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VULNERABILITY OF PULSE-WIDTH-MODULATED ADJUSTABLE SPEED DRIVES TO UTILITY CAPACITOR SWITCHING

by

Van E Wagner

A THESIS

Submitted to
Michigan State University
in partial fulfillment of the requirements
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ABSTRACT

VULNERABILITY OF PULSE-WIDTH-MODULATED
ADJUSTABLE SPEED DRIVES TO
UTILITY CAPACITOR SWITCHING

by

Van E Wagner

Surges of less than 200% magnitude caused by utility capacitor switching can disrupt PWM drives. In this thesis, the disruption mechanism of the drive is investigated by SPICE simulations and laboratory experimentation. Field tests are analyzed to understand the generation and propagation of the capacitor switching transient through the power network. It is shown that drive dc filter inductor transient saturation is the most significant mechanism of drive disruption. It is further shown that the power network bandwidth determines whether the switching transient propagates from the source to the drive.

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CHAPTER 1

INTRODUCTION

The proliferation of power electronics has introduced several incompatibilities between electronic equipment and typical variations of utility electric power. One of the most common is disruption of operation of pulse-width-modulated (PWM) adjustable speed drives due to the transient caused by closing utility capacitors (see Figure 1). A switching transient magnitude of only 1.10 to 1.30 pu can be sufficient to disrupt a drive. Equipment designers consider high frequency transients (greater than 10 kHz) as more of a threat to their equipment than low frequency utility capacitor switching transients less than 1 kHz. As a result, capacitor switching transient disruption of PWM drives is the second most common power quality complaint received by utilities, after disruption due to voltage sags.

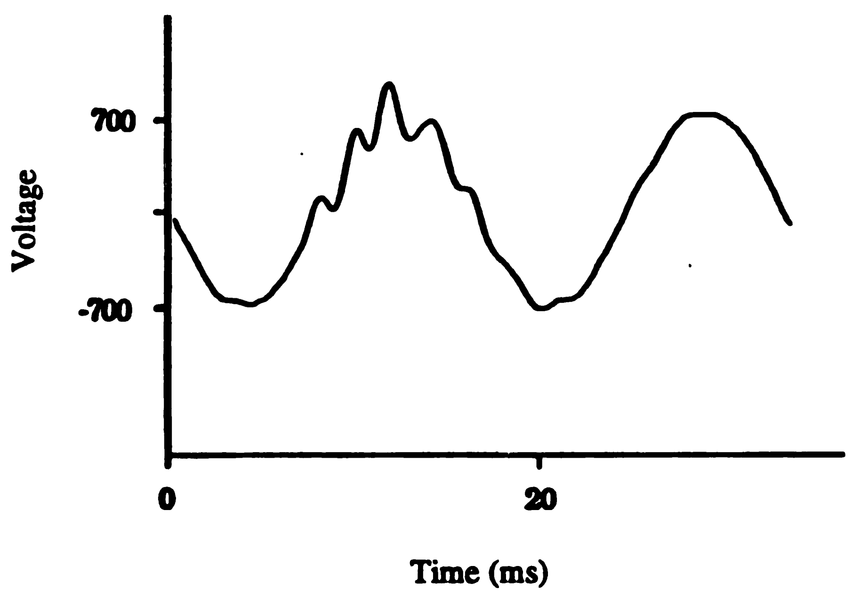

Figure 1. Capacitor switching transient at drive terminals

Historically the study of capacitor switching transients on the power system has focused on the prevention of over-voltage damage to insulation [1]. However, the disruption and damage thresholds for power electronic equipment are much lower than for the electromagnetic and electromechanical equipment previously addressed. The first discussion of this problem in the literature appeared in 1990 by Wagner et al [2] in the form of a detailed case study. In 1991 both Hensley et al [3] and McGranaghan et al [4] conducted computer simulations of the problem with the Electromagnetic Transients Program (EMTP) sponsored by the Electric Power Research Institute (EPRI). Since EMTP is primarily an electric utility modeling tool, the results provide only limited information on drive response. Despite the magnitude of the problem, there have been no further serious analyses of the problem beyond these first three papers.

Because it is a relatively recently discovered problem, industry standards have not addressed the voltage transient from either the utility or equipment perspective. A recently published standard [5] does suggest an optional 5 kHz ring wave test for low voltage equipment intended to simulate a capacitor switching transient. However, a standardized procedure has not been developed to define conformance to the ring wave test.

This thesis explores the generation, propagation, and impact on PWM drives of the utility capacitor switching transient. The primary focus is determination of the drive response mechanism by analysis, simulation, and experimentation. Although the thesis only investigates the impact of the transient on PWM drives, the results also apply to other types of power electronic equipment. The secondary focus is transient generation by analysis and field tests. The mechanism of transmission of the transient between the utility capacitor and the drive is also analyzed.

CHAPTER 2

CAPACITOR SWITCHING TRANSIENT

The generation and propagation of the capacitor switching transient is analytically developed in this chapter. First an expression for the transient is derived and then the associated characteristics are examined.

2.1 Mathematical Representation

The capacitor switching transient is the result of connecting a capacitor to a voltage source through resistance and inductance (see Figure 2). The transient response is found from

$$E \cos \omega t = L\frac{di}{dt} + Ri + \frac{1}{C}\int_0^t i\,dt \tag{1}$$

where L and R are the equivalent inductance and resistance of the network between the capacitor and the source voltage, E is the source voltage at frequency ω, and C is the capacitance of the switched capacitor. Assuming no initial inductor current or initial capacitor charge, the current, i, is

$$i = \frac{E}{Z}\cos(\omega t - \theta) + e^{-\zeta\omega_n t}\left[-\frac{E\cos\theta}{Z}\cos\omega_n t + \frac{E}{\omega_d}\left(\frac{1}{L} - \frac{\omega\sin\theta}{Z} - \frac{\zeta\omega_n\cos\theta}{Z}\right)\sin\omega_n t\right] \tag{2}$$

where the natural frequency, ω_n, of the circuit is defined as

$$\omega_n = \frac{1}{\sqrt{LC}} \tag{3}$$

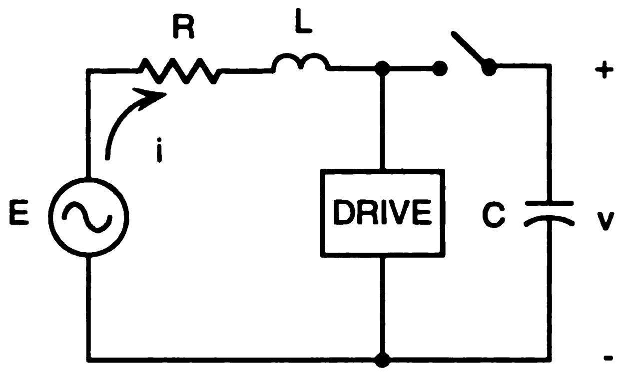

Figure 2. Circuit for analysis of the capacitor switching transient

the damping factor, ζ, is defined as

$$\zeta = \frac{R}{2}\sqrt{\frac{C}{L}} \tag{4}$$

and the damped frequency of oscillation, ω_d, is

$$\omega_d = \omega_n \sqrt{1-\zeta^2} \tag{5}$$

The circuit steady state parameters are the phase angle θ

$$\theta = \tan^{-1}\frac{\omega L - \frac{1}{\omega C}}{R} \tag{6}$$

and the impedance Z

$$Z = \sqrt{R^2 + \left(\omega L - \frac{1}{\omega C}\right)^2} \tag{7}$$

The equipment affected by the switching transient is in parallel with the capacitor. In this analysis we'll assume that the current drawn by the drive to be is much less than the current in the capacitor due to the transient. Hence, the drive does not affect the capacitor voltage or the initial conditions. The voltage during the transient applied to the drive is found by substituting the current solution (2) into (1)

$$v = \frac{E}{\omega CZ}\sin(\omega t-\theta) + \frac{E e^{-\zeta\omega_n t}}{C\left[(\zeta\omega_n)^2+\omega_d^{\,2}\right]}\left[\frac{\cos\theta}{Z}(\zeta\omega_n\cos\omega_n t - \omega_d\sin\omega_n t) - \frac{1}{\omega_d}\left(\frac{1}{L}-\frac{\omega\sin\theta}{Z}-\frac{\zeta\omega_n\cos\theta}{Z}\right)(\zeta\omega_n\sin\omega_n t+\omega_d\cos\omega_n t)\right] \tag{8}$$

The peak transient voltage is determined by the damping factor and can be derived from the voltage expression (8). However, a simpler approach to derive the same conclusion is to assume that the system voltage, E, is constant during the transient resulting in the following differential equation:

$$E = L\frac{di}{dt} + Ri + \frac{1}{C}\int_0^t i\,dt \tag{9}$$

The voltage across the capacitor from the current solution to the above equation is

$$v = \frac{1}{C}\int_0^t E\sqrt{\frac{C}{L}}\,e^{-\zeta\omega_n t}\sin\omega_d t\,dt \tag{10}$$

After integrating and rearranging terms the voltage is

$$v = E\left[1-\frac{e^{-\zeta\omega_n t}}{\sqrt{1-\zeta^2}}\sin\left(\omega_d t+\tan^{-1}\frac{\sqrt{1-\zeta^2}}{\zeta}\right)\right] \tag{11}$$

The peak voltage occurs at $t = \dfrac{\pi}{\omega_n}$ seconds and depends entirely on the damping ratio. If $\zeta = 0$, the peak voltage becomes $2E$ which is the upper limit for the circuit. Figure 3 is a graph of the maximum transient voltage verses damping factor from equation (11).

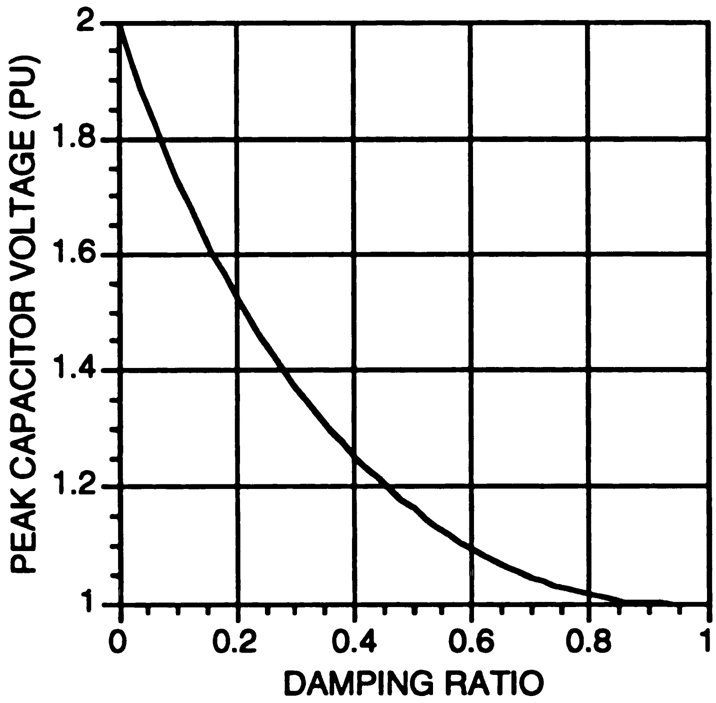

Figure 3. Effect of damping ratio on peak transient voltage.

2.2 Transient Source

The electric utility is the most common source of capacitor switching on the power system. Utility capacitors provide leading reactive power to balance the lagging reactive power generated by utility customers. They switch capacitors near their loads to improve power factor and stabilize voltage. Capacitors are gradually added and removed either automatically or manually over the course of the daily load cycle. There are seasonal trends in voltage and some capacitors, therefore, remain on during high load seasons, such as the summer, and are switched, as needed, during other parts of the year.

Capacitors are operating aids and, as such, are not left on throughout the year. A capacitor that is needed continuously indicates a system design deficiency.

The transient voltage at the utility bus as a result of the capacitor closing computed from expression (8) is shown in Figure 4. The data for the equation was from a typical 40 kV station with X = 18.7 % and R = 0.87 % on 100 MVA base closing a 12 MVAR capacitor.

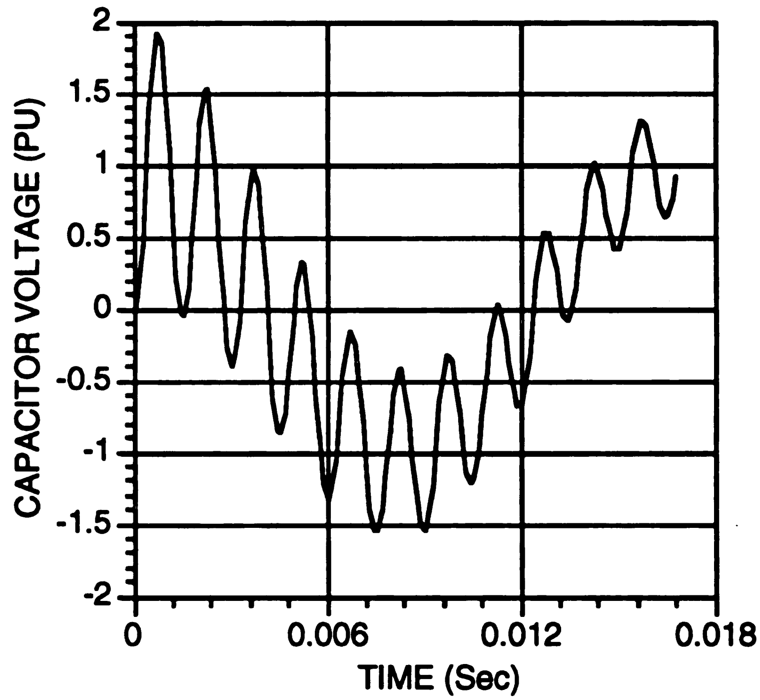

Figure 4. Computed voltage across capacitor

Another source of the capacitor switching transient is the power factor correction capacitors within facilities. These capacitors are added primarily to prevent utility power factor penalties for large utility customers. Generally, these capacitors are left on and are not automatically switched. However, if under light load conditions the capacitors cause excessively high facility voltage, the capacitors are automatically switched off.

Since utility capacitor switching is the most common source of the transient, the remainder of this analysis focuses on the utility network. Nevertheless, the results may also be applied to facility capacitors.

In the previous analysis of the voltage transient the capacitor was assumed to be fully discharged when closed into the circuit. Another mechanism that can generate even higher transient voltages is restrike of the switching device. During opening or closing of a capacitor, the arc may be momentarily interrupted. When reestablished the capacitor is charged, increasing the transient magnitude. Further restrikes can escalate the capacitor voltage to very large transient magnitudes as each reignition begins with progressively higher initial voltage across the capacitor. As utility capacitor banks increased in number form the 1950s and 1960s, restriking became more of a problem. Pflanz and Lester [6] discuss some of the equipment options to reduce restrike.

The peak restrike voltages can be derived from the differential equation (9) by setting resistance equal to zero and adding an initial capacitor voltage $V_C(0)$.

$$E = L\frac{di}{dt} + \frac{1}{C}\int_0^t i\,dt \tag{12}$$

The voltage across the capacitor then becomes

$$v = V_c(0) + \frac{1}{C}\int_0^t \left(E - V_c(0)\right)\sqrt{\frac{C}{L}}\sin\omega_n t\,dt \tag{13}$$

or

$$v = E + \left(V_c(0) - E\right)\cos\omega_n t \tag{14}$$

If the capacitor has an initial voltage $V_C(0) = -E$, the peak transient voltage is $3E$.

Air or SF_6 insulated capacitor circuit switchers in service today are designed to reduce the possibility of restrikes. The circuit switcher is a special purpose device designed for capacitor switching duty. They have no fault duty rating but are more economical than circuit breakers for frequent switching service. Although these devices minimize restrikes, they are not universally applied and can still cause restrikes as a result of malfunction.

2.3 Damping

As shown in Section 2.1, the peak transient voltage depends entirely on the damping ratio and is most sensitive to resistance. Because losses are minimized in power systems, damping is not a practical means to control the transient. For example, the damping ratio for the utility 12 MVAR capacitor closing shown in Figure 4 is 0.0152. There is no accepted peak voltage limit that would assure the transient would not cause problems. A damping ratio of 0.2 would limit the peak voltage to 1.5 pu as shown in Figure 5. Yet, this damping ratio is twice as large as any practical damping ratio for a utility network. Therefore, the natural damping provided by the utility network is insufficient to reduce the magnitude of the transient.

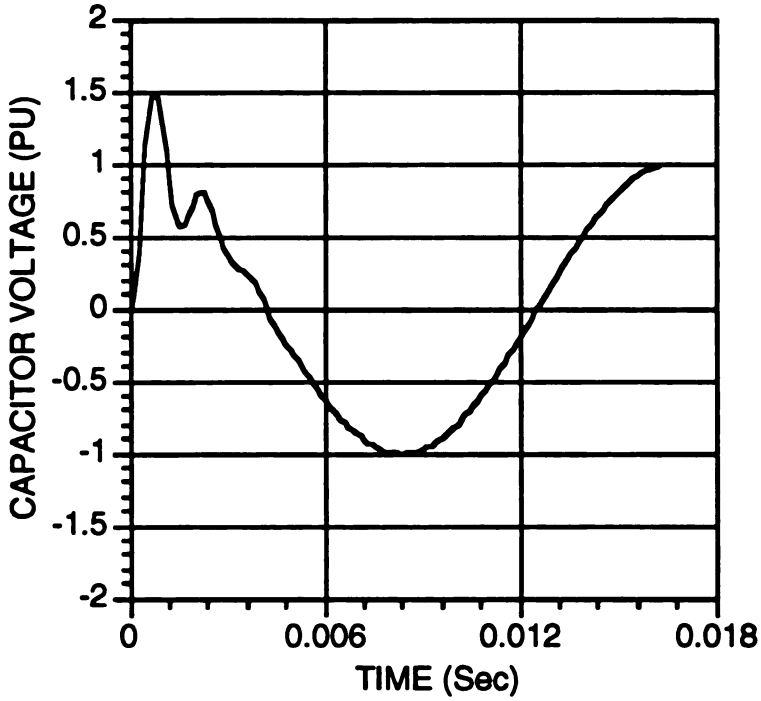

Figure 5. Capacitor voltage response with a damping factor of 0.2

2.4 Transient Modes

It has been assumed in this analysis that the utility capacitor bank and step-down transformer are the primary circuit elements that determine transient behavior. However, distinctive oscillation frequencies following closure result from the interaction of different combinations of network capacitances and inductances. Pflanz and Lester grouped the equipment and corresponding transient frequencies and the results are shown in Table 1.

Table 1. Characteristic transient frequencies and associated circuit elements

Capacitive Element	Inductive Element	System Frequency Multiplier
Switched Bank	Transformer	10
Switched Bank & Energized Bank	Transformer	10^2
Switched Bank & Energized Bank	Inter-Bank Inductance	10^3
Cable	Cable	10^4

The 10^4 times mode involves reflections due to line termination impedance mismatch. Drives are almost exclusively affected by the 10 times frequency mode which corresponds to a range of approximately 300 Hz to 900 Hz. The reason for the particular window of vulnerability of drives to this frequency band is discussed later.

Another possible mode not covered by Pflanz and Lester is a result of line switching. The switched cable or line capacitance would interact with the step-down transformer inductance to produce a transient. Since a line has approximately 1% to 2% the capacitance of the capacitor bank, the transient frequency mode is 10· times greater than the transformer-capacitor bank mode.

2.5 Transient Propagation

The mechanism of propagation of the transient away from the source depends on the transient frequency and the network. If the length of the network is less than 1/6 the wavelength of the transient, the network can be considered an extended lumped parameter circuit. If the length of the network is greater than 1/6 the wavelength of the transient, the network can be considered a high frequency transmission line. Since 600 Hz is the average transient frequency affecting drives, the corresponding wavelength is 500 km. To use the lumped parameter network model, the circuit must be less than 83 km. Most utility distribution and sub-transmission circuit lengths are under 60 km.

The number of networks between the source and the load equipment is determined by the decoupling points which are, most commonly, power transformers. If the transformer and load impedance of the sub-network is much less than the source network, the two may be treated separately. Vilcheck and Gonzalez suggest the ratio of the impedances between the two sub-networks be greater than 25 [7]. For each of the sub-networks, a separate lumped RLC (resistor, inductor, and capacitor) circuit can be developed (See Figure 6). The circuit model is derived from a series inductance and resistance representing the transformer and conductor with a shunt capacitance representing the intra-conductor capacitance or capacitor banks. The circuit behaves as a low pass filter. Power distribution networks typically resonate at between 300 Hz and 500 Hz [4] and, therefore, will attenuate transients above that frequency.

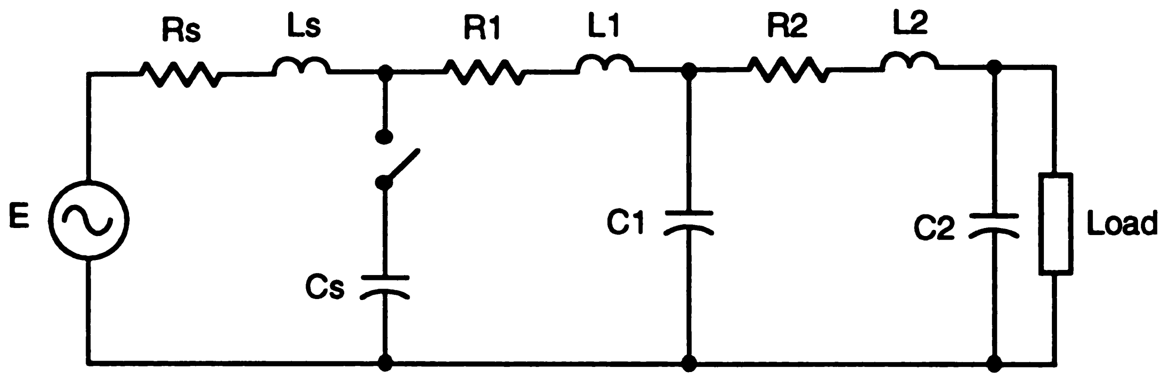

Figure 6. Lumped transient propagation model.

However, a problem can arise near the resonant frequency. An underdamped circuit can magnify a transient whose frequency is near the network's resonant frequency. Zaborszky and Rittenhouse show that the sub-network can multiply the initial transient by 11.6 times [8]. Schultz et al describe three conditions necessary for magnification of the capacitor switching transient within a sub-network [9]:

1. To be considered a sub-network, it must be decoupled from the other networks connected to it, including the source network.
2. The frequency of the transient must be close to the resonant frequency of the sub-network.
3. The sub-network must be underdamped.

The first issue has already been discussed. For the second issue, the proximity of the transient and sub-network resonant frequencies can be quantified by the quality factor, Q_0, used for filter circuits and defined as

$$Q_0 = \frac{\omega_n L}{R} = \frac{\omega_n}{\omega_{bw}} \tag{15}$$

where ω_{bw} is the bandwidth between the two frequencies on either side of the resonant frequency that are $\frac{1}{\sqrt{2}}$ times the peak magnitude. For the utility network just used, the bandwidth is 3 Hz. Because the X/R ratio at system frequency of power systems is typically 5 or greater, the bandwidth will not be more than 12 Hz. Therefore, the transient frequency must be very close to the resonant frequency of the sub-network for the transient to be sustained.

The third item is a restatement of the fact that, at resonance, the voltage across the capacitor for a series RLC circuit can be much greater than the circuit source voltage.

For the sub-network in series resonance, the current is determined by the resistance. The ratio of the magnitude of the output voltage to the input voltage, η, is

$$\eta = \left| \frac{\frac{E_{in}}{R} \frac{1}{j\omega_n C}}{E_{in}} \right| = \frac{1}{R}\sqrt{\frac{L}{C}} = \frac{1}{2\zeta} \tag{16}$$

Since ζ can vary from zero to one, the transient voltage out of the sub-network can be vary from 1/2 the magnitude of the input to infinity.

With the damping ratio typically less than 0.1 for power systems, the nearness of the transient frequency to the sub-network resonant frequency determines how the capacitor switching transient is transmitted from the source network.

CHAPTER 3

PWM DRIVE SUSCEPTIBILITY

The pulse-width-modulated (PWM) adjustable speed drive is the most common equipment to be affected by the capacitor switching transient and this chapter explores the possible reasons for the vulnerability. Although the discussion is limited to PWM drives, the mechanisms can apply to other types of drives and other power converters. First, the PWM drive is described and then the sub-parts analyzed for vulnerability.

3.1 Pulse Width Modulated Drive Description

The PWM drive varies motor speed by providing power of variable frequency and voltage to induction or synchronous motors. The essential elements of the drive are shown in Figure 7. The three phase full-wave rectifier section uses uncontrolled diodes: hence losses are reduced, power factor is increased, and control circuitry is minimized compared to thyristor drives. The inductor-capacitor (LC) filter is the dc link between the rectifier and inverter that filters the ripple and further improves power factor. The inverter section converts the dc to the required ac frequency and voltage by pulse-width-modulation at a switching frequency of 10 to 20 times the base frequency. Because of the required switching speed, the inverter section most frequently uses bipolar transistors, insulated gate bipolar transistors (IGBT), or field effect transistors (FET). These semiconductors have simpler gating requirements than thyristors and, hence, require less complex control circuitry. However, transistors do not have the current handling capability of slower semiconductors and, consequently, PWM drives are only available up to several hundred horsepower.

An advantage of the PWM drive is that it produces a sinusoidal-like voltage waveform for the motor. With reduced motor voltage harmonics, cogging is prevented and motor losses are reduced. For constant torque operation from approximately 10% to 110% synchronous speed, the volts per hertz ratio is held constant at the output of the inverter to maintain a constant stator magnetic field. If voltage is held constant as frequency is changed, torque will vary inversely with speed and the motor will provide constant horsepower over approximately a 2 to 1 speed range.

A further advantage is that the PWM drive uses a standard induction motor. The induction motor is a rugged, simple, inexpensive, and low maintenance device compared to dc or specialty motors.

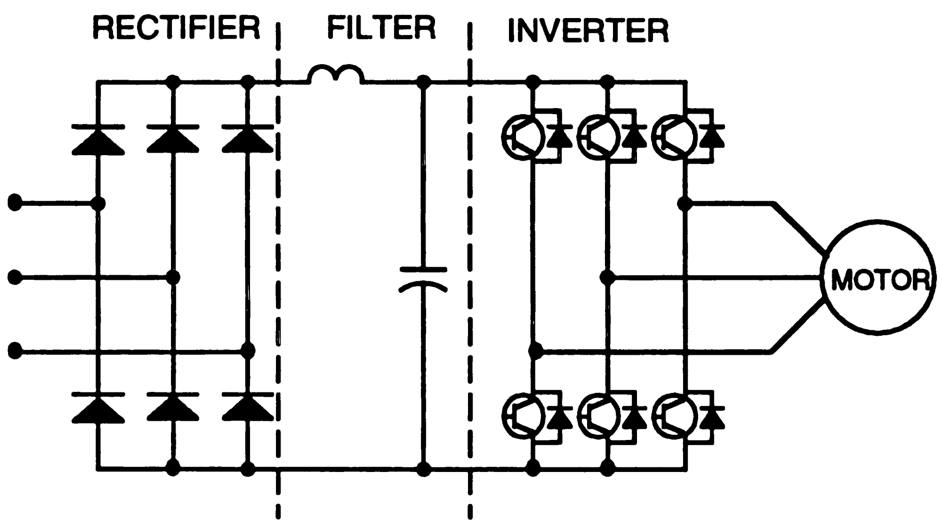

Figure 7. Diagram of PWM drive

3.2 Susceptibility

The three major sub-parts of the PWM drive are analyzed for their susceptibility to the capacitor switching transient. They are the rectifier, filter, and inverter sections.

3.2.1 Rectifier

The 60 Hz rectification section consists of a three phase full-wave bridge. The diodes conduct symmetrically about the peak of the crest of the voltage waveform. The width of the conduction interval depends on the drive load level with higher load requiring a wider conduction interval. Since when conducting the diode directly passes the voltage waveform, a damped oscillatory transient in the range of 300 Hz to 900 Hz readily passes through the rectification stage.

The rectifier section contains devices to protect the semiconductors from disruptive and damaging electrical disturbances. Transient voltage surge suppressors are connected at the ac incoming terminals. However, these devices begin operating at 200% to 300% of rated voltage. Since the capacitor switching transient can be 100% to 200% greater than rated voltage, the surge suppressors have no effect. If the surge suppressor is sized to a lower voltage, it partially conducts during high voltage conditions and will over heat. Noise filters are also sometimes added to the incoming power lines to protect the drive against conducted noise. These are ineffective since their cut-off frequencies are generally at a few kilohertz and are too high to attenuate the transient in question. Because of the harmonic currents generated by the rectifier, lower cutoff frequencies could cause excessive insertion loss.

3.2.2 Inverter

The capacitor switching transient can cause excessive voltage at the inverter input. Due to cost, the inverter transistors may have limited over voltage capability. For example, 480 V drives use 1200 V transistors where the rectified dc can be as much as 690 V. To protect the inverter transistors, an over voltage trip is provided at approximately 760 V for 400 μs that shuts down the inverter. Consequently, for nominal input voltage conditions, the inverter has only a 10% over-voltage margin. For such an event, the drive controls often provide fault code that will indicate high dc voltage as the cause. This mechanism is much less likely with 240 V drives since they use 800 to 1000 V transistors and have a 40% or greater over-voltage margin.

Some drive manufacturers have recommended that their dynamic braking option will decrease susceptibility to the transient. In situations where the mechanical load drives the motor above synchronous speed, the motor operates as an induction generator and increases the drive dc voltage. An example would be quickly decreasing the drive frequency while the motor is operating a high inertia load. The frequency corresponding to the motor's speed may, momentarily, exceed the inverter output frequency. The brake is a resistor that is electronically switched across the drive dc bus to lower the dc voltage. However, in all the cases where this was attempted, the brake did not respond sufficiently fast to limit the transient voltage.

3.2.3 Filter

Although other factors contribute toward susceptibility to the capacitor switching transient, the dc filter design is the most important. Since the filter components are relatively expensive compared to other drive components, filter performance must be balanced against cost. There are two mechanisms by which the filter can cause equipment disruption. The first is the filter presents a low impedance to the transient and causes a corresponding current transient that can either trip over-current devices or damage the rectifiers. The second is that the filter passes the voltage transient and inverter dc over-voltage protection is actuated.

The filter normally takes the form of an inductor-capacitor (LC) low pass filter (see Figure 8). Ripple current is shunted through the capacitor and ripple voltage is dropped across the inductor. The capacitor is sized to shunt most of the ripple current from the load. If the load is variable, the criterion should be applied at maximum load. If the load is assumed to be purely resistive and represented by R_L, the filter capacitor is sized by the criterion

$$R_L >> \frac{1}{n\omega C} \tag{17}$$

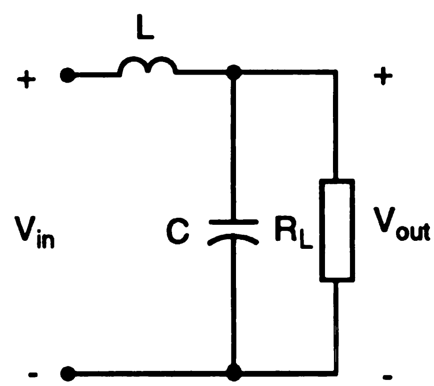

Figure 8. The LC filter.

where $n\omega$ is the ripple frequency. Setting R_L to be 10 times the capacitive reactance, C becomes

$$C = \frac{10}{n\omega R_L} \tag{18}$$

The load ripple voltage can be determined by applying the voltage divider rule across the inductor and capacitor. If V_{IN} is the rms value of the ripple voltage into the filter and V_{OUT} is the ripple voltage out of the filter,

$$V_{OUT} = \frac{\dfrac{V_{IN}}{n\omega C}}{n\omega L - \dfrac{1}{n\omega C}} = \frac{V_{IN}}{(n\omega)^2 LC - 1} \tag{19}$$

The inductor is then sized to provide the specified voltage ripple reduction from the expression

$$L = \frac{\dfrac{V_{OUT}}{V_{IN}} + 1}{(n\omega)^2 C} \tag{20}$$

In effect, the load impedance determines capacitance and the desired voltage ripple determines the inductance. Although the amount of ripple depends on the application, 1% or less is considered typical. The resonant frequency of the LC network should be chosen between the possible harmonic current frequencies of the drive to prevent resonance. The inductor also affects power factor and the current range over which the rectifiers continuously conduct. On low power equipment (for example drives under 5 hp), the inductor is frequently omitted to reduce cost.

Since the filter capacitor can be considered a short circuit at the transient frequencies under discussion, the inductor determines transient current response. If the transient current magnitude is high enough, the drive may be disrupted or damaged. For equipment that operates with variable load, the mechanism is more probable at heavy rather than light load conditions since device thermal margins are narrower. The current surge could cause the over current protection for the equipment to operate. Or, due to poor design, the rectifiers may fail before the over-current protection acts. In small motor drives without an inductor in the filter, fuse blowing is the most common response to capacitor switching.

The other mechanism of disruption is that the voltage transient passes through the filter to the inverter where over-voltage protection operates. Smaller drives where the filter inductor has been omitted are very vulnerable to this mechanism. Those filters that include the inductor would be expected to attenuate the transient since the transient frequency is above the ripple frequency. However, there are cases where an inductor is present and transient passes through the filter with little, if any, attenuation. For this to occur, the inductor must saturate because of the transient. With the inductance at an eighth or tenth of its unsaturated value, the filter would let pass transients of frequency less than 1 kHz.

McLyman describes one approach for inductor design in [10]. The maximum operating flux for the inductor may be 20% to 30% below saturation and is composed of flux due to the maximum dc current and flux as a result of the voltage across the inductor (see Figure 9). Assuming the voltage across the inductor to be sinusoidal, the resulting current through the inductor creates a flux density B_{ac}

$$B_{ac} = \frac{V_R}{4.44\omega_R A_C N \cdot 10^{-8}} \tag{21}$$

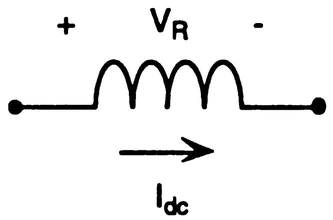

Figure 9. The currents and voltages contributing to inductor flux.

where V_R is the ripple voltage, A_c is the core cross sectional area, and N is the number of turns. The flux density due to the dc current I_{dc} is

$$B_{dc} = \frac{0.4\pi N I_{dc} \cdot 10^{-4}}{l_g} \tag{22}$$

where l_g is the core gap length. The total flux is the sum of the ac and dc components and is given by

$$B = \frac{0.4\pi N I_{dc} \cdot 10^{-4}}{l_g} + \frac{V_R}{4.44\omega_R A_C N \cdot 10^{-8}} \tag{23}$$

The ripple voltage for a three phase full wave rectifier is only 6% of the phase voltage. If the capacitor switching transient is 20% or more of the phase voltage and I_{dc} is at its rated value, the transient can easily drive the inductor into saturation. It appears that a heavily loaded drive would be more susceptible to transient saturation.

With the inductor in saturation, the filter behaves like an RC network (see Figure 10) where R is the equivalent resistance of the filter. The transient voltage integrates across the capacitor until the over voltage limit is exceeded for the inverter.

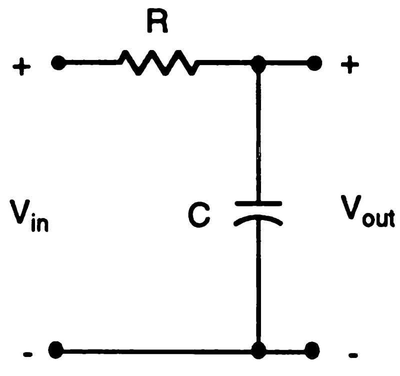

Figure 10. RC integrating network.

CHAPTER 4

EXPERIMENTAL RESULTS

Two experiments to evaluate the response of PWM drives to the capacitor switching transient are described. The first is a field test of the response of the power distribution network and two drives in an automobile engine factory when utility station capacitors are switched. The second test is application of a laboratory generated transient to a PWM drive.

4.1 Field Investigation

The field investigation was at a dual utility fed automotive engine manufacturing plant. Each 40 kV plant feed is from separate 120 kV to 40 kV utility stations (east and west) as shown in Figure 11. Plant primary voltage was stepped-down to 13.2 kV to operate large motors and numerous unit substations that further reduced voltage to 480V. The plant operates two camshaft grinders from separate primary sources; each grinder utilizes one 10 hp and one 50 hp PWM drive. The west utility station had one 40 kV 12 MVAR capacitor bank while the east station has two 18 MVAR and two 24 MVAR banks. When the west station 12 MVAR bank was energized, the 50 hp drive would frequently trip. Yet the identical drive fed from the east station never tripped when any of the four station capacitor banks were energized. The camshaft grinders were added two years prior to the discovery of the problem. During those years, the 12 MVAR bank was switched infrequently and, therefore, a correlation with the drive trips was difficult to establish.

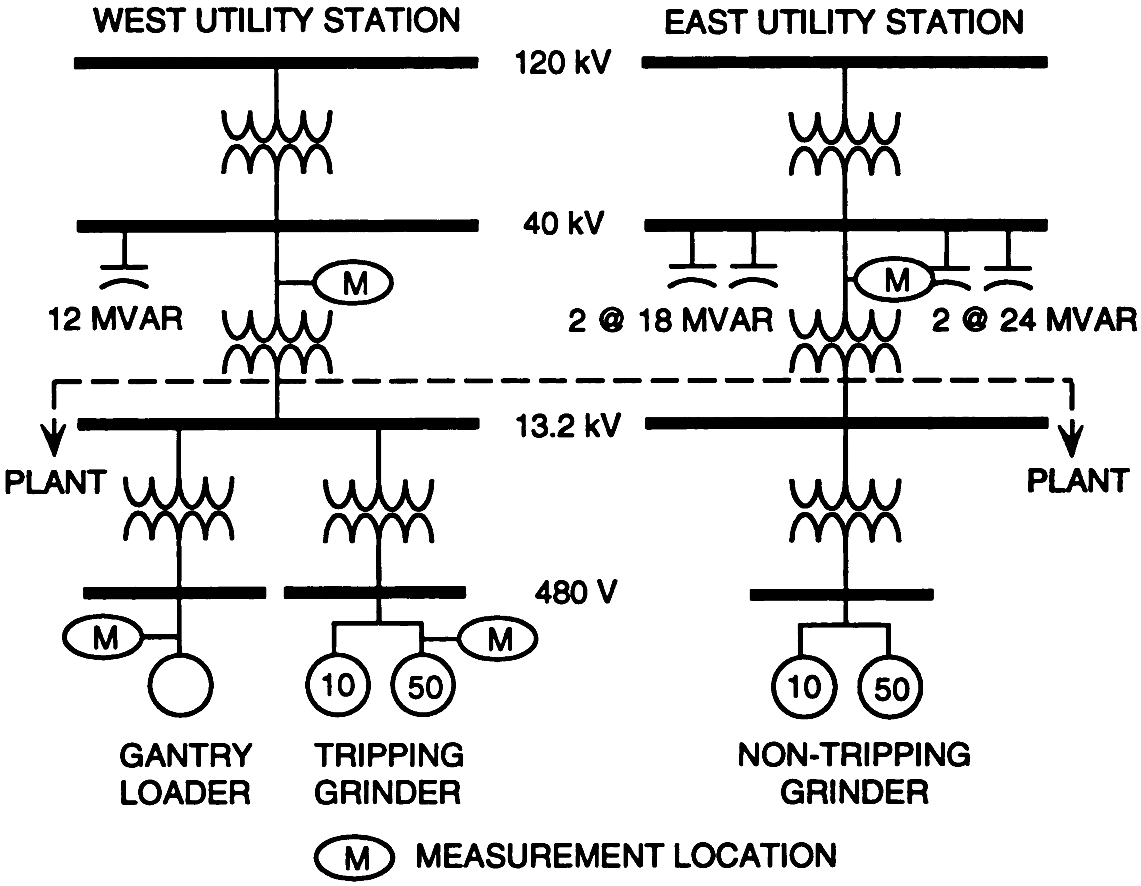

Figure 11. Diagram of utility and plant electrical network.

4.1.1 Test Description

The goal of the investigation was to identify the trip mechanism of the 50 hp drive and to determine how other 480 V buses within the facility would respond to the same transient. Since an identical 50 hp drive was not available on another 480 V bus fed from the same primary source, a gantry loader servo positioner was selected. The rectifier and filter sections of the servo positioner were similar to the PWM drive. When the 50 hp drive tripped, the diagnostic indication was high dc bus voltage. Therefore, for the test the dc voltage was recorded as well as the phase voltage for both drives while the 12 MVAR capacitor bank was switched (see Figure 12). The loader and grinder were operated without work pieces to prevent damage to the machines during drive trips.

Drive operating power levels were not measured. The voltages were recorded at separate locations simultaneously by dual channel digital oscilloscopes connected through voltage isolators. The isolators were necessary to isolate the oscilloscope chassis from line potential.

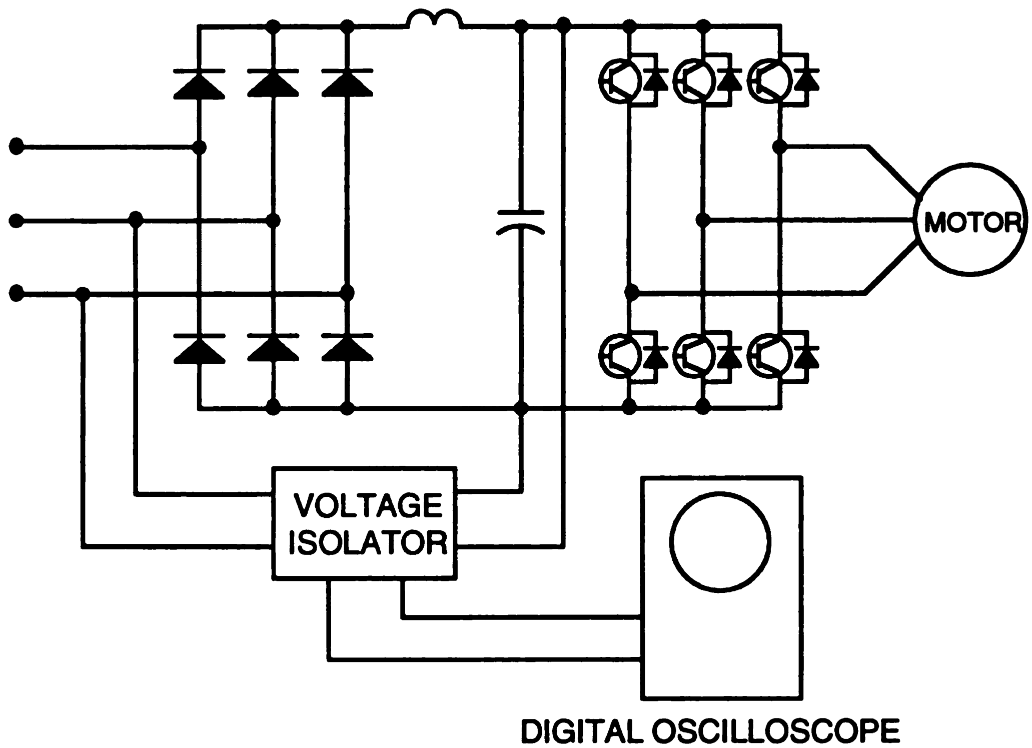

Figure 12. Method for measuring drive ac and dc voltage

In separate tests, the transients were measured at the primaries of the plant's two 40 kV transformers with power quality disturbance instruments monitoring at the utility revenue meter potential transformers (see Figure 13).

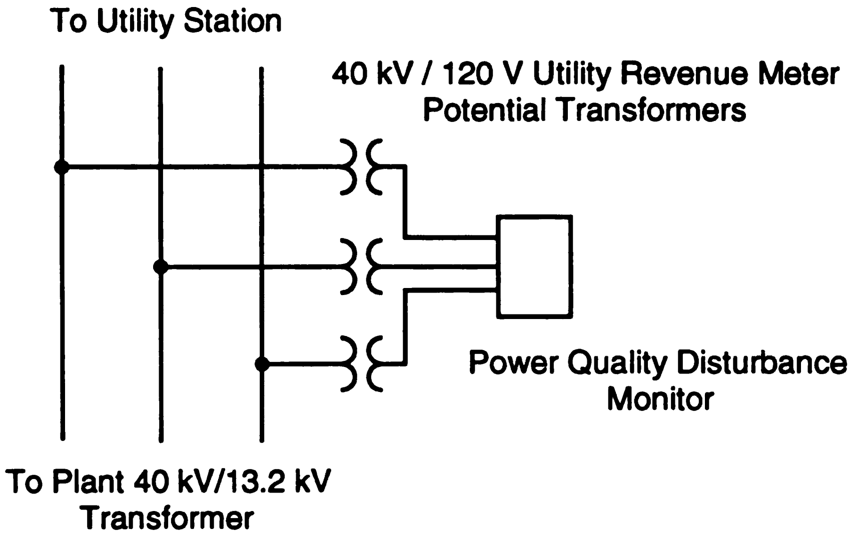

Figure 13. Disturbance monitor/utility interface

4.1.2 Network Analysis

The measured frequencies of the transient at several points in the network are shown in Table 2. The frequency of the transient at the plant's two 40 kV transformers for the table were taken from the measurements shown in Figures 14 and 15, which were conducted independently of the 480 V measurements. From the discussion on transient propagation in Chapter 2, it follows that the sub-network response depends on the how close the transient frequency is to the sub-network resonant frequency.

The resonant frequency for the 40 kV utility bus is calculated from (5) and utility data. The computed value for the east station frequency is identical to the measured value. However, there is a significant difference between the computed and measured values for the west station. The only possible explanation for the mismatch is that the utility data for the west station is less accurate than the data for the east station. Since the tests were conducted at an operating plant, there was little opportunity for further investigation.

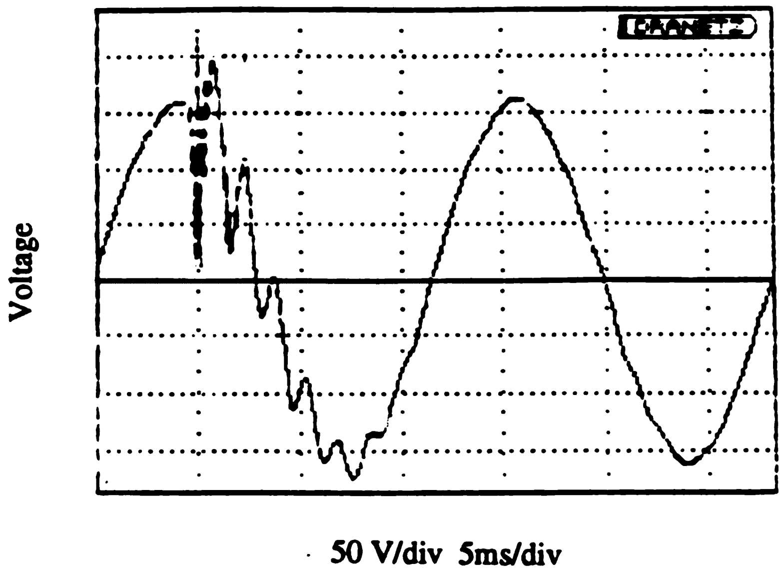

Figure 14. Recorded transient at plant east utility feed

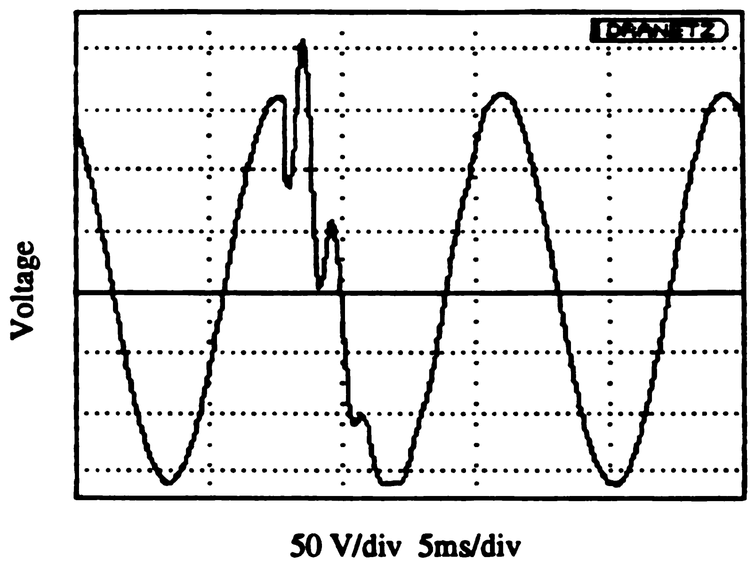

Figure 15. Recorded transient at plant west utility feed

Table 2. Computed and measured transient frequencies

Station	Computed 40 kV ω_d	Measured at Plant 40 kV ω_d	Measured at Grinder 480 V ω_d	Measured at Gantry 480 V ω_d
East	667 Hz	660 Hz	-	-
West	525 Hz	420 Hz	450 Hz	400 Hz

The transient was only measured at the 480 V level on buses fed from the west utility station. Since the identical grinder drive fed from the east utility station was not affected by capacitor switching, it would be difficult to capture the transient at that grinder. The transient frequencies at the grinder and gantry 480 V feeds are close to the measured 40 kV frequency. The slight difference among the 40 kV and the 480 V frequencies is likely due to the error in reading the frequencies directly from the voltage plots.

Table 3. Computed and measured transient peaks

Station	Computed at 40 kV Capacitor (pu)	Measured at Plant 40 kV (pu)	Measured at Grinder 480 V (pu)	Measured at Gantry 480 V (pu)
East	1.9	1.2	-	-
West	1.9	1.2	1.24	1.08

The transient response at the utility capacitor banks were computed from equation (8) and are shown in Figures 16 and 17. These figures can be compared to the measured transients shown in Figures 14 and 15 taken at the plant transformer primaries. As shown in Table 3, the peak transient voltage of both of the plant feeds is approximately 1.2 pu whereas the computed peaks are 1.9 pu. The plant 40 kV and utility 40 kV bus are all part of the source network: there are no intervening sub-networks between the two locations to attenuate the transient. Nevertheless, there are two reasons for a lower value at the plant. First, since the plant is several kilometers from the utility station, part of the difference may be due to voltage drop on the utility lines. The second is that the disturbance monitor that recorded the transient was coupled through potential transformers. The potential transformers may not have sufficient bandwidth to accurately reproduce the transient. The other measurements for the field investigation were performed with directly connected monitors and avoid this problem.

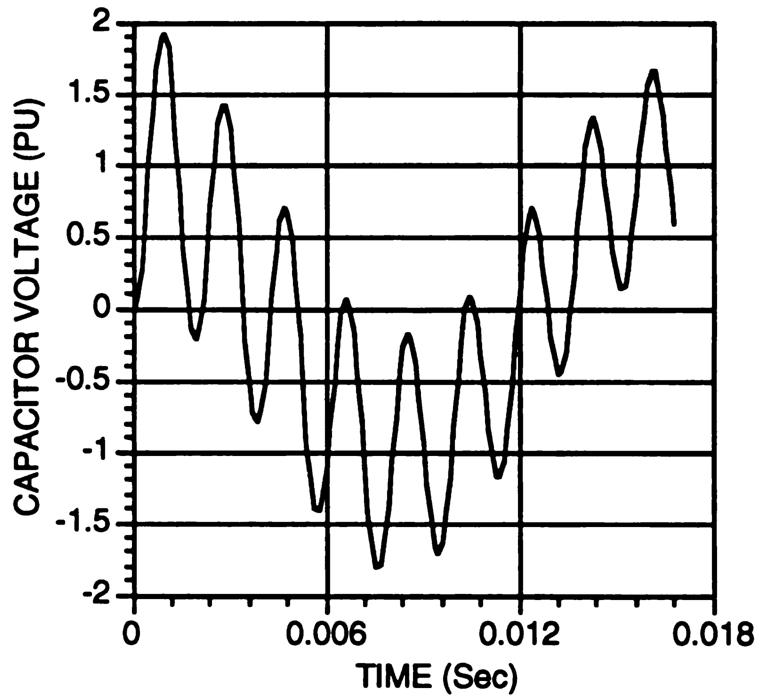

Figure 16. Computed west utility capacitor response

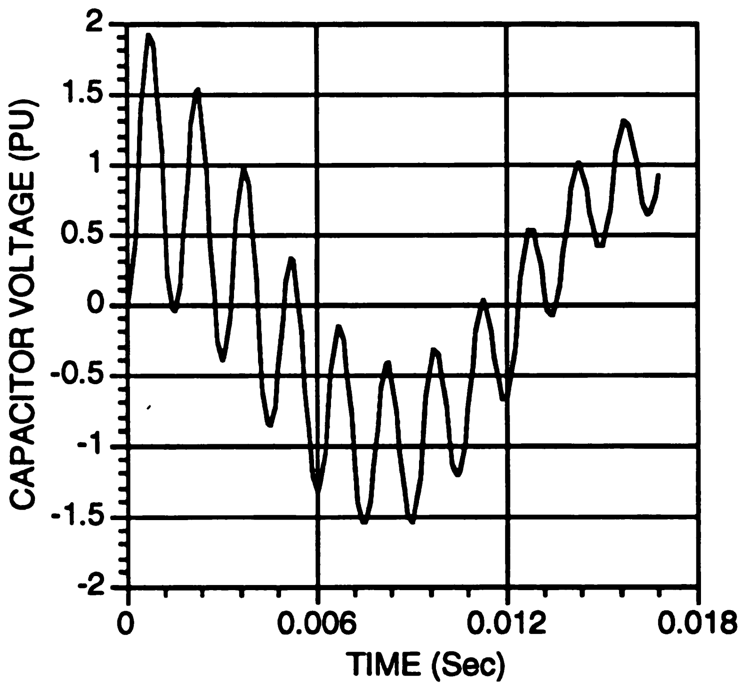

Figure 17. Computed east utility capacitor response

Table 3 shows the peak magnitudes of the transient corresponding to the test points in Table 2. The per unit magnitude of the transient at the 480 V grinder was equal to the 40 kV magnitude. The per unit magnitude of the gantry drive 480 V transient was lower than at the other two measurement points.

The ac magnitude of the phase voltage transient at the two drive locations showed the different responses of the two 480 V networks to the same input transient. The magnitude of the transient at the gantry drive was 1.08 pu of nominal voltage and it was 1.24 pu at the grinder drive terminals. Indeed, it appears the grinder drive would not trip if it were fed from the same 480 V bus as the gantry drive. The difference in response may be due to different resonant characteristics. The design and operating characteristics of the two networks were identical. Each had 60 kVAR of power factor correcting capacitors, the same nominal transformer impedance, and less than 10% loading during the test. At the time of the investigation, the 480 V network resonance characteristics

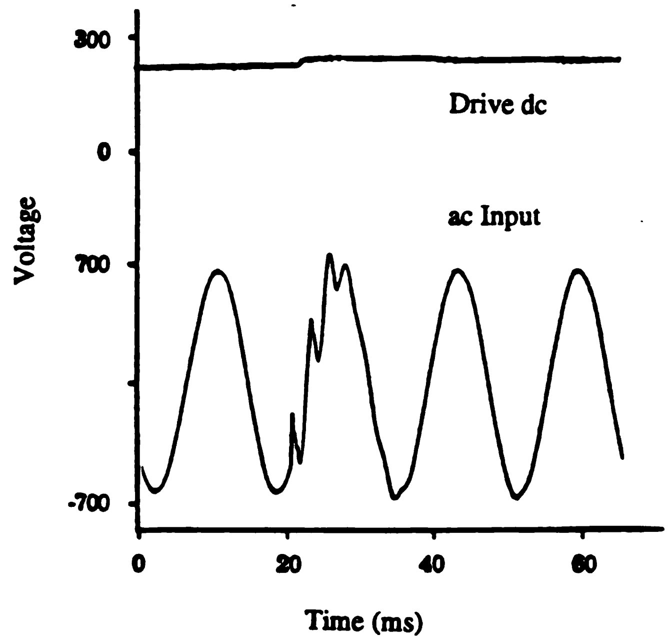

Figure 18. Gantry drive transient waveforms

were not considered an issue and, therefore, were not examined. However, the maintenance condition of the two networks were thought to be relevant and were compared. The bus, cables and associated equipment feeding the grinder were inspected and tested for ground impedance, voltage imbalance, connection integrity, open capacitor fuses, and harmonics. The results showed the network was in very sound condition and the cause for the separate responses of the two networks was not investigated further.

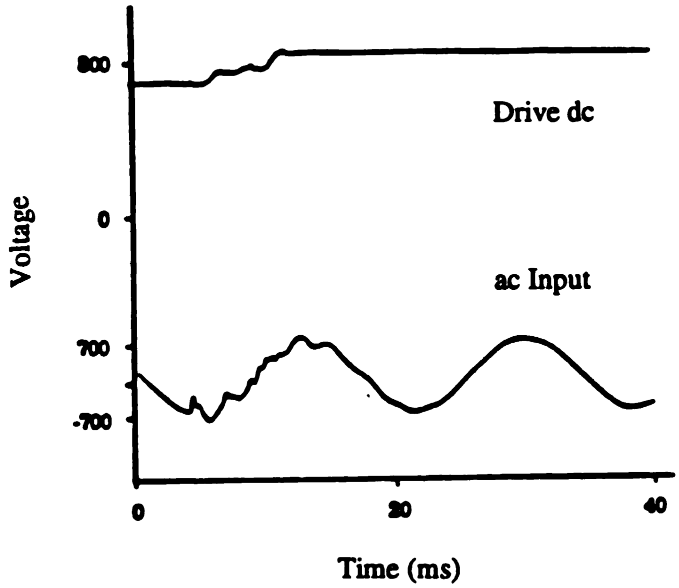

Figure 19. Grinder drive transient waveforms

4.1.3 Drive Analysis

The ac and dc responses at the two pieces of equipment are shown in Figures 18 and 19. Due to the scale of the figures, the relative amplitudes of the transients are actually opposite to what they appear.

The magnitude of the transient response on the dc buses closely followed the magnitude of the transient on the phase voltage on both of the drives (see Table 4). The dc transient magnitude was 1.20 pu on the 50 hp grinder drive and 1.06 pu on the gantry drive. The station capacitor bank was closed three times for the test and in each case the grinder drive tripped on high dc voltage and the gantry drive did not. The dc trip voltage for the grinder drive is 760 V while the peak transient voltage shown on the response plot is 843 V. It was not known where exactly the grinder drive tripped on Figure 19.

Table 4. Drive ac and dc transient peaks

Drive	480 V Transient Peak (pu)	Drive dc Transient Peak (pu)
50 hp Grinder	1.24	1.2
Gantry Loader	1.08	1.06

The response out of the LC filters of the two drives were very different. There was a small step change of dc voltage at the gantry loader drive due to the transient. The dc voltage at the time of the transient at the grinder drive appears from Figure 19 to be the result of integration of the transient voltage. The LC filter of the 50 hp grinder drive has a cutoff frequency of 107 Hz and should have attenuated the 450 Hz transient by 22 db. As discussed in Chapter 3, the mechanism operating in this case may be saturation of the LC filter inductor.

To determine if the inductor was saturating, the inductors from two drives were tested. One was from the 50 hp grinder drive and the other was from a 10 hp PWM drive known not to trip from the capacitor switching transient. The inductors were driven by a voltage source 2 kW audio amplifier at 450 Hz (see Figure 20). Unlike an actual drive filter, there was no dc bias current through the inductor. Saturation was indicated by the cross-over to distorted voltage across the inductor by an oscilloscope. The inductor from the 10 hp drive saturated at 90 Vrms and the inductor from the 50 hp drive saturated at 14 Vrms. From these results, it is evident that inductor saturation is the most probable cause of the 50 hp drive trip due to the capacitor switching transient.

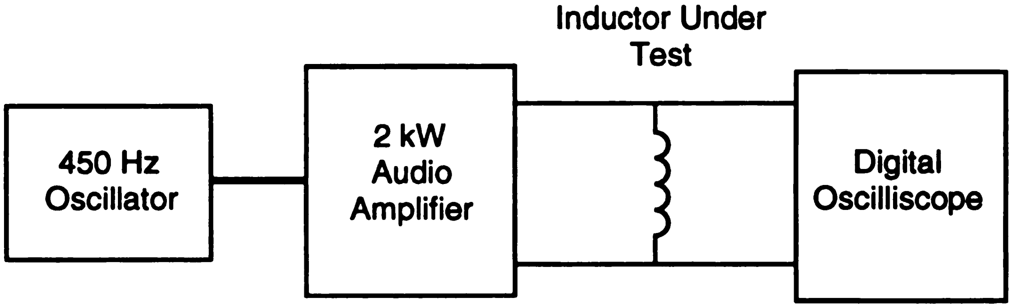

Figure 20. Inductor saturation test configuration

4.2 Laboratory Tests

A laboratory test of the response of a PWM drive to the capacitor switching transient was performed to gain a broader understanding of the response. The test set-up for the drive is shown in Figure 21. The 10 hp drive operates from 480 V three phase with a series notch filter on one of the phases tuned to the frequency of the transient. A 500 Hz damped transient generated by the function generator drives the 2 kW audio amplifier. The amplifier is isolated from line voltage by a 4:1 step-up transformer. The filter's high impedance at the transient frequency causes the transient current to flow through the ac lines to the drive. The equivalent utility source impedance is very low at this frequency. The transient frequency was chosen to be between the drive 8th and 9th current harmonics to prevent excessive voltage drop across the filter. To evaluate the impact of the transient at several motor load levels, the motor load is controlled by an eddy current dynamometer. The voltage is monitored through two voltage isolators that attenuate the signal and isolate the measured voltage from the oscilloscope ground. Only two oscilloscope channels were available for the test so that the results show two separate events.

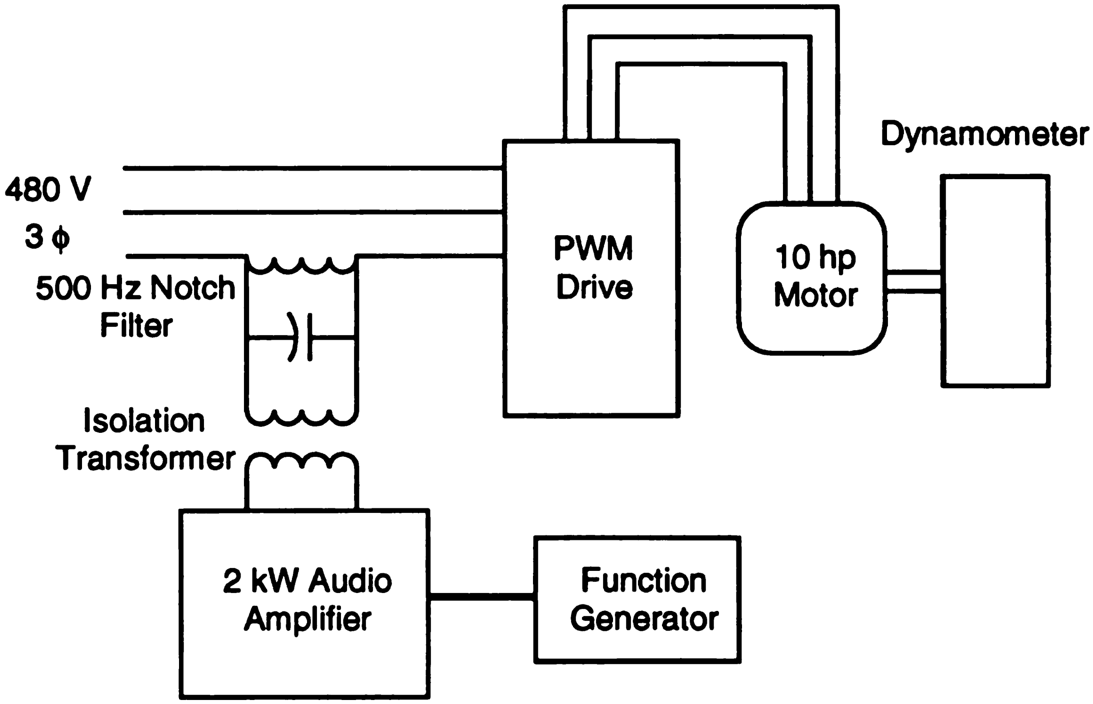

Figure 21. Laboratory simulation of capacitor switching transient

Results of the test are shown in Figure 22 for the ac input, rectifier output, and LC filter output with an applied transient magnitude of 0.38 pu. In no case did the transient cause the drive to trip. The transient is clearly present in the rectifier output but almost entirely absent from the LC filter output. There is a slight step increase of dc voltage similar to the increase observed in the gantry loader drive (compare to Figure 18). Although the test results shown in Figure 22 were with the drive lightly loaded, results were independent of drive load. The test verifies that this drive would not be vulnerable to the capacitor switching transient.

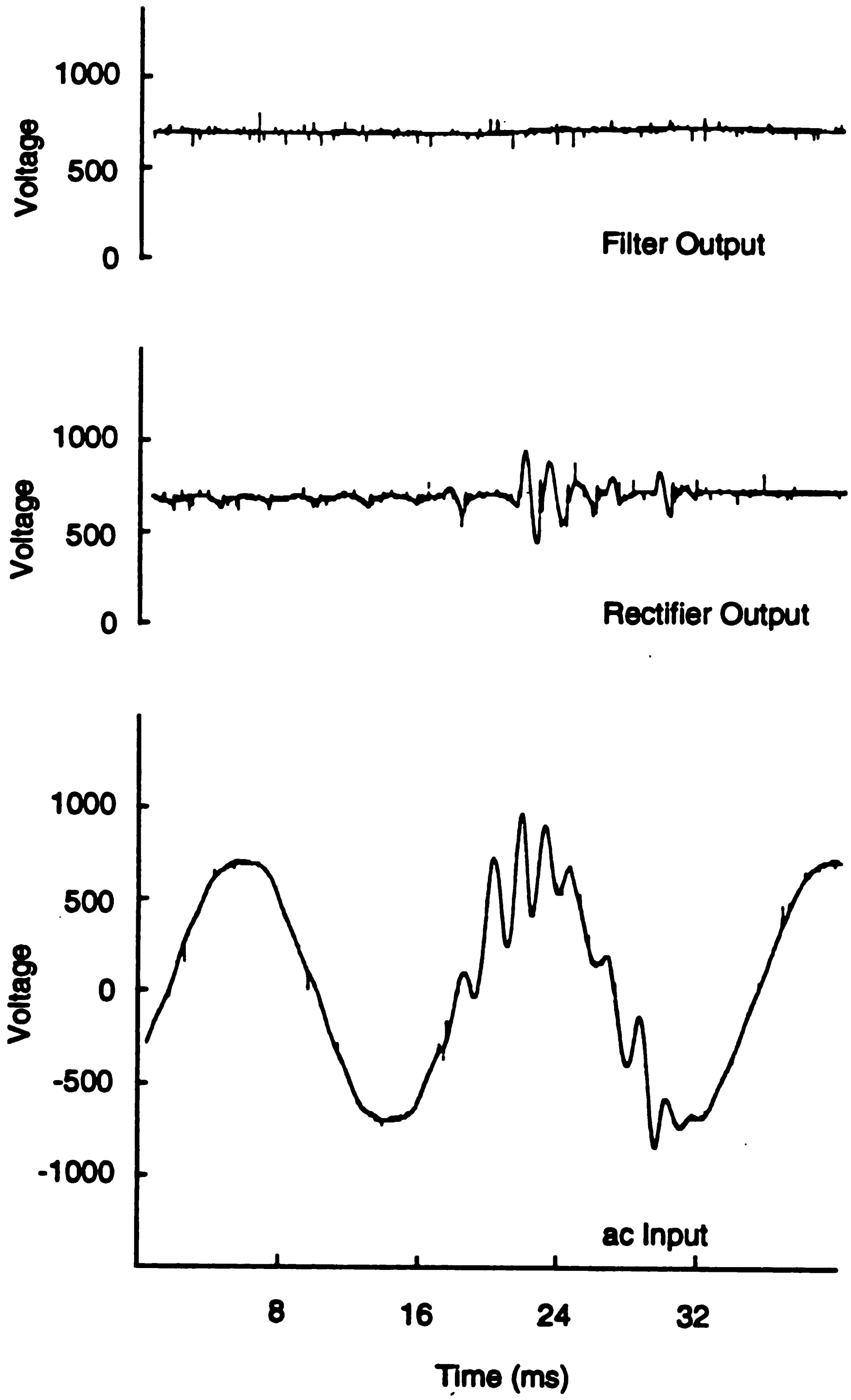

Figure 22. Laboratory test response of PWM drive

CHAPTER 5

SIMULATION RESULTS

To evaluate the effect of the capacitor switching transient on PWM drives, a drive was modeled with the SPICE electronic simulation program. The simulation provides a flexible means of exploring circuit behavior without the cost and time required for the equivalent hardware tests. First, the SPICE model is described and then the results of the simulations are presented. Finally, the simulation results are compared to the experimental results from Chapter 4.

5.1 Circuit Model

SPICE was originally developed as simulation program for low power analog circuits. For power electronic applications, two problems arise. First, the program must readily handle three phase sources with the flexibility and accuracy of single phase sources. Second, device models must be available in the current and voltage ratings common for power electronic applications.

The SPICE program selected for the drive modelling is a Macintosh based version from Intusoft compatible with Berkeley SPICE 2G.6. Intusoft provides a three phase voltage source, GEN3, that allows control of frequency, phase, amplitude, and balance. However, the number of high voltage and current semiconductor device models is limited. Consequently, the simulations were scaled for 240 V from the actual supply voltage of the field tests of 480 V.

The circuit modelled for the simulation is based on a Vee-Arc 10 hp PWM 480 V induction motor drive. To scale the drive for 240 V operation without adjusting the component current ratings, the motor size was reduced to 5 hp. The Vee-Arc has no field history or laboratory test evidence of capacitor switching transient susceptibility. The challenge of the simulation is to perturb a well performing drive so that it behave like a highly susceptible drive.

The circuit modelled in SPICE is shown in Figure 23 and includes the electrical source, the drive and the motor. An input listing for the simulation is contained in Appendix A. GEN3 creates a balanced three phase 240 V rms source for the circuit with the neutral resistor added to terminate the neutral node from the source. The source impedance is represented by the RS and LS phase elements and is identical to the source impedance for the 50 hp grinder drive from the field investigation described in Chapter 4. The capacitor switching transient is created by three voltage sources in series with GEN3. Seventeen milliseconds into the simulation, the transient voltage sources add a damped 24% transient at 450 Hz (see Figure 24). This transient is the same as the transient measured at the 50 hp grinder drive during the field investigation. The 17 ms delay allows the circuit to settle into steady state operation.

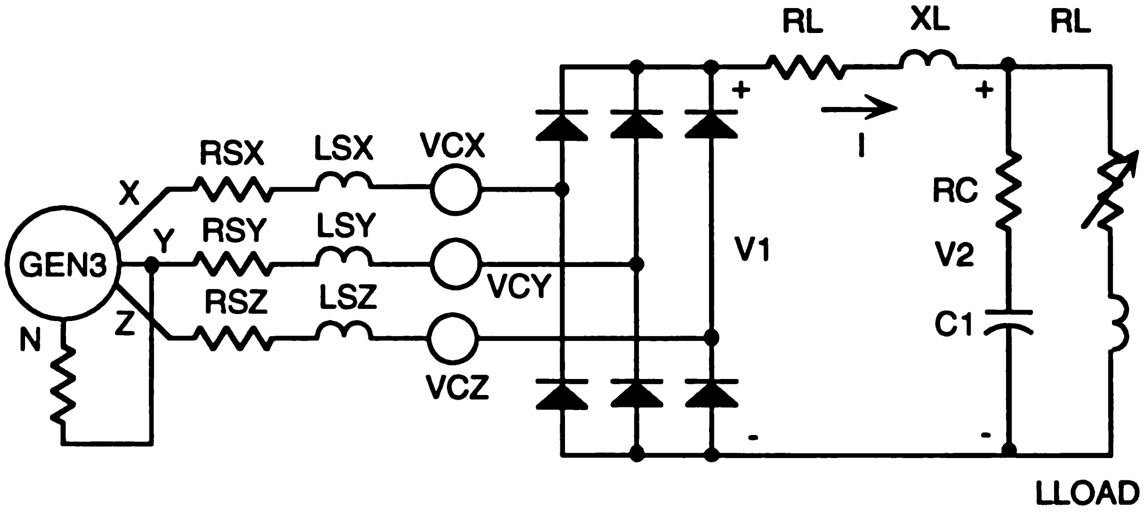

Figure 23. SPICE circuit diagram

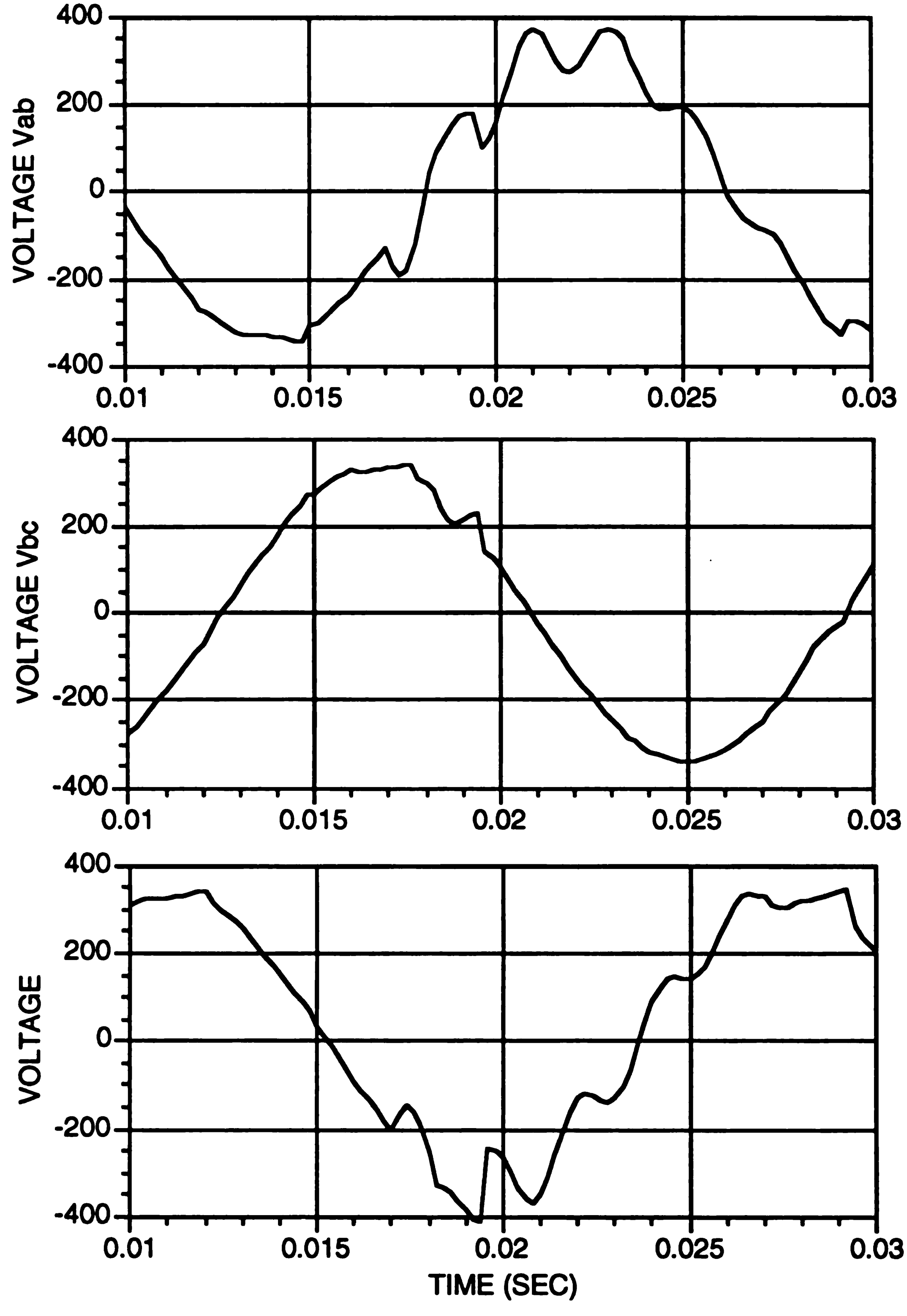

Figure 24. Drive ac input voltage

The current then enters a three phase full-wave bridge with uncontrolled rectifiers.

The LC filter uses a series inductor shunted by a capacitor with actual inductor and estimated capacitor losses included as equivalent resistances. Inductor saturation appears to be a factor in the transient performance of the field tested drive. Therefore, the simulation circuit includes a saturable inductor model for the filter inductor developed by Rumsey [11]. The desired behavior is obtained from linear circuit elements as shown in Figure 25. The voltage controlled current source G and capacitor CB integrate the applied voltage to produce an output voltage proportional to the flux at node 3. The voltage controlled voltage source E isolates the integration elements from subsequent stages. The current out of E is the inductor current which, when unsaturated, is determined by resistor RB. This current drives the current controlled current source F which produces the inductor's terminal current. Element F isolates the model's circuitry from the inductor terminals. When the current through RB exceeds the inductor saturation level, the voltage across the resistor is high enough to exceed VP and VN and forward bias diodes D1 and D2. The subsequent saturation current is determined by RS. Frequency dependent losses are modelled by the parasitic capacitance across diodes D1 and D2. That capacitance, viewed through the integration network, is seen as a frequency dependent resistance.

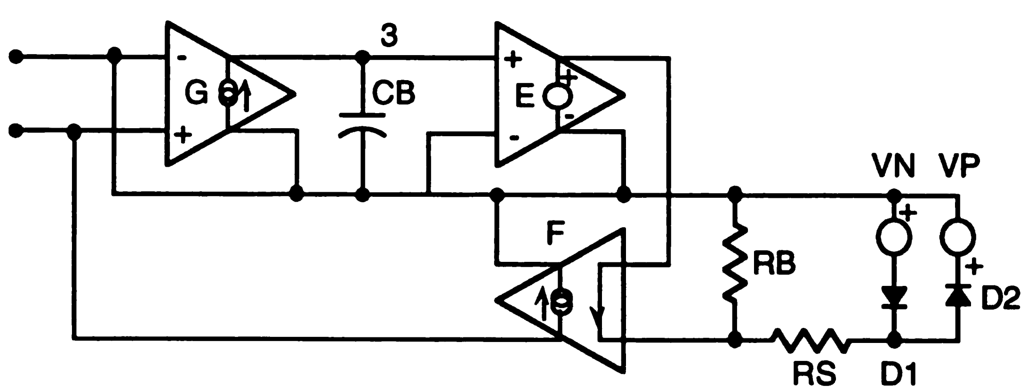

Figure 25. Circuit model for saturating inductor

Since voltage out of the LC filter is of primary interest, the drive inverter and motor were not directly modelled. Instead, a series resistor and inductor provides an equivalent load representing the 5 hp motor's mechanical output and leakage inductance. Two of the quantities of interest from the simulation results are input and output voltages for the LC filter (V1 and V2 in Figure 23) to evaluate the effectiveness of the filter and whether the inverter dc over-voltage limit is exceeded. The other quantity of interest is the inductor current (I in Figure 23) during the transient to determine if the diodes are damaged or over-current protection operates.

5.2 Simulation Results

The results of three cases are presented with different LC inductor operating modes: linear operation, saturation, and no inductor.

5.2.1 Drive with Linear Inductor

The results of the first case to be considered with the drive fully loaded and a linear inductor in the LC filter are shown in Figures 26 and 27. The input voltage to the LC filter exceeds 400 V yet the output shows little effect of the transient. Scaling down from 480 V drives, the typical dc trip level to protect the inverter would be 380 V. Over voltage protection would not be actuated in this case. There is a low frequency oscillation apparent in both the filter output voltage and the inductor current. This oscillation is due to the interaction of the filter capacitor and inductor which have a combined resonant frequency of 70 Hz. The 70 Hz voltage component does not appear at the ac source but does appear at the motor.

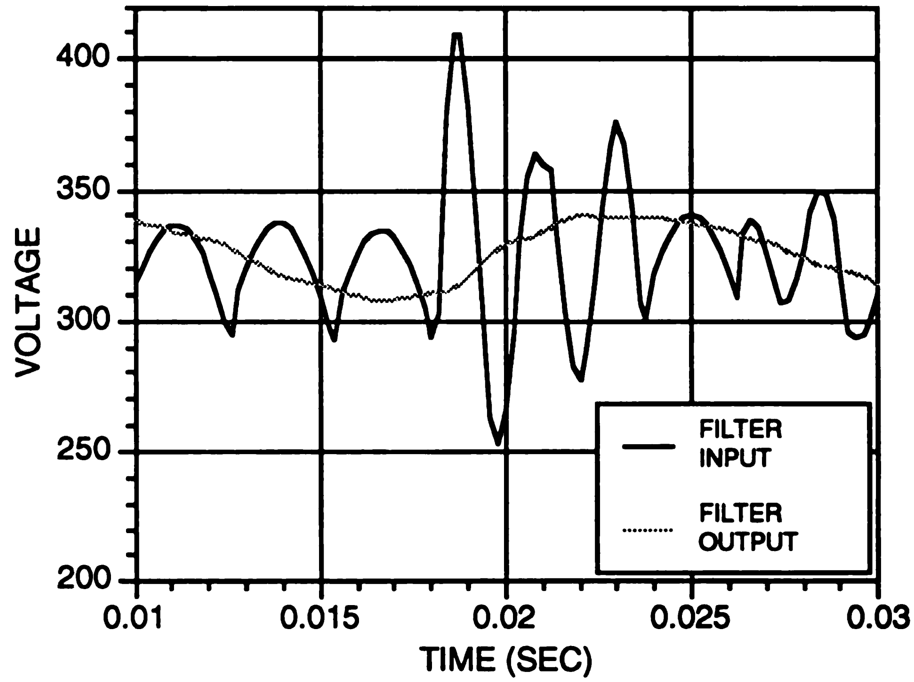

Figure 26. Filter transient voltage response with linear inductor

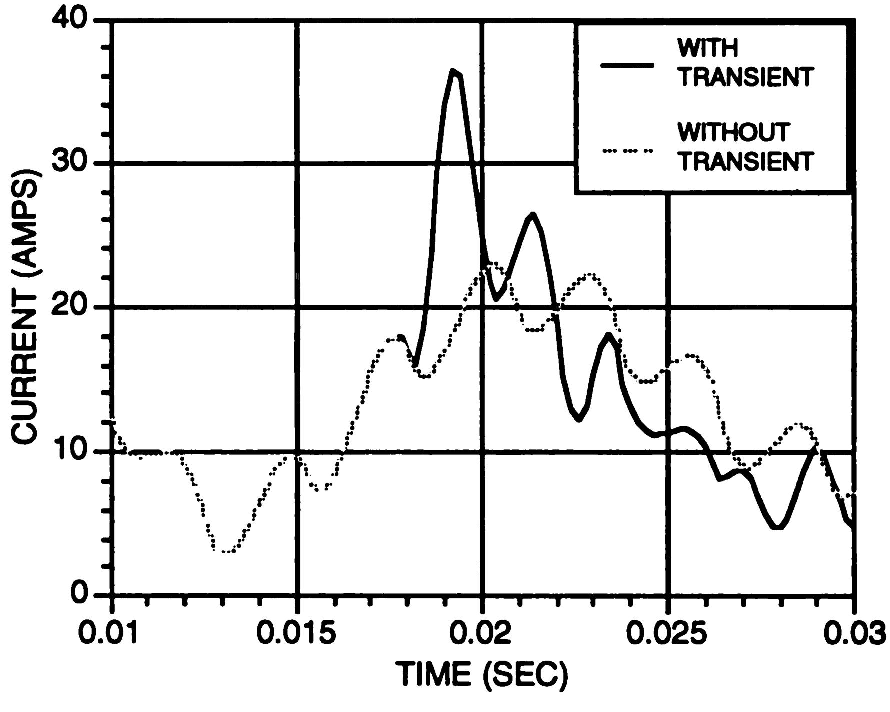

Figure 27. Inductor current with and without transient

To better demonstrate the effect of the transient, the inductor current graph shows the current with and without the transient. The capacitor switching transient at 17 ms shows a corresponding 8 A transient current. Since ac input current for the drive is about 14 A at full load, the peak would not cause over-current protective devices to trip the drive. However, because the filter capacitor acts as a short circuit at the transient frequency, the magnitude of that transient current is entirely dependent on the filter inductor. The transient current would be

$$i_t = \frac{V}{\omega L} \tag{24}$$

where V is the transient voltage (80 V in this case) at the transient frequency ω (450 Hz). For the above simulation, i_t would be 8.4 A which agrees with the results. If the LC filter were designed for a higher ripple than the Vee Arc, the inductor would be smaller and the current transient would be larger. Depending on the ac over-current devices utilized, the resulting transient current could trip the feed to the drive.

5.2.2 Drive with Saturating Inductor

The next case examines the consequence of designing the filter inductor to operate just below saturation at full load. A transient voltage applied to the inductor could readily force the core into saturation. An inductor was created with the Intusoft model with the resulting magnetization characteristics at the transient frequency (450 Hz) shown in Figure 28. The vertical axis is relative flux density and the horizontal axis is inductor current. The inductor core is assumed to be United States Steel M19 transformer iron with a saturation flux density of 1.0 T. The actual inductor has 20 cm^2 core area and 64 turns. The saturation level of the inductor was adjusted by varying the number of turns.

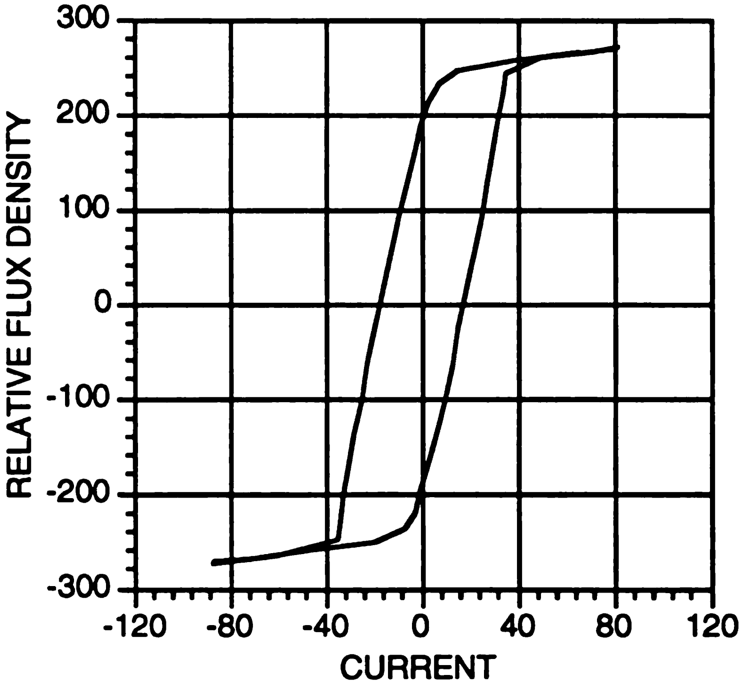

Figure 28. SPICE inductor B/H curve

The simulation result for the LC filter input and output voltages are shown in Figure 29. In this case the peak filter output voltage is 385 V and the over-voltage protection limit of 380 V is exceeded for 500 μs. The drive will trip because the voltage applied to the inverter section exceeds the over-voltage protection limits for magnitude and duration. Furthermore, since the capacitor voltage exceeds the peak LC filter input voltage, the transient voltage integrated across the filter capacitor. As described in Chapter 3, with the inductor saturated, the filter acts as a RC integrating network. The peak dc voltage is still less than the peak three phase transient voltage of 420 V (see Figure 24).

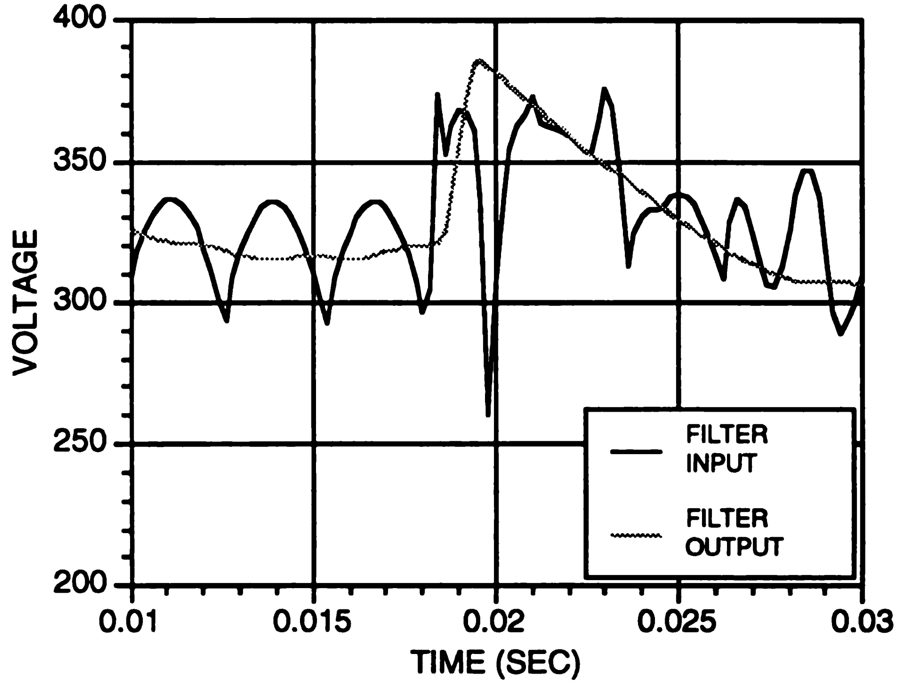

Figure 29. Filter transient voltage response with inductor saturation

The inductor current is shown in Figure 30 to peak at 148 A as a result of the transient. With the inductance reduced to almost a tenth of its non-saturated value, a very low impedance is presented to the voltage transient. The magnitude of the current transient is sufficient to operate over-current protection or damage diodes.

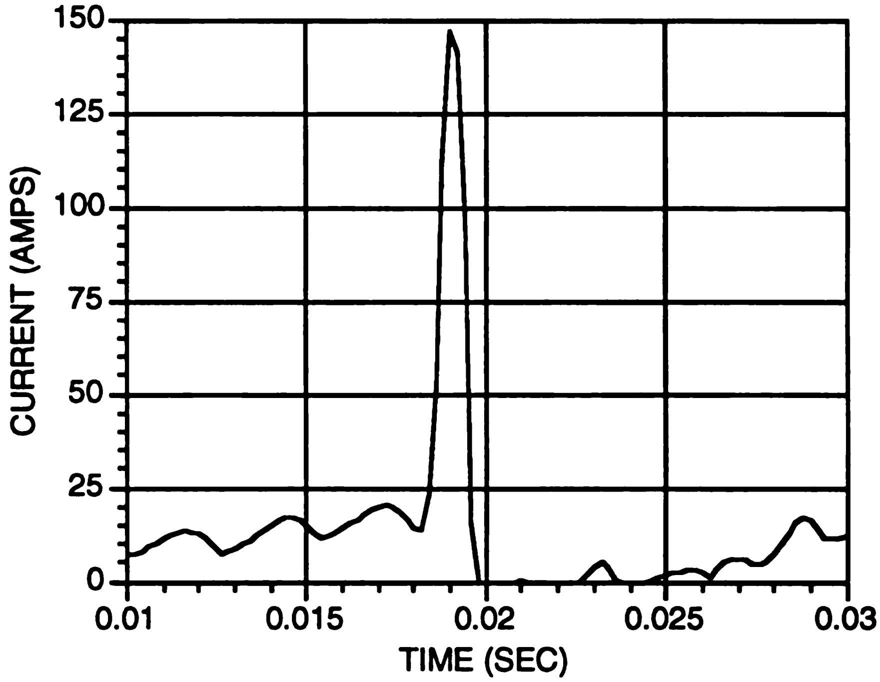

Figure 30. Inductor transient current with saturation

At light load the inductor current is reduced and the capability to operate with transient voltages without saturating is improved. Shown in Figure 31 are the LC filter input and output voltages with the same applied capacitor voltage transient except the drive is operating at 10% load. There is a slight increase in filter output voltage as the capacitor is charged to a greater voltage but not sufficient to interfere with drive operation. The 70 Hz oscillation is absent is this case because the diodes are not conducting continuously at light load.

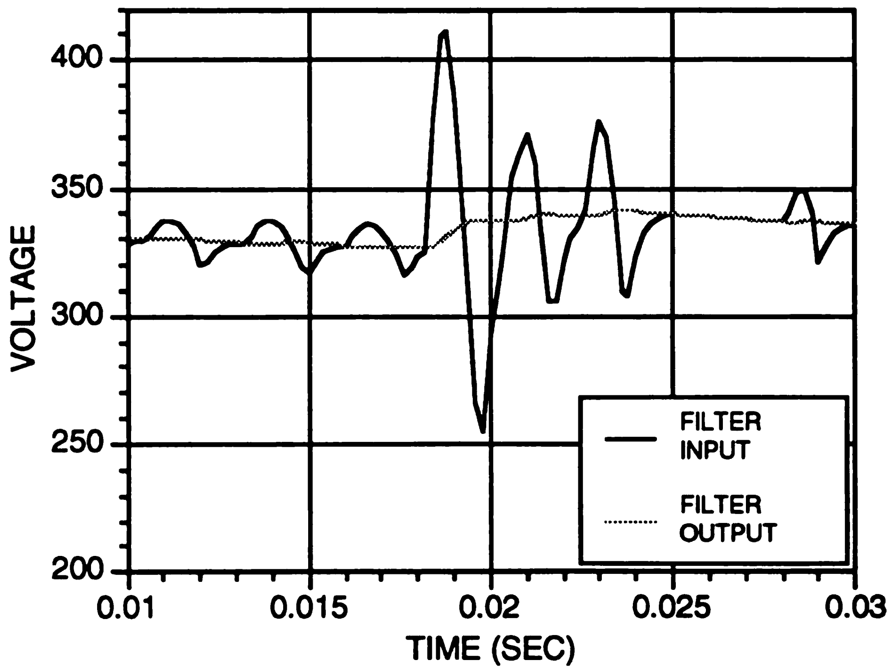

Figure 31. Filter transient voltage response with saturable inductor at light load

5.2.4 Drive with No Inductor

The last case to be considered is a drive with no filter inductor. The inductor is sometimes omitted in small drives to decrease cost. Although none of the field tests involved such a drive, they are sufficiently common to merit examination. Here, the filter input and output voltages are identical since there is only a shunt filter capacitor. Figure 32 shows the capacitor voltage as a result of the transient with other conditions identical to the previous cases. As in previous cases, the filter capacitor and parasitic resistances behave like an RC integrating network for the transient voltage. The dc voltage peaks at 420 V which is the magnitude of the three phase transient peak and more than enough to operate the inverter over voltage protection.

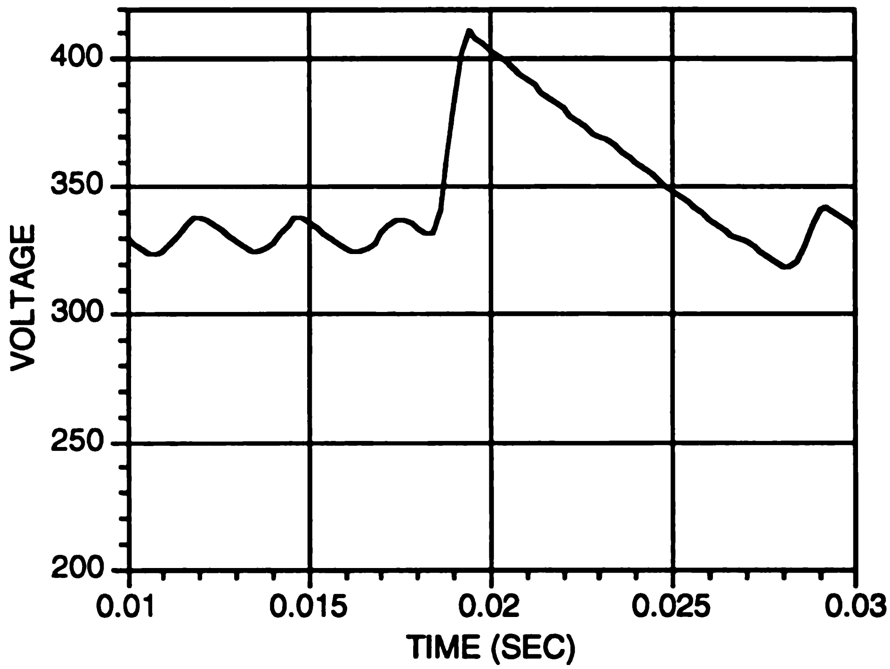

Figure 32. Filter transient voltage response without an inductor

The resulting dc current is shown in Figure 33. Since the filter capacitor presents a very low impedance to the transient frequency, the source impedance determines the resulting transient current. In this case it is very high and could operate over current protection or damage diodes. A drive without the filter inductor is very vulnerable to over voltage and over-current problems due to capacitor switching transient.

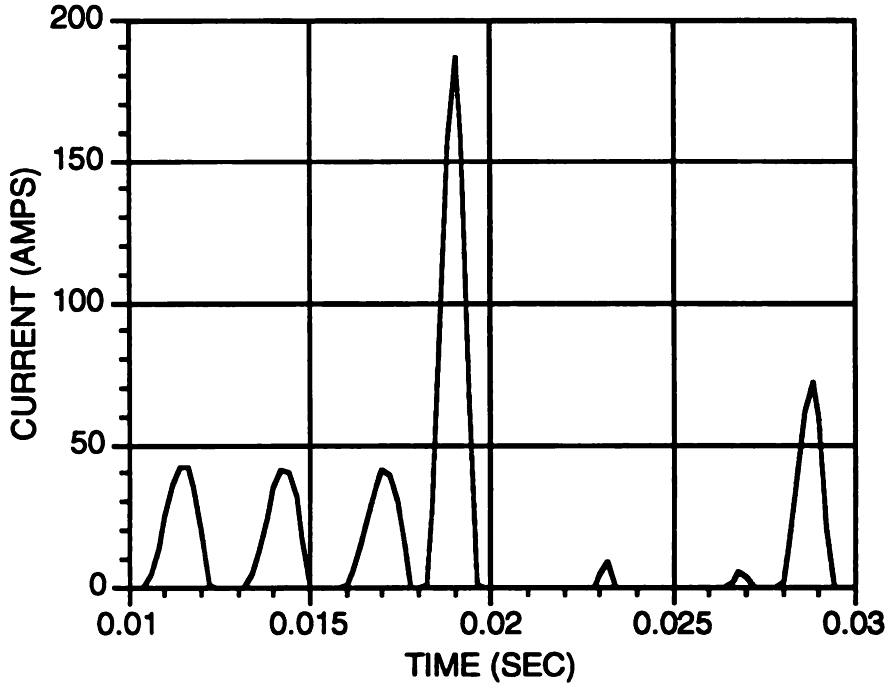

Figure 33. Filter transient current without an inductor

5.3 Comparison with Experimental Results

The simulation results from the previous sections are now compared to the experimental results from Chapter 4. The linear and saturated inductor simulations are compared to the results form field and laboratory tests. Since there were no tests of drives without a LC filter inductor, this simulation case is not compared.

5.3.1 Drive with Linear Inductor

If the linear inductor drive simulation results are compared with gantry drive results and the Vee-Arc laboratory test results, the dc transient voltage at the LC filter output is similar. The results from the gantry and Vee-arc drives are repeated here for the convenience of the reader in Figures 18 and 22. The 70 Hz voltage oscillation present in

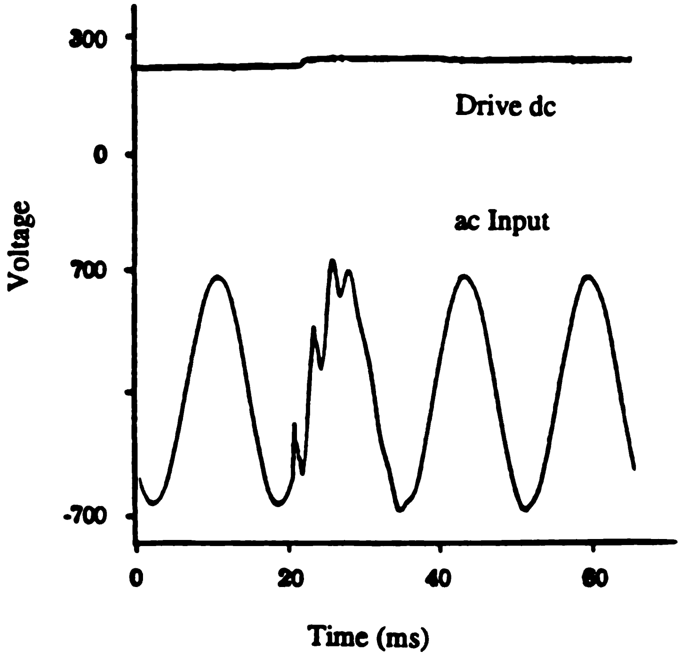

Figure 18. Gantry drive transient waveforms

the simulation output is absent from the experimental results. However, the gantry and Vee-Arc drives were tested at light load and the simulation showed that the 70 Hz oscillation is absent when the diodes are not continuously conducting at light load. If the experimental results are compared to the simulated light load case (Figure 31), the results are identical. The increase in filter output voltage is greater in the simulation than the laboratory test of the Vee-Arc drive. This may be because the voltage transient was only applied to one phase in the laboratory test whereas it was applied to all three phases in the simulation.

5.3.2 Drive with Saturating Inductor

The simulation of a drive with a saturating filter inductor tested the hypothesis that saturation was responsible for the observed response of the grinder drive from the field investigation to the capacitor switching transient. The LC filter output dc voltages for the

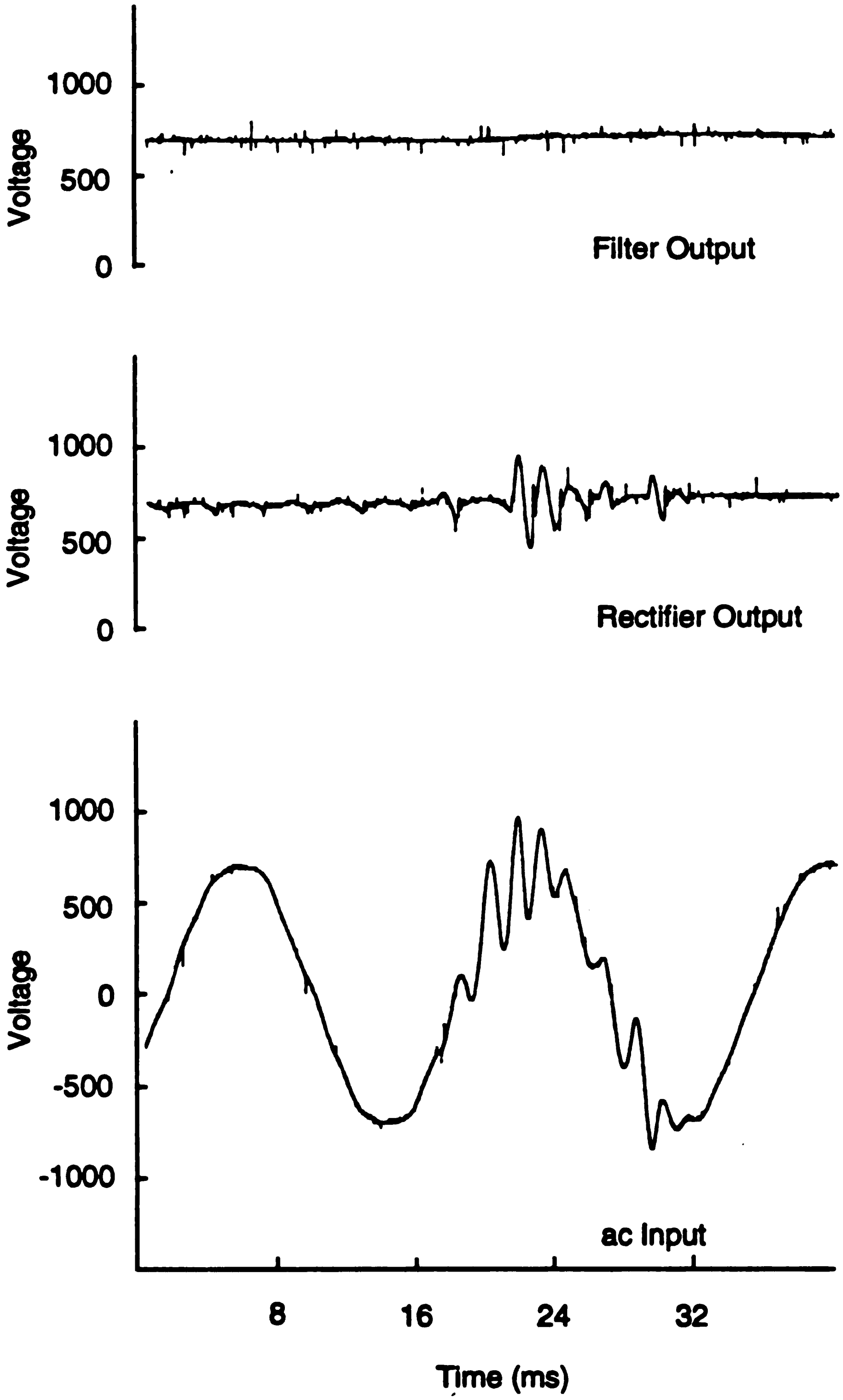

Figure 22. Laboratory test response of PWM drive

50 hp grinder drive and the corresponding simulation are repeated here in Figures 19 and 29. The simulation verified that when the voltage transient forces the filter inductor into saturation, the LC filter effectively passes the transient to the inverter section. After the dc peak voltage was reached in the simulation, the voltage quickly decayed to normal. For the actual drive, the over-voltage protection immediately trips the inverter and the peak dc voltage decays over a much longer time.

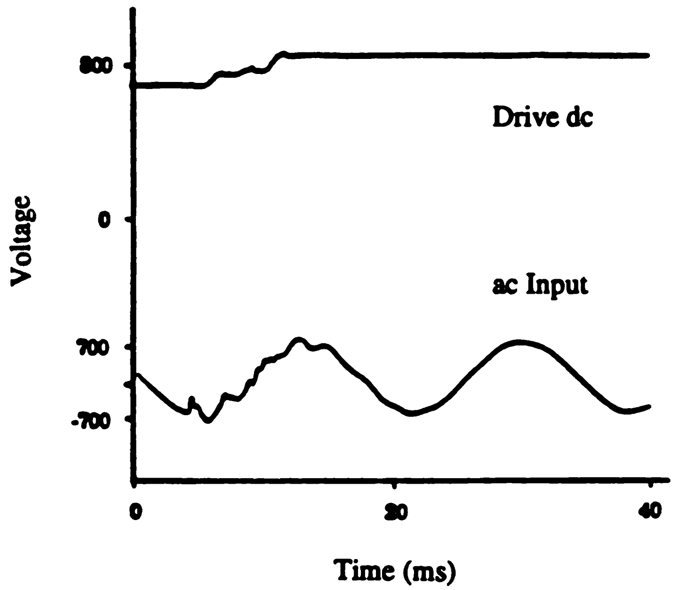

Figure 19. Grinder drive ac and dc transient waveforms

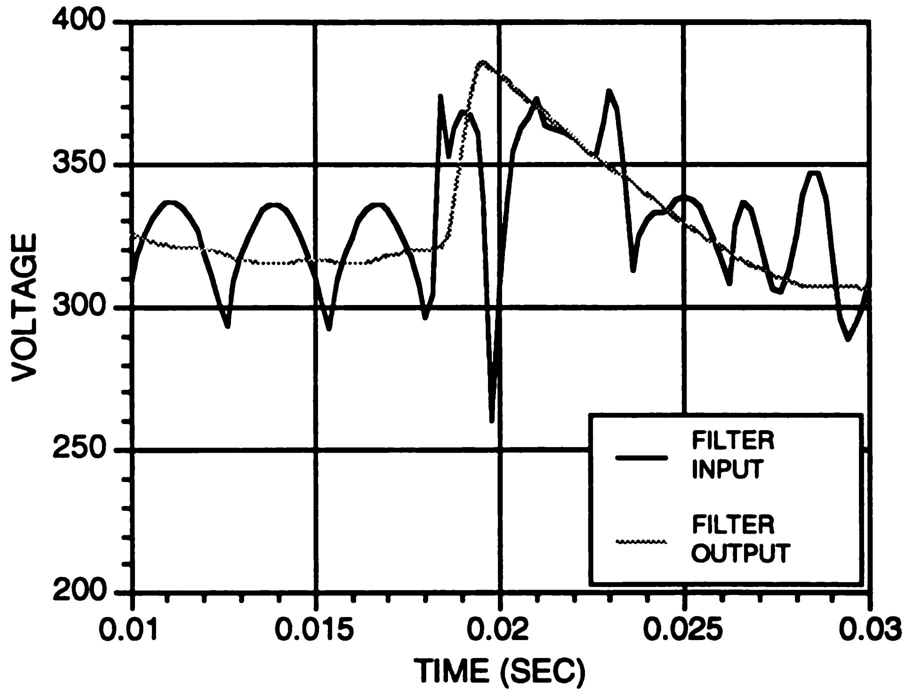

Figure 29. Filter transient voltage response with inductor saturation

CHAPTER 6

MITIGATION

There are three approaches to mitigating capacitor switching transient problems. The first is to reduce the magnitude of the transient at the utility capacitor that is switched. The second is to design the drive with the necessary characteristics to reduce susceptibility to the transient. Third, is to add a filter ahead of each drive to reduce the magnitude of the transient. The method selected depends on whether the issue is addressed at the equipment design, selection, or operating stage. Each of the methods is considered separately.

6.1 Utility Switching Options

The utility can reduce the magnitude of the transient with different capacitor switching methods. As utility capacitor banks increased in size in the 1950's and restrikes became an increasing problem, a common method of reducing the switching transient was to use pre-insertion resistors. These resistors were momentarily placed in series with the capacitor as the switch was closed and shorted afterwards to reduce losses (see Figure 34). However, situations where the capacitor bank was frequently switched (such as for testing) sometimes exceeded the thermal duty cycle rating of the resistors and caused failure. The development of special purpose capacitor switchers that eliminated the restrike problem eliminated as well the need for the resistors.

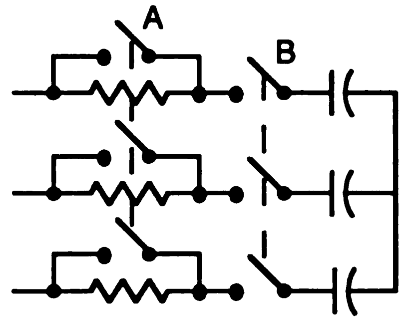

Figure 34. Operation of pre-insertion resistors

As the effect of the low level capacitor switching transients on equipment became apparent to utilities, they sought a solution with readily available equipment. One inexpensive solution was the pre-insertion inductor developed to prevent the capacitor bank inrush current for large banks from exceeding the switch capability. This device operated in the same manner as the pre-insertion resistor without the thermal problems. Rather than increase damping, the inductor shifts the transient frequency away from the resonant frequency of the intermediate networks that propagate the transient. As utility experience has shown, the change in transient frequency may not be sufficient in all cases to eliminate tripping of drives. The pre-insertion inductors for 40 kV capacitors are 3 mH and cause the greatest frequency change on low impedance networks. For the network feeding the automotive plant discussed in the field investigation, the pre-insertion inductor would decrease the resonance frequency by 15 % from the calculated value of 525 Hz. Clearly, pre-insertion inductors do not completely eliminate the problems due to capacitor switching for the utility

Although more expensive and not available for all voltage classes, a complete solution is offered by the synchronous capacitor switcher. This device closes the three phases separately with a vacuum switch as the voltage crosses zero almost eliminating the transient. The resulting transient can be calculated from equation (1) by adding the closing angle, τ, in the differential equation and solving for the capacitor voltage.

$$E\cos(\omega t-\tau)=L\frac{di}{dt}+Ri+\frac{1}{C}\int_{\tau}^{t} i\,dt \tag{25}$$

$$\begin{aligned} v=&\frac{E}{\omega CZ}\sin(\omega t-\theta-\tau)+\frac{Ee^{-\zeta\omega_n t}}{C\left[(\zeta\omega_n)^2+\omega_d^{\,2}\right]}\left[\frac{\cos(\theta+\tau)}{Z}(\zeta\omega_n\cos\omega_n t-\omega_d\sin\omega_n t)\right.\\ &\left.-\frac{1}{\omega_d}\left(\frac{\cos\tau}{L}-\frac{\omega\sin(\theta+\tau)}{Z}-\frac{\zeta\omega_n\cos(\theta+\tau)}{Z}\right)(\zeta\omega_n\sin\omega_n t+\omega_d\cos\omega_n t)\right]\end{aligned} \tag{26}$$

For example, a synchronous switch installed at the east utility station from the field investigation would produce the transient shown in Figure 35 plotted from (26). The plotted closing angle is 0.3 ms later than the voltage zero crossing which is the switch manufacturer's specified tolerance for operation. Such a device would reduce the capacitor switching transient to such a low level, that it would no longer be a problem.

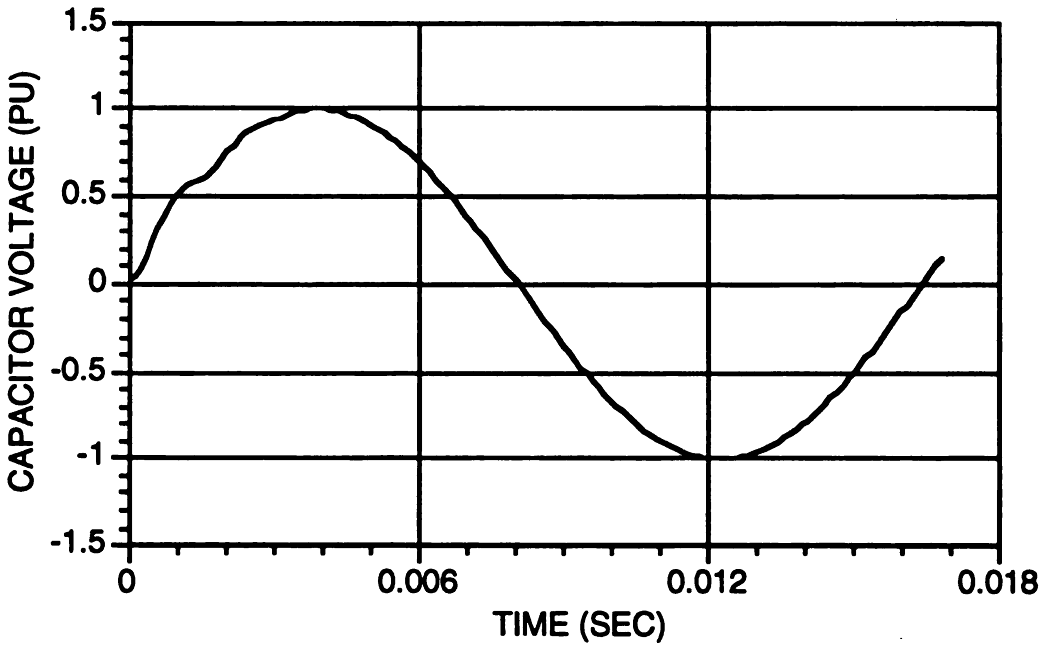

Figure 35 Utility capacitor voltage with synchronous closing

6.2 Drive Design Options

It is possible to design the drive to reduce susceptibility of the capacitor switching transient. The most important design factor in the performance of the drive is the filter inductor capability. The results of the drive simulation in Chapter 5 demonstrate how the inductor affects transient voltage and current. If the inductor is present and does not saturate, the transient voltage and current are limited to reasonable levels that should not cause the drive to malfunction. If the inductor saturates during the transient, the transient voltage integrates across the filter capacitor until it exceeds the dc over-voltage trip threshold. The resulting transient current may operate over-current protection or damage diodes. If the filter inductor is omitted, the same over-voltage and over-current problems exist. The inductor should be designed according to equation (23) which is repeated here.

$$B = \frac{0.4\pi N I_{dc} \cdot 10^{-4}}{l_g} + \frac{V_R}{4.44\omega_R A_C N \cdot 10^{-8}} \tag{23}$$

The inductor dc current, I_{dc}, should be the maximum expected under maximum load and low ac voltage conditions. The ripple voltage, V_R should perhaps be assumed to be a maximum of 1 pu of the nominal ac voltage at the ripple frequency. From (23), the core area, A_C, affects the ripple voltage capability without changing the dc flux density. By selecting the appropriate core area, the vulnerability of the drive to the capacitor switching transient can be minimized.

6.3 Transient Filters

There are situations where a susceptible drive is already in use and the source of the capacitor switching transient is unalterable. In this case a filter may be added to the drive in the form of an ac reactor ahead of the drive. Unlike noise filters with cut off frequencies of approximately 10 kHz, this reactor must filter frequencies that closely correspond to harmonic current frequencies of the drive. To prevent excessive insertion loss, the reactor must be customer engineered for the drive and sized from 2 % to 3 % of the drive base impedance.

To illustrate, a SPICE simulation was conducted with ac reactors, designated LR, shown in Figure 36. These were sized to 2% impedance which corresponds to 562 μH and are between the transient voltage source and the drive The transient voltage into the drive terminals from the simulation is shown in Figure 37. The transient magnitude has been reduced especially at the voltage crests where diode conduction occurs. However, the voltage waveform has been flat-topped by the reactors. Since diode conduction occurs at the voltage crest, the corresponding voltage drop across the reactors causes the flat-topping. As a result of the flat-topping, the drive is operating at reduced voltage, which reduces the drive's capability to withstand momentary voltage sags without tripping.

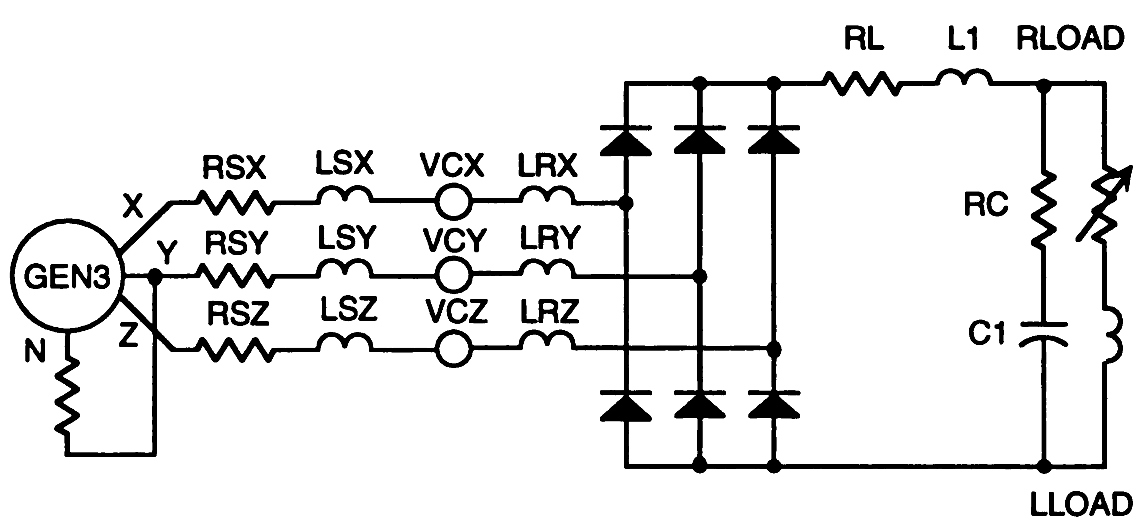

Figure 36. SPICE model with ac reactors added

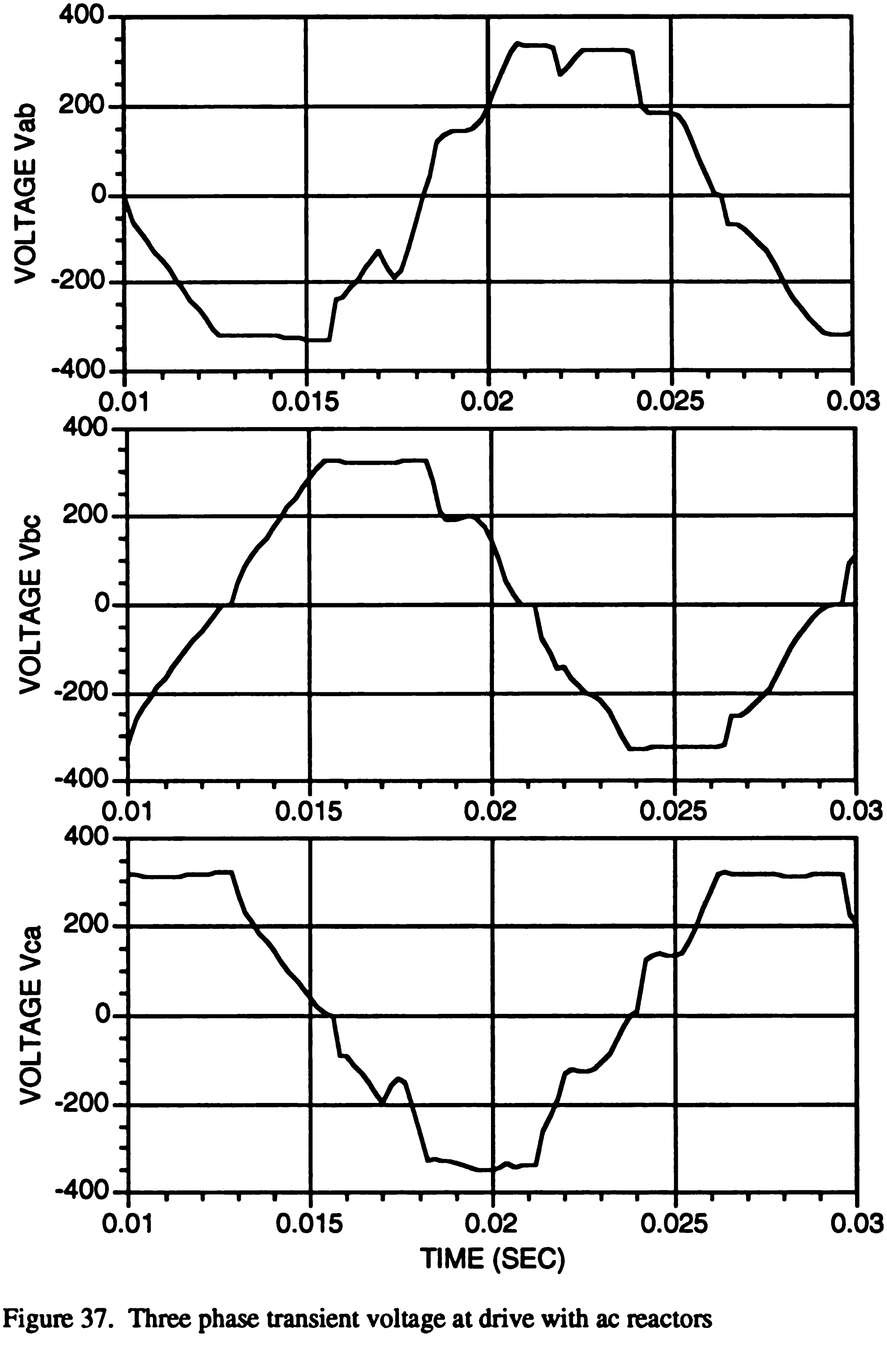

Figure 37. Three phase transient voltage at drive with ac reactors

Two SPICE simulations were run to determine the effect of the ac reactors to protect the drive from the capacitor switching transient. The first was the fully loaded drive with saturating filter inductor and the second was a fully loaded drive without a filter inductor. The drive dc voltages and currents as a result of the transient are shown in Figures 38 through 41. In each case the dc voltage entering the inverter was well below the over-voltage trip threshold and transient currents were below levels that could cause damage to the drive.

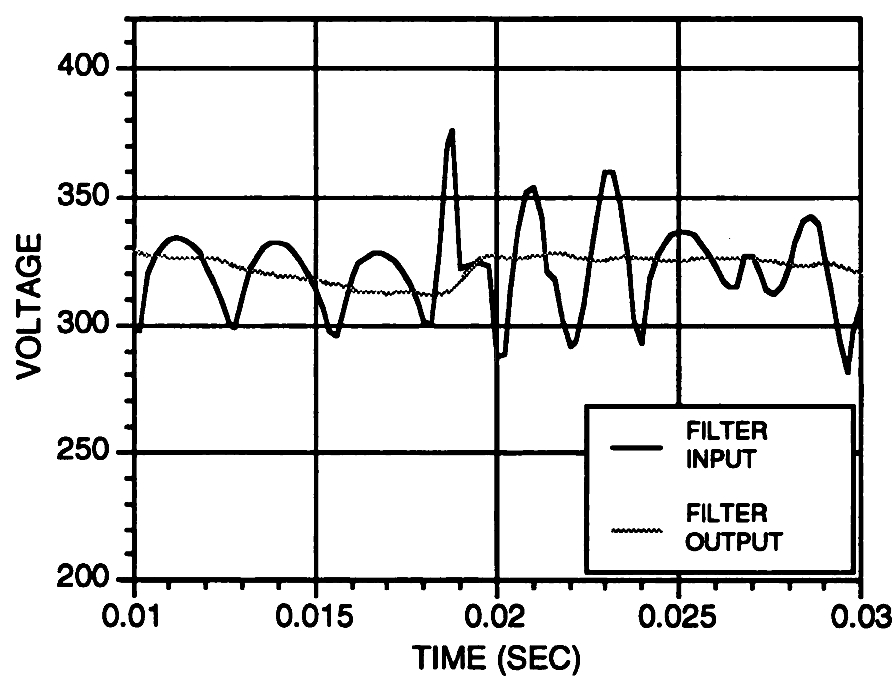

Figure 37. Filter transient voltages with a saturable dc inductor and ac reactor

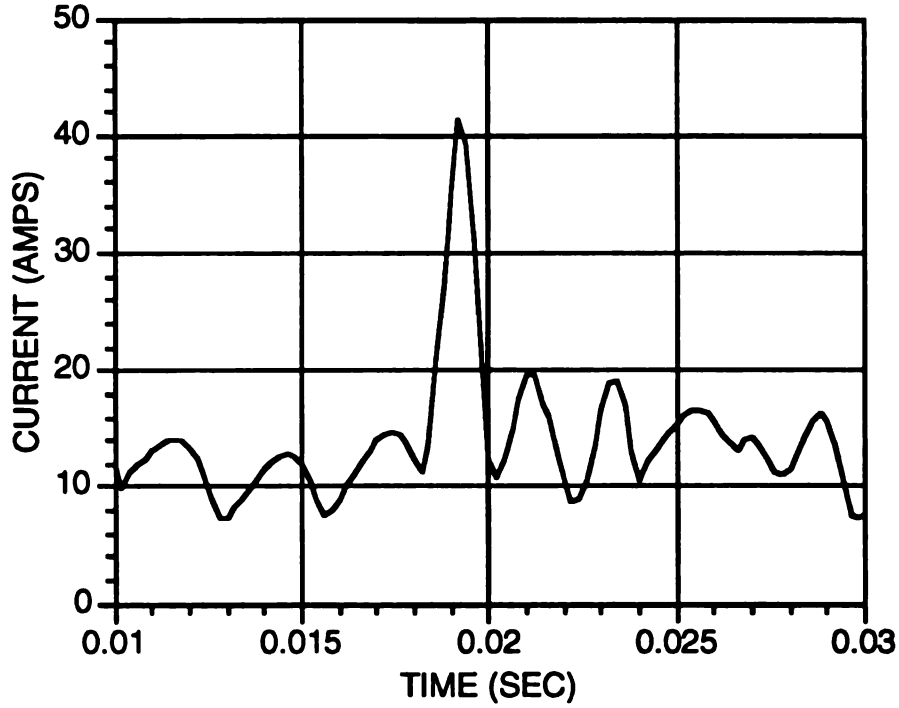

Figure 39. Drive dc current with saturable dc inductor and ac reactor

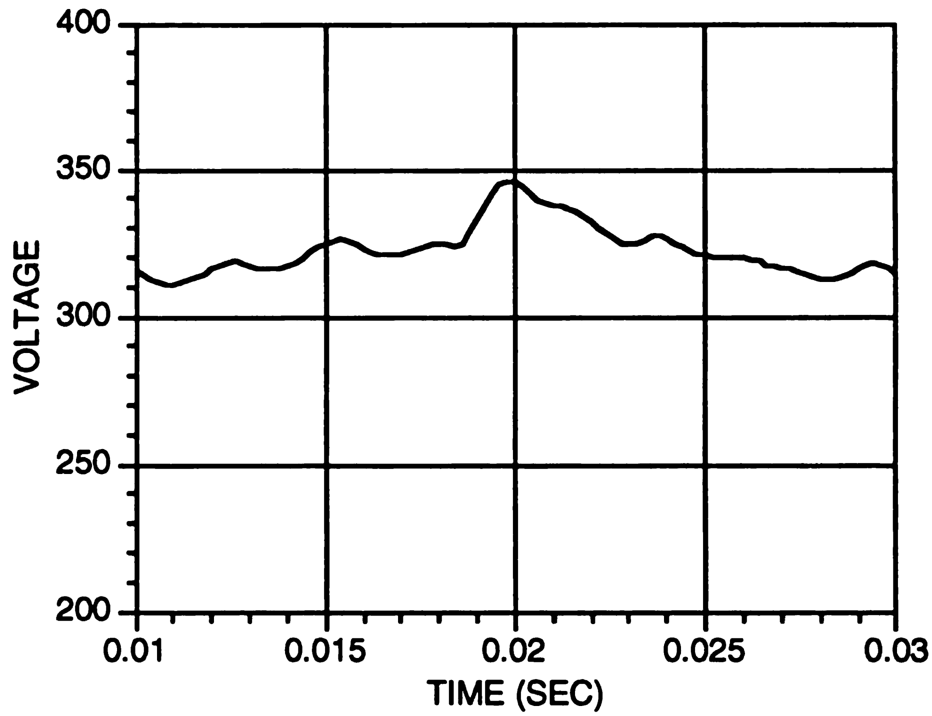

Figure 40. Drive dc transient voltage with ac reactor and no dc inductor

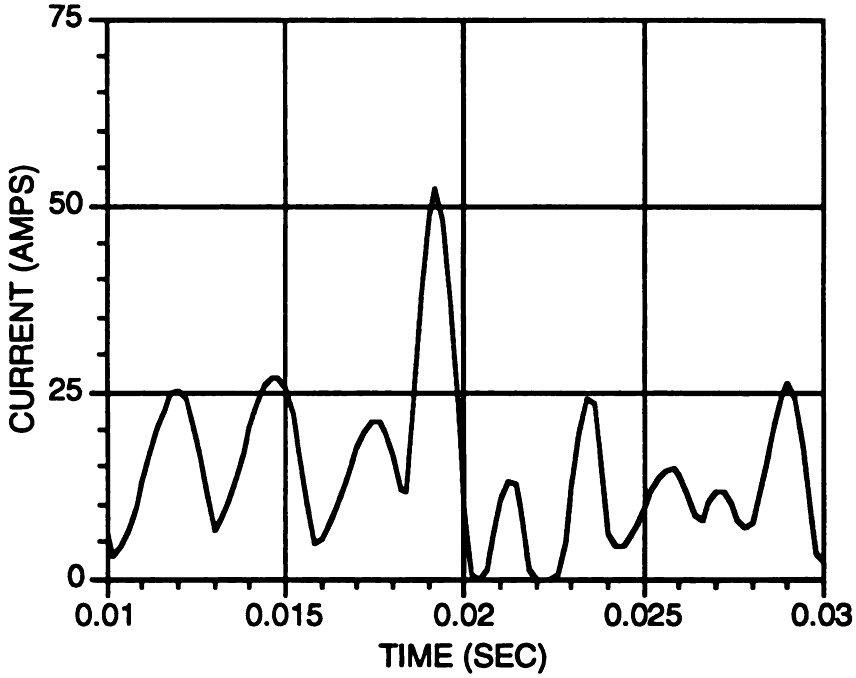

Figure 41. Drive dc current with ac reactor and no dc inductor

CHAPTER 7

CONCLUSIONS

The preceding analysis, experimentation, and simulation explored how ac drives with uncontrolled rectifiers respond to the capacitor switching transient. It also addresses how the transient is generated at the utility capacitor and propagates to the drive. The results may be summarized by the following:

- Due to the low value of damping inherent in utility systems, the peak value of the capacitor closing transient is always near the maximum theoretical value of 2 pu at the capacitor bus.

- The transient propagates only if the transient frequency is within the bandwidth of the successive voltage step-down networks.

- PWM drives can be disrupted by the transient either by dc over-voltage at the inverter input or by a current surge through the drive filter capacitor. The over-voltage triggers an over-voltage trip to protect the inverter transistors. The current surge can operate over-current devices or damage diodes.

- A properly designed filter inductor that does not saturate during the transient minimizes susceptibility of the drive. If the inductor saturates during the transient, the transient voltages integrates across the filter capacitor. If the filter inductor has been omitted from the drive, the filter capacitor operates in a similar manner. In both cases, a large current surge is generated by the filter capacitor as a result of the voltage transient.

- The most effective method to reduce the magnitude of the transient at the utility capacitor is to use synchronous switchers that close each phase separately at the voltage zero-crossing.

- A less desirable method is to add ac reactors ahead of the drive to filter the voltage transient. Although these reduce the transient at the drive, they also cause flat-topping of the ac voltage which increases the drive susceptibility to under-voltage problems.

APPENDIX A

APPENDIX A

SPICE MODEL

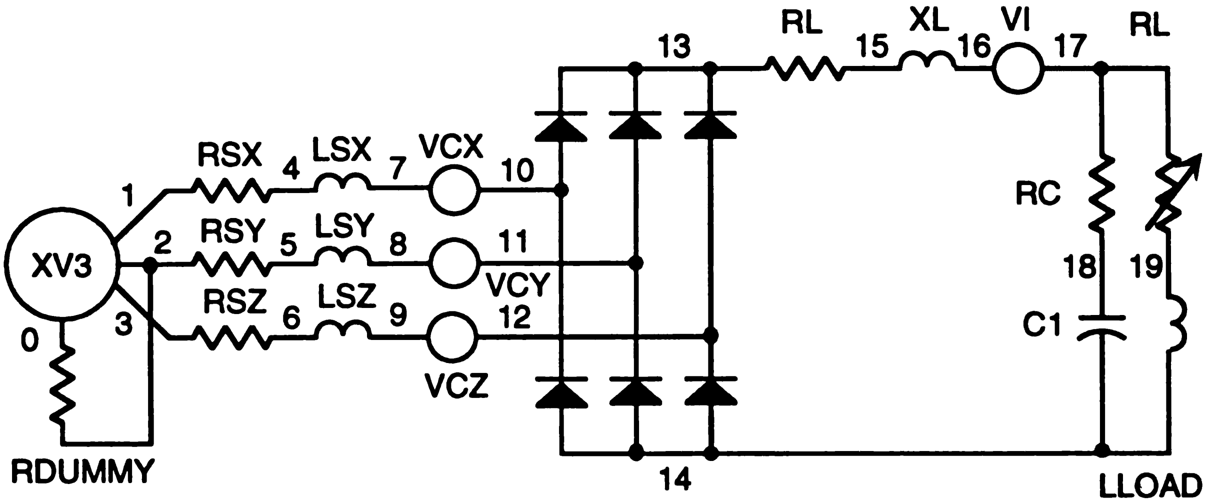

Figure 42. SPICE PWM drive circuit with node numbers

Table 5. SPICE PWM drive circuit data

```
PWM CAP SWITCHING TRANSIENT RESPONSE
*LAST NODE: 20
*
* 9/19/92 - full load
*
*INCLUDE DIODE.LIB
*INCLUDE SIGNAL.LIB
*INCLUDE DEVICE.LIB
*OTHER THAN LIMPTS, THESE OPTIONS AID PWR ELECTRONICS SOLUTIONS
.OPT ABSTOL=1UA VNTOL=1UV LIMPTS=505 METHOD=GEAR RELTOL=.03 ITL4=500
+ ITL5=14000
.TRAN 0.2MS 30MS 5MS 0.05MS UIC
.PRINT TRAN V(13,14) V(18,14) I(VI), V(20,16)
```

Table 5 (cont'd)

```
*
*3-PHASE VOLTAGE SOURCE SUBCIRCUIT CALL
XV3 1 2 3 0 GEN3 {VGEN=196 FREQ=60 MAGERR=0 PHASE=0}
*
*THIS ADDS THE CAP SWITCH TRANSIENT
VCX 7 10 SIN 0V 80V 450HZ 17MS 125
VCY 8 11 SIN 0V -40V 450HZ 17MS 125
VCZ 9 12 SIN 0V -40V 450HZ 17MS 125
*
VI 16 17
*
*RDUMMY CONNECTS THE NEUTRAL WHICH IS NOT USED IN THE DRIVE
RDUMMY 2 0 1K
D1 10 13 DN3495
D2 11 13 DN3495
D3 12 13 DN3495
D4 14 10 DN3495
D5 14 11 DN3495
D6 14 12 DN3495
*
*CAPACITOR WITH LOSSES (RC)
C1 18 14 0.0015 IC=310V
RC 17 18 .05
*
*NONLINEAR INDUCTOR CALL WITH LOSSES (RL)
*RLD CONNECTS THE FLUX TEST POINT
XL 15 16 20 CORE {VSEC=.130 IVSEC=.000 LMAG=3.36MH LSAT=43UH FEDDY=1KHZ}
RL 13 15 .041
RLD 20 16 100K
*
*THIS SIMULATES MOTOR THRU THE DRIVE INVERTER
RLOAD 17 19 22
LLOAD 19 14 29.1MH IC=10A
*
```

Table 5 (cont'd)

```
*SOURCE IMPEDANCE
RSX 1 4 2.56M
RSY 2 5 2.56M
RSZ 3 6 2.56M
LSX 4 7 0.02MH IC=0A
LSY 5 8 0.02MH IC=-10A
LSZ 6 9 0.02MH IC=10A
.END
```

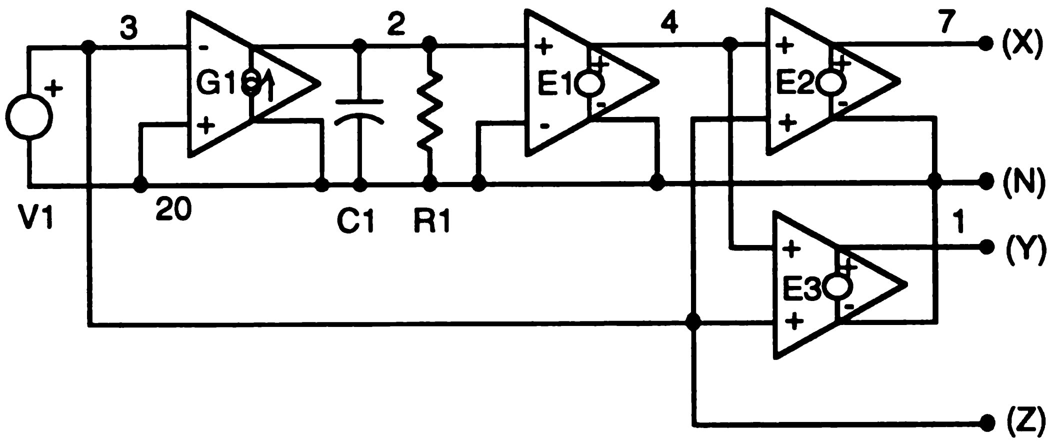

Figure 43. SPICE three-phase voltage source GEN3

Table 6. SPICE three-phase voltage source GEN3 data

```
*SYM=GEN3
.SUBCKT GEN3 3 7 1 20
*3 PHASE GENERATOR
* FREQ = {FREQ}
* AMPLITUDE = {VGEN}
C1 2 20 {1/(6.28319K*FREQ)} IC={VGEN}
R1 2 20 1E6
I1 20 2 PULSE {VGEN*1U} 0
* MAKES UIC UNNECESSARY
E1 5 20 20 2 1
V1 3 20 SIN 0 {VGEN} {FREQ}
```

Table 6 (cont'd)

```
E2 7 20 POLY(2) 5 20 3 20 0 -866.00M -500.00M
E3 1 20 POLY(2) 5 20 3 20 0
+ {(1+.01*MAGERR)*(.866*(1-.5*(.0174533*PHASE)^2)-
.5*.0174533*PHASE*(1+.166667*(.0174533*PHASE)^2))}
+ {(1+.01*MAGERR)*(-.5*(1-.5*(.0174533*PHASE)^2)-
.866*.0174533*PHASE*(1+.166667*(.0174533*PHASE)^2))}
G1 20 2 20 3 1M
R2 7 0 100MEG
R3 1 0 100MEG
R4 3 0 100MEG
R5 5 0 100MEg
.ENDS
```

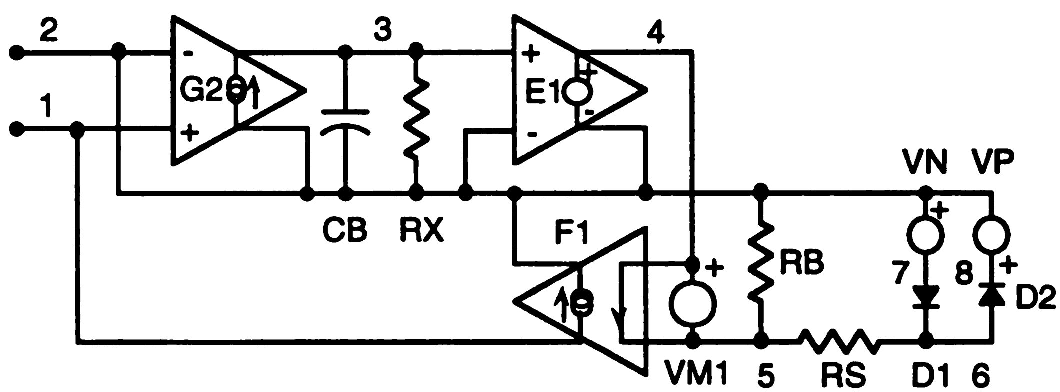

Figure 44. SPICE nonlinear inductor model CORE

Table 7. SPICE nonlinear inductor model CORE data

```
*SYM=CORE
*PARAMS ARE
*  VSEC, VOLT-SEC CORE CAPACITY (2Bm)
*  IVSEC, INITIAL CONDITION
*  LMAG, MAGNETIZING INDUCTANCE
*  LSAT, SATURATION INDUCTANCE
*  FEDDY, FREQUENCY WHEN LMAG REACTANCE = LOSS RESISTANCE
* EXAMPLE CALL
*  XCORE1 5 6 7 CORE {VSEC=50U IVSEC=-25U LMAG=10MHY LSAT=20UHY
```

Table 7 (cont'd)

```
  + FEDDY=20KHZ}
*
.SUBCKT CORE 1 2 3
RX 3 2 1E12
* EXAMPLE MAKES CB=1E-7, IC=-250
CB 3 2 {VSEC/500} IC={IVSEC/VSEC*500}
F1 1 2 VM1 1
G2 2 3 1 2 1
E1 4 2 3 2 1
VM1 4 5
* EXAMPLE MAKES RB=100K
RB 5 2 {LMAG*500/VSEC}
RS 5 6 {LSAT*500/VSEC}
* EXAMPLE MAKES RS=200
VP 7 2 250
D1 6 7 DCLAMP
VN 2 8 250
D2 8 6 DCLAMP
*EXAMPLE MAKES CJO=238PF, MULTIPLIER 3 AND VJ=25 GO TOGETHER
.MODEL DCLAMP D(CJO={3*VSEC/(6.28*FEDDY*500*LMAG)} VJ=25)
.ENDS
```

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