fuse in the faulty phase blows first, the rectifier will continue to operate in single-phase mode. However, if one of the healthy fuses blows first, the fault will continue until the fuse in the faulty phase blows. This will cause total interruption of power to the rectifier. The sequence in which the fuses blow depends on the fault initiation point on the supply voltage waves. This will be explained in the following analysis.

Let us consider that the fault is initiated at the point F as shown in Fig. 4(a). At this instant, diodes D_1,D_2 were conducting as indicated in Fig. 5(a).

Although the fault occurs at F, the fault current starts building up only at the point A when the diode D_3 becomes forward-biased as shown in Fig. 5(b). The equation for fault current \( i_f \) in this mode (mode 1) is

\[ 2L \frac{di_b}{dt} = v_b - v_a = v_{ba}, \]
where $L$ is the phase inductance. The supply voltage can be given as

$$v_a = \sqrt{\frac{2}{3}} V_L \cos \omega t,$$

$$v_b = \sqrt{\frac{2}{3}} V_L \sin \left( \omega t - \frac{\pi}{6} \right),$$

$$v_{ba} = \sqrt{2} V_L \sin \left( \omega t - \frac{\pi}{3} \right),$$

where $V_L$ and $\omega$ are line rms voltage and supply angular frequency, respectively. Therefore, from (11) and (14), the fault current in phase $a$ or $b$ can be solved as

$$i_a = i_b = \frac{V_L}{\sqrt{2} \omega L} \left(1 - \cos \left( \omega t - \frac{\pi}{3} \right)\right),$$

where $\omega t \geq \pi/3$.

In mode 1, the voltage across $D_a$ is $v_c - 0.5(v_a + v_b)$, i.e., $1.5v_c$. Therefore, $D_a$ will become forward-biased and will start conduction at the point B shown in Fig. 4(a). In this mode (mode 2), three-phase symmetrical short circuit occurs across the supply phases as indicated in Fig. 5(c).

Fig. 4. Fault mode voltage and current waves for rectifier diode short-circuit fault.

The corresponding currents equations can be given as

$$L \frac{di_b}{dt} = v_b = \sqrt{\frac{2}{3}} V_L \sin \left( \omega t - \frac{\pi}{6} \right),$$

$$L \frac{di_c}{dt} = v_c = \sqrt{\frac{2}{3}} V_L \sin \left( \omega t - \frac{5\pi}{6} \right),$$

$$i_a = i_b + i_c,$$

where $\omega t \geq 5\pi/6$. $V_L$ and $L$ are the line rms voltage and the line inductance per phase, respectively.

The fault current waves for the three phases are sketched in Figs. 4(b), 4(c), and 4(d). Since phase $a$ current is the same as that of phase $b$ in mode 1, the corresponding $i^2t$ stress on line fuses will be the same. We define the $i^2t$ stress on fuse in mode 1 (only two diodes conducting) as critical $i^2t$. The fault current profiles on phase $a$ and phase $b$ indicate that if the fuses survive in mode 1, then the phase $a$ fuse will definitely blow in mode 2, causing single-phase rectifier operation. However, it can be shown that if the fault is initiated beyond the negative zero crossing point (b) of the $v_b$.
wave, the phase $b$ will not conduct until the point $A$ of the next cycle. As the fault initiation point approaches the point $C$, the fault current contributed by phase $a$ and phase $c$ and the corresponding critical $i^2t$ is reduced. If the fault develops after the point $C$, the fault current does not build until the point $A$ of the next cycle. Therefore, we can conclude in a qualitative manner that a fault occurring near the point $B$ or $C$ will cause only the phase fuse to blow, resulting in single-phase operation as discussed in Section III-A. However, if a fault occurs outside these regions, faulty or healthy phase fuse will blow randomly. As discussed before, blowing a healthy phase fuse will cause total disruption of power. These phenomena will be investigated further in simulation study.

C. Transistor Base Drive Open Fault ($F_3$)

The inverter transistors are normally controlled by isolated base drive amplifiers. Malfunctioning of one of these units can result in a missing base drive as indicated by an open switch $F_3$ in Fig. 1. Since the transistor $Q_3$ is now inoperative, the phase $a$ of the machine is connected to the positive dc bus through the bypass diode $D_1$. The machine phase $a$ voltage is then determined by the polarity of current and the switching pattern of transistor $Q_3$.

Assume that the drive system attains a new steady state following a transistor base drive open fault and also assume that the base drive control pattern remains the same before and after the fault. If the small stator resistance is neglected, the applied voltage across the machine will be supported by leakage inductance and the magnetizing inductance. This means that in a steady-state post-fault condition, the machine cannot support any dc voltage. If a dc offset is impressed in the beginning, the dc current will build up until at steady state the offset voltage is biased off. The healthy phase legs $b$ and $c$ cannot generate any dc voltage across the machine. For phase leg $a$, if the polarity of phase current $i_{as}$ is positive, the phase voltage will be clamped to the negative bus. On the other hand, if $i_{as}$ is negative, the voltage will be that of the negative bus when $Q_3$ switches on, and that of the positive bus when $Q_3$ is off and $D_1$ is on. With the sinusoidal base drive pattern of $Q_3$, the phase $a$ voltage dc offset can be zero only if the polarity of the current is biased to the negative value for the whole cycle. Since the phase voltages are balanced with the sinusoidal PWM modulation before and after the fault, the phase currents will be balanced sinusoidals (neglecting ripple) with dc offset after the fault as indicated in Fig. 6.

The above conclusion is valid only under the assumption of magnetic linearity and infinite rotor inertia. The dc offset current in phase $a$ will be equally divided between the phase $b$ and phase $c$ because of symmetry in these phases. With the phase current waves shown in Fig. 6, the developed torque in the machine can be analyzed by considering the effect of ac and dc current components separately and applying the superposition principle. The ac component will contribute motoring torque whereas the dc component will produce braking torque. It can be shown that the net torque is given by

$$ T_{net} = 3 \left( \frac{L_m}{L_r} \right)^2 \left( \frac{1}{\omega_s} - \frac{1}{\omega_r} \right)^2, \quad (19) $$

where $L_m$, $L_r$, and $R_r$ are per-phase magnetizing inductance, rotor self-inductance, and rotor resistance of the machine, respectively, $\omega_s$ and $\omega_r$ represent rotor frequency and slip frequency (in elec. rad/s). $l$ is an ac component (rms) of phase currents.

Since $\omega_r < \omega_s$, the net torque will remain positive. A few additional comments can be given for post-fault steady-state condition as follows.

i) Interaction between the dc component of stator flux and the fundamental frequency rotor current will cause fundamental frequency pulsating torque that can be particularly harmful at low operating frequency and low shaft inertia.

ii) The injected dc offset in the machine phase currents worsens the current stress of the inverter switching devices. Since the motoring torque is generated by the ac component of the phase current only, the maximum average torque capability of the drive is substantially reduced in comparison to normal healthy condition due to additional loading of the devices.

All of the above effects will later be demonstrated by simulation study. However, the current waves will be somewhat different from Fig. 6 due to nonzero stator resistance and finite rotor inertia.

D. Transistor Short-Fault ($F_5$)

This type of fault puts extreme stress on the inverter switching devices and therefore requires immediate attention of the protection circuit. Unfortunately it is also a commonly occurring fault. A transistor might fail due to
current stress or voltage stress. The failure due to current occurs when the device is carrying load current whereas the voltage failure occurs when the device is switching off the load current or blocking forward voltage. For such a fault, base drive to the healthy transistor of the same leg should be immediately suppressed in order to prevent a shoot-through fault. Even then, the phase current will continue to grow if base drive to other healthy devices is kept unaltered.

For the fault shown, the machine phase \( a \) will be tied to the positive bus. Assuming that the base drive of the healthy transistors remains unaltered (except that of \( Q_4 \)), the phase voltages with respect to the negative bus can be given as

\[
v_{ds} = v_{ds},
\]

\[
v_{bs} = 0.5v_{dd} + \tilde{v}_{bs},
\]

\[
v_{cs} = 0.5v_{dd} + \tilde{v}_{cs},
\]

where \( v_{bs} \) and \( v_{cs} \) are the respective alternating components of voltage. Substituting (20)–(22), the corresponding stationary \( d-q \) stator voltages can be derived as

\[
v_{pq}^* = \frac{v_{dd}}{3} - \frac{1}{3}(\tilde{v}_{cs} + \tilde{v}_{bs}),
\]

\[
v_{dq}^* = \frac{1}{\sqrt{3}}(\tilde{v}_{cs} - \tilde{v}_{bs}).
\]

Fig. 7 shows the machine \( d-q \) equivalent circuits in stationary frame with these applied voltages. If the fault persists, the dc component of the \( q \)-axis applied voltage will introduce a dangerously large dc component (limited only by the small stator resistance) into the \( q \)-axis current \( (i_q) \). In practice, as mentioned before, the drive overcurrent protection will suppress base drive to all the healthy devices as soon as the faulty phase current exceeds the threshold value. With the five diodes of the inverter remaining healthy, it can be shown that for positive \( i_{ds} \), the phase \( b \) and \( c \) will be shorted through the diodes \( D_3 \) and \( D_2 \) causing symmetrical short circuit across the machine. The speed and the residual flux linkage in the machine will create counter emf and the machine will experience dynamic braking (energy being dissipated in stator and rotor resistances) until the flux decays to zero. These phenomena were investigated in detail in the simulation study. Note that any overcurrent that may be generated in the process flows through the freewheeling diodes which have higher short time current capability. The healthy transistors and their snubbers are not subjected to any stress. Although base drive suppression protects the healthy devices from overcurrent, it is not possible to reenergize the drive without opening the faulty phase. If the faulty phase can be opened quickly, single-phase motor operation can be continued at partial load with degraded performance. These phenomena will be investigated later in detail.

IV. SIMULATION STUDY OF CONVERTER FAULTS

A systematic simulation study was conducted on a typical volts/hertz controlled drive system under the four fault conditions discussed in the previous section. The motor drives a constant torque type load. The simulation study substantiates the preliminary analysis, and determines quantitatively the voltage and current stresses on the devices in transient and steady-state conditions. These data will be helpful for fault diagnosis and evaluation of satisfactory operating regions of the drive on the torque–frequency plane. The system under simulation study is indicated in Fig. 1, except that only four specific faults, \( F_1, F_2, F_3, \) and \( F_4 \), have been selected for simulation. The simulation parameters for the system are shown in Table I.

The modeling of a faulty rectifier and inverter is somewhat complex because the symmetry is lost at fault condition. Both the converters were modeled as ideal switching networks in simulation. To determine the effective circuit topology at a particular instant, the voltage across each switch and the status of its base drive signal were examined. Depending on the current circuit topology, the differential equations were described and solved at each instant of simulation. The machine modeling under fault condition is equally important. The machine was modeled in a stationary reference frame \( d-q \) axis using the stator and rotor flux linkages as state variables. The magnetic saturation was modeled by considering magnetizing reactance as a function of total flux linkage according to a saturation curve. Although a more detailed saturation model is desirable for accurate prediction of machine performance with nonsinusoidal voltages, the simplified model helps prediction of approximate performance for dc and subharmonic saturation. In addition to the converter and machine models, the simulation system incorporates two additional modules that calculate rms and average values of several system variables. A simplified transient thermal model of the inverter system with a heat sink was used to estimate the peak device junction temperature when the faulty drive operates at steady state. The IBM personal-computer-based Simmon language was used for the simulation study.

Fig. 8 shows post-fault steady-state simulation waveforms for the input supply single line to ground fault (\( F_1 \))
The load torque was adjusted so that the resulting single-phase rectifier diodes carry the rated rms current. Consequently, this operating point falls on the steady-state safe operating boundary of the faulty drive configuration. The average dc link voltage does not change appreciably after the fault, but there is a large increase of the ripple voltage. This voltage ripple contributes to large current and torque ripple as shown in Fig. 8. The dominating ripple frequency of machine current is 70 Hz (ω/2 = ω) and the corresponding torque ripple is at 120 Hz (ω - d) as explained in Section III-A. The speed ripple contributed by the torque ripple depends on machine inertia and is shown in Fig. 8(c). The fault simulation was repeated for the machine operating frequencies of 10, 20, 30, 40, and 60 Hz. In each case, the post-fault machine loading was adjusted so that the rectifier diodes always carry the rated rms current whereas all other devices carry current within their ratings. This set of operating points defines a new boundary of safe operating area of the drive-in torque-frequency plane. Fig. 9(a) shows safe operating regions for healthy and faulty drives. Fig. 9(b) shows the diode rms currents as the operating point on the torque-frequency plane moves along ABC and ABD and compares with the pre-fault current. The corresponding torque ripples in pre-fault and post-fault conditions are shown in Fig. 9(c). This difference is a quantitative measure of the performance degradation due to this fault.

From this simulation study, we conclude that the faulty drive can safely operate over a significant area of the healthy drive's operating region in the torque-frequency plane. The large dc link voltage ripple causes some torque and speed pulsation, but these effects can be minimized by overdesigning the filter capacitor or by using a sophisticated PWM algorithm to reject dc link voltage ripple. In many applications, such as pumps and fans, a small speed ripple will be preferred to a complete shutdown of the drive.

Next, the rectifier diode short-circuit fault (F1) was investigated. The short circuit across the diode D, as shown in Fig. 1 was studied. Fig. 10 shows the input line current waves and dc link voltage for this fault and the effect of subsequent clearing. The exact instant of fault clearance was determined by comparing the i²t stress on the phase fuse with its melting i²t. The capacitor voltage (vC) after fault clearing is determined by the net charging/discharging current flowing through the capacitor. In the beginning, the voltage decreases because of the loading, but the large stored energy in the line inductances creates the voltage overshoot, as indicated. The fault was simulated at different points along the line voltage wave. The critical i²t as defined in Section III-B was computed in each case and plotted in Fig. 11. The fuse melting i²t is also shown on the same figure.

As predicted earlier, there are two regions −(5π/9) ≤ θ ≤ 10π/9 and 4π/3 ≤ θ ≤ 5π/3 where critical i²t is less than fuse melting i²t. Therefore, if the fault occurs within these intervals, the faulty phase fuse will blow and the drive will continue to run with a single phase rectifier
as discussed before. In this case, the performance shown in Figs. 8 and 9 will remain valid. However, if the fault occurs outside these intervals, the fuse in either the faulty or healthy phase may blow first. If a healthy phase fuse blows first, fault current will take two paths until blowing another phase fuse completely interrupts the power. The zones where critical $i^2t$ falls below the fuse melting $i^2t$ are widened if the line inductances are higher. Therefore, higher line inductance will make this fault self-clearing in most cases.

The results for base drive open fault for transistor Q1 are shown in Figs. 12 and 13. If the current waves in Figs. 12(a) and 12(b) are compared with the corresponding waves in Fig. 6, the analogy is obvious except here the faulty current wave is clamped to zero during part of the cycle. This difference is due to nonzero stator resistance that was neglected in the analysis. Nevertheless, the conclusion regarding the dc injection into machine phase currents is apparent from these waveforms. Large fundamental frequency torque pulsation is particularly harmful at low frequency due to a possible mechanical resonance problem. Another problem due to the dc offset in machine current is the unequal current stress in the upper and lower transistors of a phase leg. Therefore the rms values of machine phase currents do not directly correspond to the worst thermal stress on the devices. This point should be considered while designing a drive protection system. On the basis of the maximum allowable thermal stress on a transistor, Fig. 13 shows the steady-state safe drive operating area for healthy and faulty drives. The boundary line $A'B'C'$ on the torque–frequency...
plane was obtained by simulation at different operating points such that the transistor junction temperature remains within the safe limit. It is interesting to note that the faulty drive permits constant torque generation above a minimum threshold frequency. However, the available torque rapidly decreases as the operating frequency is reduced. This can be explained as follows. At a low operating frequency, the frequency of torque pulsation is also low. This causes a large fluctuation of the machine speed. The speed fluctuation generates new current harmonics that further increase inverter loading. Therefore, the maximum available torque diminishes at low speed. The simulation study was carried out down to the minimum output frequency of 10 Hz. The dotted portion of AB' in Fig. 13(a) simply indicates the trend of the profile. The rms torque ripple and the corresponding speed ripple as a function of the operating frequency are indicated in Figs. 13(b) and 13(c). For the load inertia considered in the simulation study, the speed ripple appears to be considerably high. In low inertia drive where such speed ripple is not permitted, the operating zone has to be severely restricted. Of course, the present study limits the operating zone only on the basis of inverter capability.

As indicated before, the short circuit of an inverter transistor is a commonly occurring fault. In order to avoid a shoot-through fault, base drive to the healthy transistor of the same phase leg should be immediately suppressed. This is not possible, however, if a transistor is conducting and the fault occurs in the blocking transistor in the same phase leg. The large short-circuit current, in this case, will destroy the healthy device. Short circuit across the dc link will be ultimately cleared by the inverter input fuse shown in Fig. 2. In our discussion, we will not consider this case. Instead, we will assume that base drive to a healthy transistor of the same phase leg is suppressed immediately after the fault. Fig. 14 shows the behavior of machine current and the corresponding speed due to transistor short-circuit fault.

Two situations have been considered in our study. The first situation corresponds to inhibiting the base drive of the healthy device in the faulty leg but maintaining the usual base drive for other devices. In such a case, the phase currents rise to a dangerous level due to dc offset as discussed before. The machine speed drops sharply due to large braking torque. In practice, the instantaneous overcurrent protection of the inverter will immediately suppress all the base drive. This is the second case in the simulation study. As is evident in Fig. 14, the phase currents are well restricted in this case. The study was repeated for faults at different points of the machine phase voltage waves. It was observed that in some operat-
Fig. 14. Simulated waves for transistor short-circuit fault at normal base drive and inhibited base drive condition. (Load torque = 20 N-m; supply frequency = 50 Hz). (a) Machine phase a current ($i_a$). (b) Machine phase b current ($i_b$). (c) Machine speed ($\dot{\theta}$ = 3000 rpm).

ing points, the machine currents can exceed the safe level of the switches. However, this current now flows through bypass diodes that have larger short time current carrying capability. The protection system, therefore, saves the inverter from any further damage. The drive is deenergized and cannot be restarted without opening the faulty phase. If a disconnect device (e.g., solid-state circuit breaker) is used at the inverter output which can be opened at the first zero crossing of the faulty phase current, single-phase inverter operation may continue at a reduced load. This area will be investigated in the next phase of our study.

V. CONCLUSION

The paper describes systematically the effect of different types of fault in a voltage-fed PWM inverter induction motor drive that uses the open loop volts/hertz speed control method. The machine internal faults are, however, excluded from the study. The important fault types are identified in the beginning that is then followed by preliminary analysis of the selected fault types. Then, systematic simulation study has been made to substantiate the analytical study. The extensive simulation study permits one to define the zones of operation where the drive can continue to operate safely in a degraded mode. This is extremely important in high-reliability process control applications. The study of fault performance of the drive system is extremely complex. The complexity is further aggravated due to a modeling problem of the machine under saturation and unsymmetrical condition. Again, the unsymmetry of converter operation makes the simulation program development somewhat difficult. However, such a study is extremely important to determine the device stress, to optimally design the protection system, and to determine post-fault drive operating capability. The results of this study can be extended to other converter configuration or drives with other types of control. It also provides insights for fault tolerant control development of the drive that will be undertaken in the next phase of our study.

REFERENCES


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