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ON DYNAMIC VOLTAGE RESTORATION AND ITS INTERACTIONS WITH LOADS

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School of Electrical and Electronic Engineering

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SUMMARY

Power quality problems have received increasing attention in recent years because they impact greatly on the operations of electrical equipment. Often the consequence would be significant economic losses to customers. One method to enhance electricity supply quality is based on the use of series Custom Power compensation technique, such as the Dynamic Voltage Restorer (DVR). A restorer is intended for use in a distribution system, with rating of between a few hundred kVA to 2 MVA typically. Central to its design is a voltage-source inverter connected to a DC energy-storage device. The restorer is designed to be installed upstream of loads which are to be protected against short-duration voltage disturbances.

The basic operating principle of a DVR is by injecting a voltage component in series with the load voltage so that the load-side voltage is maintained at its nominal level during a voltage disturbance. Due to its quick response and high reliability, the DVR is considered to be an effective device in mitigating the impacts of voltage sag and can provide good power quality for customers. The main focus of this project is to examine issues closely related to the design and applications of such a series compensator (SC).

Currently, most studies on series compensation are focused on applying the technique in mitigating the effects of voltage sags. However, power quality problem due to voltage swell has not been so closely investigated. This thesis therefore begins with a detailed analysis concerning the more general case of the mitigation of the effects of voltage sag/swell through a series compensator. Based on the analysis it is noticed that most of voltage swell events can cause the compensator to absorb energy from the external system. It would then result in a rise in the dc-link voltage of the inverter. By permitting phase adjustment in the injected voltage, a generalized compensation method which is suitable for both sag and swell is proposed. The new voltage injection technique applies to both balanced as well as unbalanced voltage disturbances. It allows the magnitude of the positive phase sequence component of the compensated load voltage to be restored to its pre-sag/swell level.

Despite the generalized voltage compensation strategy is shown to be advantageous as it can increase the protected load ride-through capability for sag/swell mitigation, a step-type change in the load terminal voltage phase is inevitably introduced. A study on the load response under such voltage phase shift is then described. Analysis shows that static linear load and induction motor will exhibit drastically different dynamic responses under the phase shift. Analytical expressions have been developed to quantify the load response in term of the voltage injection magnitude and load parameters. From the analysis and supported by simulation and laboratory measurements, it is seen that over the interval of the voltage injection, the induction motor stator current is up to some 5-6 times larger than its rated value. In order to reduce this large transient current, one solution called “an exponential phase shift method” has been proposed. Simulation results indicate that using the suggested phase shift method, the generalized compensation strategy can increase the ride-through capability afforded by the SC while minimizing the load current oscillations.

Due to convenience and simplicity, the step-type voltage injection technique is nevertheless commonly used in SC. Earlier analysis shows that it inevitably leads to large load transients being introduced into the power system. Since the restorer is series-connected to the load, it must also experience similar load transient currents. The consequence is increased losses in the switching elements of the restorer. This is undesirable as it may damage the DVR components. A computational method has been proposed for which the transient currents and the associated losses are determined. The derived thermal model then permits an assessment to be made on the resulting temperature rise within the restorer. Based on the suggested method, the switching devices in the DVR can be sized readily in order to alleviate the thermal problem due to the load transient currents. The results of this work would be particularly useful for the design and sizing of an SC.

Voltage sags are usually caused by two main types of events: short circuit faults or the switching on of large loads such as induction motors. Presently, most of the studies concerning SC applications are based on the assumption that the voltage sags are due to short-circuit faults. Sag compensation due to motor-start has also been studied in this project. During a start-up, an induction motor might draw some five to seven times larger current than normal. This may damage the SC due to the thermal loading on the inverter

switching devices or it can cause a severe depletion of the stored energy in the restorer. A method to estimate the peak starting current and the required energy for supporting the motor-start is given. The proposed method allows one to size the restorer to ensure that it is able to mitigate such type of voltage sag.

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List of Symbols

α :	Phase angle of $\vec{V}_{inj}$ with respect to $\vec{V}_{sag}$.
ϕ :	Load power factor angle.
θ :	Phase shift between $\vec{V}_{pre}$ and $\vec{V}_s$.
ψ :	Phase difference between $\vec{V}_{pre}$ and $\vec{V}$.
φ :	Phase angle of the sustained source-side voltage ($\vec{V}_{sag}$) with respect to the load current $\vec{I}$ during the sag.
β :	Phase difference between $\vec{I}$ and $\vec{V}_{inj}$.
P_{inj} :	Injected active power from the DVR.
V_d :	DC-link output voltage of the DVR.
$V_{ip1}I$:	Active power injection from DVR for in-phase compensation.
$V_{ip2}I$:	Active power injection from DVR for energy-saving compensation.
$\vec{V}_{pre}$:	Pre-sag load-side voltage.
$\vec{I}_{pre}$:	Pre-sag load-side current.
$\vec{V}_L$:	Load-side voltage phasor during voltage sag/swell.
$\vec{V}_s, \vec{V}_{sag}$:	Source-side voltage (or point-of-common coupling bus voltage).
$\vec{V}_{s,pps}$:	Positive phase sequence component of $\vec{V}_s$.
$\vec{V}_{inj}$:	Injected voltage phasor from DVR.
$\vec{I}$:	Load-side current phasor during voltage sag/swell.
$\vec{Z}_s$:	Source-side impedance.
$\vec{Z}_{Load}$:	Load-side impedance.
$\vec{Z}_f$:	Fault impedance.
$\vec{V}_1, \vec{V}_2, \vec{V}_0$:	Positive-, negative-, and zero-sequence voltage respectively at the PCC;
$\vec{Z}_{s1}, \vec{Z}_{s2}, \vec{Z}_{s0}$:	Positive-, negative-, and zero-sequence source impedance.
$\vec{Z}_{f1}, \vec{Z}_{f2}, \vec{Z}_{f0}$:	Positive-, negative-, and zero-sequence feeder impedance.

$\vec{I}_1, \vec{I}_2, \vec{I}_0$:	Positive-, negative-, and zero-sequence fault current.
$\vec{E}$:	Positive sequence source voltage.
$\vec{V}_{Apre}, \vec{V}_{Bpre}, \vec{V}_{Cpre}$:	Pre-fault voltages of phase A, B, C respectively. phases and where
$\vec{V}_A, \vec{V}_B, \vec{V}_C$:	Standing voltages during sag/swell of phase A, B, C respectively The
$\theta_A, \theta_B, \theta_C$:	Phase shifts between pre-fault voltage and standing voltage during sag/swell of phase A, B, C respectively.
v_a, v_b, v_c :	Phase voltages of phases a, b, c respectively (instantaneous).
v_d, v_q :	Phase voltages of d, q axis respectively (instantaneous).
ω_x :	Rotating speed of the reference frame.
γ_0 :	Phase shift between the stator magnetic field and the axis of phase a of the stator of the three-phase machine.
ω_1 :	Synchronous speed (corresponds to the power supply frequency).
γ :	Initial angle of the phase voltage.
ω_m :	Motor mechanical speed.
$\vec{V}_s, \vec{V}_r$:	Space phasors of motor stator and rotor voltages.
$\vec{I}_s, \vec{I}_r$:	Space phasors of motor stator and rotor currents.
$\vec{\lambda}_s, \vec{\lambda}_r$:	Space phasors of motor stator and rotor flux linkages.
R_s, R_r :	Motor stator and rotor resistances.
L_s, L_r :	Motor stator and rotor leakage inductances.
L_m :	Motor magnetizing inductance.
m_a :	Amplitude modulation ratio of PWM converter.
V_{tri} :	Triangular waveform of PWM converter.
$V_{carrier}$:	Carrier wave of PWM converter.
f_c :	Switching frequency of PWM converter.
f_1 :	Desired fundamental frequency of the converter voltage output.
Generally,	
	it should be equal to the power supply frequency.
$V_{control}$:	Control signal waveform in PWM converter.

$\vec{V}_{\text{ref}}$:	The reference voltage phasor of DVR control system.
$V_{\text{ce(sat)}}$:	Saturation value of collector-emitter (on-state voltage drop of the active IGBT).
$V_{\text{ce(TO)}}$:	Static collector-emitter threshold voltage of IGBT.
r_{int} :	Internal resistance of IGBT.
I_L :	Current through the IGBT.
$t_1(i), t_2(i)$:	Switching on and off time while the IGBT conducts for the i^{th} time interval.
$t_p(i)$:	Pulse width while the IGBT conducts for the i^{th} time interval.
$t_{d(\text{on})}$:	Turn-on delay time of IGBT.
t_r :	Turn-on rise time of IGBT.
V_{GS} :	Gate-emitter voltage of IGBT.
$t_{d(\text{off})}$:	Turn-off delay time of IGBT.
t_f :	Turn-off fall time of IGBT
$W_{\text{con-T}}$:	Conducting energy loss of IGBT.
$W_{\text{on-T}}$:	Turn-on energy loss of IGBT.
$W_{\text{off-T}}$:	Turn-off energy loss of IGBT.
W_{Tloss} :	Total energy loss during IGBT conducting.
$W_{\text{con-D}}$:	Conducting energy loss of diode.
$W_{\text{off-D}}$:	Turn-off energy loss of diode.
W_{Dloss} :	Total energy loss during diode conducting.
V_F :	Forward on-state voltage (on-state voltage drop of the active diode).
V_{FO} :	Threshold voltage of the diode.
r_d :	On-state resistance of the diode.
V_D :	Diode forward voltage.
L_{st} :	Stray inductances of the converter.
$(\frac{di}{dt})_d$:	Current gradient of diode.

P_{Tloss} :	Average power loss during IGBT conducting.
P_{Dloss} :	Average power loss during diode conducting.
Z_{jc} :	Thermal impedance between the junction and case of converter.
Z_{ch} :	Thermal impedance between the case and heatsink of converter.
Z_{ha} :	Thermal impedance between the heatsink and ambient of converter.
r_{Tk} :	Steady-state thermal resistances of the IGBT.
r_{Dk} :	Steady-state thermal resistances of the diode.
τ_{Tk} :	Thermal time constants of the IGBT.
τ_{Dk} :	Thermal time constants of the diode.
C_{Tk} :	Steady-state thermal capacities of the IGBT.
C_{Dk} :	Steady-state thermal capacities of the diode.
$T_{coup/D}, T_{coup/T}$:	Thermal coupling of IGBTs and diodes with their corresponding anti-paralleled elements.

Chapter 1

Introduction

1.1 Motivation

In recent years, due to the proliferation of sensitive power electronic equipments and the rapid expansion of information technology business, power quality problems have attracted more and more attention by the utility industry. This is because poor power quality could cause the shutdown or malfunction of these equipments and lead to considerable economic loss. A safe, reliable and price-competitive power supply to these industries is therefore a prerequisite to ensure their profitable operations [1-21].

Generally the quality of power can have a direct economic impact on many industrial consumers. This is especially so following the increasing use of electronics-based device in industry; electrical equipment will become more prone to system disturbances. The electric utility is also concerned about power quality issues as well. Meeting customer expectations and maintaining customer confidence is a strong motivator. With today's move toward greater competition between utilities, it is more important than ever. The loss of a disgruntled customer to a competing power supplier can have a significant impact financially on a utility. Therefore, industry customers and electricity utilities are both now more acutely aware of minor disturbances in the power system, obvious financial impacts will be associated with power quality problems [1, 2, 6-8].

Among the many power quality problems, voltage sags are considered to be the most devastating disturbances that can affect supply quality of distribution systems. Industry surveys in recent years reveal that 80% to 90% of the customers' complaints pertaining to power supply quality are due to voltage sags [2, 3, 6, 9].

A voltage sag is defined as a decrease in voltage magnitude of between 0.1 to 0.9 p.u, and its typical duration is from 0.5 cycle to 1 min. Voltage sags are mainly caused by faults

on the transmission and distribution systems, or by the energization of heavy loads such as the starting of large motors. The magnitude and duration of voltage sag at a specific site vary according to system electrical stiffness, types of disturbances or faults, fault-clearance practice and disturbance/fault locations [2, 6, 15, 16].

Presently, there is considerable interest in finding ways to mitigate the effects of the sags. The conventional devices used to mitigate voltage sags include Uninterruptible Power Supplies, Ferro resonant transformers, Magnetic Synthesizers, Electronic Tap changer, and motor-generator sets. However they have their limitations and may, in some cases, be used economically at low power ratings [2, 7, 8].

A more recent development in this field is the use of Customer Power compensation technique. This is achieved by utilizing advanced power electronics technology to ensure a high quality of supply voltage. It is considered to be a very useful way for mitigating the effects of a supply voltage sag on sensitive loads, one example of which is the Dynamic Voltage Restorer (DVR) [22-53]. The Dynamic Voltage Restorer is originally designed to compensate for voltage disturbances on distribution system. It is a form of series compensation device. As will be shown in the following Chapters, the fundamental principle of DVR operation is that when voltage sag occurs upstream of the system, the DVR will inject a voltage component to restore the load-side voltage to its nominal level as soon as possible. DVR is considered to be a promising and effective device for sag mitigation due to its quick response and high reliability. Indeed, the focus of this thesis is on the design and applications of the series compensator.

On the other hand, voltage swell, which is another type of power quality problem, has not been so closely investigated as in the case of voltage sag. A prolonged and excessive voltage swell is also damaging to equipment and therefore requires corrective actions. Similar in definition as with a voltage sag, a voltage swell is that of a voltage level of over 1.1 p. u., and its typical duration is also from 0.5 cycle to 1 min [2, 8].

As with sags, swells are usually associated with system fault condition. In fact, voltage sag and voltage swell could occur simultaneously. One example is that during a Single Line Ground (SLG) fault, a voltage sag will appear on the faulted phase and swell could

be introduced on the unfaulted phases. Voltage sag and swell might also occur consecutively. A typical disturbance scenario could begin with a severe fault that causes a deep voltage sag, and those electrically close-by loads may be automatically disconnected from the grid by their own under-voltage protections. Upon fault clearance, this partial load rejection event, coupled with the fact that generators and/or reactive power compensators requiring some time to adjust their outputs to control network voltages, means that there could be a momentary and/or temporary over-voltage or swell in the power system [54].

Using series compensation technique and similar in principle for sag compensation, a series compensator can also be used to inject a voltage component which is in series with the load voltage so as to reduce the load voltage during a voltage swell. Details of such scheme will be described in Chapter 3. Central to the proposed technique is the adjustment of the injection voltage phase angle, with the objective of restoring the load voltage level while exercising some control on the voltage of the energy storage device within the series compensator.

Generally, when a sag/swell occurs in the power system, not only the magnitude of network voltage changed, but also a jump in the phase angle of the voltage could appear. It is well known that some kinds of loads such as induction motor or adjustable speed drives are particularly sensitive to voltage magnitude and/or phase deviations. The phase shift will introduce a certain level of dynamic transients in the loads. This could have undesirable impacts on the system. Therefore, it is very necessary to study the load response under the voltage phase shift. In this project, linear static load, induction motor and adjustable speed drives are selected as the typical loads under study. Their responses under the voltage phase shift are closely investigated in Chapter 4.

Since in most cases of DVR applications, instantaneous voltage injection is still being used because it is convenient to apply. Therefore, the load transient current introduced into the system will induce in the series compensator similar transients. As power converter forms the main component of the restorer, the increase in the current in the converter would result in increased losses in the switching devices within the converter. The switching devices could be damaged if the losses are excessive. Chapter 5 describes

a method to evaluate the transient currents, to quantify the losses and to calculate the switching device junction temperature rise within the converter. It forms the basis from which the switching devices in the DVR can be sized readily in order to alleviate the thermal problem due to the load transient currents.

Currently, most of research works on series compensator applications are concentrated on voltage sags as consequence of system faults. On the other hand, switching on of large loads such as the starting of large motors is another common cause of voltage sags. In Chapter 6, series compensation of sags due to motor-start will be studied.

1.2 Major contributions of the Thesis

As a result of the research work, the following original contributions have been made:

1. **A generalized voltage compensation strategy for mitigating the impacts of voltage sag/swells:** As can be seen in the literature, voltage sag compensation through series compensation device (such as a DVR) has been studied by many researchers. It is shown that during a voltage sag compensation process, the restoration process almost inevitably requires the injection of energy from the compensator to the external system. Unfortunately, the impacts of voltage swell have not been so closely investigated. Using series compensation technique and similar in principle for sag compensation, the DVR can also be used to reduce the load voltage during a voltage swell by in-series injecting a voltage component into the system. However, the studies to be described in latter Chapters show that under most voltage swell condition, the DVR will absorb active power from the external system. It will lead to a voltage rise in the energy storage device in the DVR. Excessive over-voltage in the DC-link will damage the normal operation of the restorer. Therefore a compensation strategy suitable for swell should also address the aforementioned problem. Based on the detailed analysis of three existing series compensation techniques, a generalized compensation scheme for mitigating both voltage sag and swell including unbalanced voltage conditions is described. Under the proposed method, the load-side voltage can be restored for

sag and swell while the voltage of the DVR energy storage device can also be controlled.

2. **Analysis of load response under voltage phase shift:** Associated with the occurrences of sag/swell is the unavoidable phase shift on power supply voltage. The proposed generalized compensation method for sag/swell will also introduce the phase shift on the load supply voltage. Appreciable phase shift could have undesirable effects on the proper functioning of loads. The effect of voltage phase shift on the loads is also examined in this investigation. Linear static loads, induction motor and adjustable speed drive are selected as the typical loads to study because they are most commonly found in power distribution systems. Results obtained so far show that both the linear static load and the induction motor load will exhibit drastically different dynamic responses under this kind of phase shift, and analytical expressions indicate that load transient current responses are dependent of the magnitude of phase shift angle as well as load parameters.

For adjustable speed drives, when a voltage disturbance occurs, the corresponding DC-link voltage will have a sudden change in magnitude. However since the voltage phase shift is instantaneous, the DC-link voltage change will also be of short-duration oscillations. Therefore, the effect of the phase shift on the output voltage of an adjustable speed drive is seen to be minimal.

The more practical condition when many parallel loads are subject to the same phase shift is also examined. It is clear that the magnitude of the impedance between the load terminals and the DVR has a great influence on the load transient current. If the DVR is installed electrically close to the loads, the load transient response would be the sum of each single load response, and expressions relating to the response have been derived.

The exponential voltage injection scheme to mitigate large transient current due to voltage phase shift is then proposed. The results show that the suggested method could reduce the magnitudes of the load transients effectively. This method

balances the need for achieving acceptable load swings while maximizing the saving in the stored energy of the series Custom Power device.

3. **Analysis of the impact of load transient current on DVR.** Based on the results of the analysis of load response under voltage phase shift, the impact of the load transient current on the DVR device has also been investigated. As the DVR is connected in series with the load, it must experience similar transient load response under the voltage phase shift. The large current transients will cause excessive heating in the DVR power electronics elements such as the IGBT and diodes. Research related to the issues of the effects of device temperature rise on the operation of the DVR or any other custom power device has not been reported in the literature thus far. In this thesis, a complete analysis on how to calculate the resulting temperature rise within the switching devices of the DVR is given. Simulation result shows that the temperature rise is not negligible. The proposed method can be adopted in the design of Custom Power device, by incorporating the thermal loading effect on the converter power module whenever a voltage phase shift occurs.

4. **Sag compensation due to motor-start.** Most of the studies concerning series compensation applications are based on the assumption that the voltage sags are due to short-circuit faults. Switching of large loads such as motor-start is another common cause of sag occurrences. The compensation of these two type sags are different since for the later, the motor starting current may typically last for a longer duration compared to that caused by faults, even though the sag magnitude could be smaller. The consequence could be damages to the DVR device, again due to the thermal loading effect on the switching devices. It may also result in the depletion of the energy in the energy storage device of the restorer. A method to estimate the motor peak starting current and the required starting energy of motor starting is given. The suggested method can be utilized by DVR designers to size the DVR, to guarantee the DVR can mitigate such kind of sag without any damage.

1.3 Organization of the Thesis

As was explained, Chapter 1 provides a brief summary of the motivation and lists the major contributions made in this project. Chapter 2 will contain a description of common power quality problems and how these could impact on loads. A literature review pertaining to the research area on series compensator, particularly that on the DVR, will also be included.

Original contributions of this research are described in details in Chapter 3- Chapter 6. In Chapter 3, the problem formulation pertaining to sag/swell compensation will be considered first. The analysis is then extended to the case of unbalanced voltage disturbances. From the analytical results obtained, a generalized compensation strategy for voltage sag/swell disturbances will be proposed. Under this new method, the load voltage can be restored for sag and swell while the DC-link voltage can also be controlled.

The analysis on load response under the voltage phase shift will be given in Chapter 4. Static linear load, induction motor and adjustable speed drive are selected as they are the most widely used loads. With the developed load models, their responses under the voltage phase shift will be analyzed. The exponential voltage injection scheme to mitigate the transient current due to voltage shift is also given in this Chapter.

Based on the results in Chapter 4, the impacts of load transient current on the series compensator is investigated in Chapter 5. A complete method on how to calculate the temperature rise within the compensator converter following the voltage shift is given.

Series voltage compensation of sags due to motor-starts is studied in Chapter 6. A method to estimate the peak current and required starting energy for the motor load starting is given. Based on the suggested method, suitable DVR components in the energy storage device could be selected.

The main findings of the research and recommendations for future works are given in Chapter 7.

Chapter 2

Power Quality

As pointed out in Chapter 1, power quality problems have attracted increasing attention of the utility industry in recent years. In this Chapter, a general review of the various issues concerned power quality will be given. Solutions for power quality enhancement based on Custom Power technology will also be presented. Since voltage sag is the most devastating power quality disturbance that can produce undesirable impacts on distribution systems, a detailed analysis of voltage sags and its impact on different kinds of loads will also be given. As the Dynamic Voltage Restorer (DVR) is one of the most promising and effective devices for the improvement of power quality at the distribution level, a detailed description about the DVR is included.

2.1 A Review of Power Quality Issues

2.1.1 What is Power Quality?

In broadest sense, the term “Power Quality” should be interpreted as service quality, encompassing the three aspects of reliability of supply, quality of power offered, and provision of information [8]. A more restrictive interpretation, widely used in the recent literature, relates to:

The ability of a power system to operate loads without disturbing or damaging them, a property mainly concerned with voltage quality at points of common coupling.

The ability of loads to operate without disturbing or reducing the efficiency of the power system, a property mainly, but not exclusively, concerned with the quality of the current waveform.

In general, Power Quality is the combination of voltage quality and current quality [8]. However, in almost all practical situations, power supply system can only control the quality of the voltage; it has no control over the currents that particular loads might draw. Therefore, the standards in the power quality area are devoted to maintaining the supply voltage within certain limits. In some cases, one can say: Power Quality is also Voltage Quality [8].

2.1.2 Growing Concern in Power Quality

Both electric utilities and industry end-users of electricity have increased concern on the quality of electric power supply since the late 1980s. There are several major reasons for the growing concern [1]-[8]:

- An increasing proportion of load equipment is becoming more sensitive to power quality variations than the equipment used in the past. Many new load devices contain microprocessor-based controls and power electronic devices that are particularly sensitive to many types of disturbances.
- The increasing emphasis on overall power system efficiency has resulted in a continued growth in the applications of devices such as high-efficiency, adjustable-speed motor drives and shunt capacitors for power factor correction. The emphasis is to reduce losses. This has resulted in higher harmonic levels within power systems and has led to concern on the future impact of the harmonics on system capacities.
- Increased awareness of power quality issues by the end users. Utility customers are becoming better informed about such issues as interruptions, sags, and switching transients and are challenging the utilities to improve the quality of power delivered.
- Utilities also want to deliver a good product. Most utilities have been committed to deliver a high quality power supply for many decades. Designing a system with a high reliability of supply, at reasonable cost, is always a technical challenge.

- Power supply has become so reliable that long interruptions are rare in most industrialized countries. This means that consumers are very sensitive on phenomena such as voltage sags and harmonic distortions. Conversely, in countries where the electricity supply has low availability, e.g., 2 hours per day, power quality does not appear to be such a big issue as in countries with availabilities of well over 99.9%.

2.1.3 Power Quality Terms and Definitions

Power quality standards vary between countries. Needless to say that poor quality power affects almost all consumers. It is therefore important to provide a clear understanding of the terms and definitions that are used in the literature before one can proceed to consider detailed analysis work. In particular, one should consider the following [2][7][8]:

Voltage sags (Dips): Voltage sag is considered the most common power quality phenomenon in practice. A voltage sag is defined as a decrease in voltage to between 0.1 and 0.9 p.u. in *rms* value at power frequency for a duration of 0.5 cycles to 1 minute. Normally, voltage sags are caused by faults on adjacent feeders or the switching of heavy loads. Since many types of load equipments are quite sensitive to voltage sags, they often could be tripped when the *rms* voltage drops below 90% for longer than one or two cycles. This could lead the economic loss to the customers.

Voltage interruptions: Voltage interruptions mean the complete loss of voltage of below 0.1p.u on one or more phase. Momentary interruptions are defined as those incidents lasting between 0.5 cycles and 3s, temporary interruptions have a time span between 3s and 60s, and sustained interruptions are those that last for a period longer than 60s.

Voltage swells: As the opposite of voltage sag, a voltage swell is defined as an increase to between 1.1 and 1.8 p.u. in *rms* value at the power frequency for durations from 0.5 cycle to 1 min. Similar to sags, swells are usually associated with system fault conditions. For example, a swell can occur on the unfaulted phases during a single-line-grounded (SLG) fault. On an ungrounded system, with infinite zero-sequence impedance, the line-to-ground voltages on the ungrounded phases will be 1.73 p.u. under a SLG condition.

Swells can also be caused by switching off a large load or energizing a large capacitor bank.

Transients: Transients is a phenomenon of supply voltage which is varying between two consecutive steady states. It can be a unidirectional impulse of either polarity, or damped oscillatory wave with the first peak occurring in either polarity. Although most transients are usually generated near the user due to the operation of other equipment, switching operations on the utility network can pose a more serious problem. For example, capacitor switching can lead to transients with a magnitude of 2~3 p.u. and the high energy levels can considerably shorten the life of surge protection devices.

Overvoltage: Similar with voltage swell, overvoltage is also defined as an increase in rms voltage at the power frequency, but the main difference between these two terms is that the later is usually used to describe a specific type of long-duration variation. Normally this duration could last more than 1 minute. The other difference is the range of voltage swell is about 1.1 to 1.9 pu while the typical values of overvoltage are 1.1 to 1.2 pu.

Undervoltage: Comparing to voltage sag, undervoltage also means a value of less than the nominal voltage despite its lasting time is generally greater than one minute. Its typical values are 0.8 to 0.9 p.u.

Harmonics: Harmonics is another common power quality issue in the utility industry. It is defined that sinusoidal voltages or currents having frequencies that are multiples of the fundamental power frequency. Generally harmonics are caused by nonlinear characteristics of power system devices and loads. For example, the switches of power electronics devices in the adjustable speed drives could inject higher harmonics components into the power system and leads to the distorted waveforms in the current. In recent years, in particular with the growth of non-linear loads such as personal computers and power electronic devices, the combined effect of many such devices can result in harmonic distortion of voltage/current greater than 5~10% levels and exceed the normal industry regulation standards.

Notches: Periodic voltage disturbances lasting less than 0.5 cycles. Mainly power electronics devices cause notching when the current is commutated from one phase to another during the momentary short circuit between the two participating phases.

Voltage fluctuations/flickers: Voltage fluctuations are systematic variations in the envelope or a series of random voltage changes with magnitudes that do not normally exceed the voltage range of 0.9 to 1.1 p.u. Typical cause of flickers includes the operations of power converters on electrical devices.

The above voltage disturbances are illustrated in Figure 2-1.

Figure 2-1 Waveform distortions associated with poor power quality [8]

2.1.4 CBEMA And ITIC Curves

The well-known Computer Business Equipment Manufacturer Association (CBEMA) curve, shown in Figure 2-2, can be used to evaluate the voltage quality of a power system with respect to voltage interruptions, sags or undervoltages, swells or overvoltages. This curve was originally produced as a guideline to help CBEMA members in the design of the power supply for their computer and electronic equipment.

The curve relates the magnitude and duration of voltage variations which can be tolerated for such equipment connected to a power system. The region between the two boundaries is the tolerance envelope within which electronic equipment is expected to operate reliably. Rather than noting a point on the plot for every measured disturbance, the plot can be divided into small regions with certain range of magnitude and duration. The number of occurrences within each small region can be recorded to provide a reasonable indication of the quality of the system.

The CBEMA Curve from FIPS No. 94.

Figure 2-2 CBEMA curve

CBEMA has been renamed “Information Technology Industry Council (ITIC)”, and the new curve shown in Figure 2-3, has been developed to replace CBEMA. However, due to the prominence of the CBEMA between the computer and electronic industries, the ITIC curve is being regarded as the new CBEMA curve within the high technology circle. The main difference between them is that the ITIC version is piecewise and hence easier to digitize than the continuous CBEMA curve. The tolerance limits at different durations are very similar in both cases.

Figure 2-3 ITIC curve

2.2 Custom Power Technologies

2.2.1 Introduction

Causes of the power quality problems may be due to various reasons such as system fault, switching of large loads, or utility switching. Even if the power supply system has been designed for maximum reliability and is of high quality, such power quality problems cannot be completely eliminated. This is because the widely distributed energy supply process is subjected to both atmospheric influences and unpredictable component failures. Only additional measures, taken in the power supply system or at the users' end, can protect critical loads from disturbances of that kind. For example, on the customer's side, Uninterruptible Power Supply (UPS) to protect small machines may be used. However, the rapid proliferation of load equipment with microprocessor-based controls and power electronic devices has resulted in the necessity to protect an entire plant, an entire feeder, or even a block of customers or loads in some cases. In these cases, customer-side solutions are not always the best or the most economical way to remedy degradation in power supply. Thus a special subset of power conditioning devices used at the voltage levels ranging from 1kV to 38kV have become one of the most active topics in the power system research arena in the past decade. This special class of the power conditioning devices is called Custom Power devices [1, 2, 4, 7, 8].

The concept of Custom Power was first introduced by N. G. Hingorani in [1]. Just like flexible ac transmission systems (FACTS) for bulk power transfer, the term Custom Power pertains to the use of power electronic controllers for distribution systems. Just as FACTS improves the reliability and quality of power transmission by simultaneously enhancing both the level of power transfer and stability, the Custom Power enhances the quality and reliability of power that is delivered to customers. Under this scheme a customer receives a pre-specified quality power. This pre-specified quality may contain a combination of specifications of the following [7, 8]:

- Frequency of rare power interruptions.
- Magnitude and duration of over and undervoltages within specified limits.
- Low harmonic distortion in the supply voltage.
- Low phase unbalance.

- Low flicker in the supply voltage.
- Frequency of the supply voltage within specified limits.

Custom Power devices, or controllers, are devices that include static switches, inverters, converters, injection transformers, master control modules, and/or energy storage modules. These devices will have the ability to perform current interruption and voltage regulation functions in a distribution system, for the purpose to improve system reliability and/or power quality.

The key technology, which has made Custom Power devices possible, is the turn-off solid-state switches. Developments in Gate Turn-Off thyristor (GTO) and Insulated Gate Bipolar Transistor (IGBT) mean that devices with operational capabilities suitable for high power applications are now available at a cost that makes them economically viable at distribution power levels. Also important to realizing this technology has been the advances made in micro-controllers, signal processors, fiber optic communications and techniques to series connect the solid-state switches [1, 8].

2.2.2 Main Custom Power Devices

Custom Power devices could be broadly divided into compensating type or network reconfiguring type. The compensating devices either compensate a load, i.e., correct its power factor, unbalance etc. or improve the quality of the supplied voltage. These devices are either connected in shunt or in series or a combination of both [8]. The devices include the following:

- **Dynamic Voltage Restorer (DVR)** [22-23] is used to protect sensitive loads from sag/swell or disturbances in the supply voltage. Generally it will be series connected to the system and act as a controlled voltage source. A more detailed description of the DVR will be given in Section 2.4.
- **Distribution Static Compensator (D-STATCOM)** [8, 55] is used to protect the distribution system from the effects of a polluting, non-linear harmonics

producing load. Generally, D-STATCOM will operate in current control mode, therefore its ideal behavior is represented by the current source. Figure 2-4 shows a schematic diagram of a distribution system compensated by an ideal shunt compensator (D-STATCOM). If the source-side current $\vec{I}_s$ flowing through the PCC bus is unbalanced or distorted, the D-STATCOM will generate a current $\vec{I}_{inj}$ such that it would keep the current $\vec{I}_s$ on a desirable value and within an acceptable level of distortion. From Figure 2-4, it shows that: $\vec{I}_s = \vec{I}_L - \vec{I}_{inj}$.

Figure 2-4 Schematic diagram of ideal load compensation by D-STATCOM

A basic configuration of D-STATCOM is shown in Figure 2-5. The main component D-STATCOM is an IGBT-based three-phase voltage source converter. The function of the converter is to act to synthesize the injection voltage/current to mitigate the current disturbances. IGBT switch is chosen because it could offer low loss switching at fairly high frequencies ($>1\text{kHz}$). It allows relatively clean waveforms to be generated using pulse width modulation (PWM) techniques. Converters using IGBT have been widespread and have been in reliable use for traction and drive applications for many years, although generally not at the same voltages or power levels as for the distribution systems considered here. A control system is designed to control the converter, which constantly monitors the distribution line and compares the data with a reference signal. The filter scheme is also necessary since it filters out the high-order harmonics produced by the converter. The shunt injection transformer is used to boost and to couple the injection current to the distribution system. Its main functions are current boosting and electrical isolation. The energy storage device is necessary to offer the required energy for the compensation. It would be made up of ultra-high energy density capacitors (e.g. capacitors using double layer technology), advanced batteries, flywheel energy storage or superconducting magnetic energy storage.

Figure 2-5 Basic Configuration of D-STATCOM

The other kind of Custom Power devices is network reconfiguration type. It is also usually called switchgear [8]. It includes current limiting, current breaking and current transferring devices. The solid-state or static versions of the devices are called:

- Solid-state current limiter (SSCL)
- Solid-state circuit breaker (SSCB)
- Solid-state transfer switch (SSTS)

Solid-state current limiter (SSCL): The schematic diagram of a solid-state current limiter is shown in Figure 2-6. It consists of a pair of opposite-poled thyristor switches in parallel with the current limiting inductor L_{lim} . Generally, the current limiter is connected in series with a feeder. In the healthy state, the opposite poled switch remains closed. When a downstream fault is detected, the switches open so that the fault current will

flows through the current limiting inductor and therefore the current would be limited. In addition a series RC combination (R_{snubber} and C_{snubber}) is connected in parallel with the opposite poled switch. This RC combination can reduce the high transient current during the switches open/close operation.

Figure 2-6 Schematic diagram of a solid-state current limiter

Solid-state circuit breaker (SSCB): A SSCB has almost the same topology as that of an SSCL except that the limiting inductor is connected in series with an opposite poled thyristor pair. Once a fault is detected, the bidirectional switch is switched off, simultaneously the thyristor pair is switched on, this will force the fault current to flow through the current limiting inductor in the same manner as discussed above. The thyristor pair is blocked after a few cycles if the fault persists. The current through the thyristor pair will cease to flow at the next available zero-crossing of the current.

Solid-state transfer switch (SSTS): The schematic diagram of a solid-state transfer switch (SSTS) is shown in Figure 2-7. This device, which is also known as a static transfer switch (STS), is used to transfer power from the preferred feeder to the alternate feeder in case of voltage sag/swell or fault in the preferred feeder. The transfer switch would be used to protect sensitive loads. An SSTS contains two pairs of opposite poled switch. Generally the switch is made of thyristors. These switches are denoted by S_{w1} and S_{w2} in Figure 2-7. Suppose the preferred feeder supplies the power to the load through the switch S_{w1} while S_{w2} remains open. If a sudden voltage sag/swell or fault develops in the preferred feeder, the SSTS then closes S_{w2} such that current starts flowing through the alternate feeder to the load. S_{w1} is then switched off. Therefore the

sensitive load could be protected from voltage disturbances or faults on the preferred feeder.

Figure 2-7 Schematic diagram of a static transfer switch

Figure 2-8 shows how these devices can be deployed on the distribution system to provide power quality improvement at the distribution feeder level for sensitive loads.

Figure 2-8 Custom Power products for distribution systems [56][45]

2.3 A Review of Voltage Sag

Since voltage sag is considered the most common power quality occurrence and could cause enormous economic loss to the electric utility customers, an explanation of the causes, characteristics of the sag, and its effect on different types of sensitive loads will be given in this section.

2.3.1 Causes of Voltage Sags

Voltage sags are mainly caused by short circuits and switching on of large loads such as the starting of large motors [2].

Sags due to faults:

To explain the origin of a voltage sag due to a short-circuit fault in power system, the voltage divider model, shown in Figure 2-9, can be used. It may appear to be a simple model, especially for transmission systems. However, as will be seen in the course of this and later chapters, it has turned out to be a rather useful model to predict some of the properties of sags.

Figure 2-9 Voltage divider model for the study of a voltage sag

Suppose that the source voltage is exactly at $1\angle 0^\circ$ p.u, thus $|\vec{E}| = 1$ p.u. Before the fault, the voltage on PCC bus is nearly equal to the source voltage since the source impedance $\vec{Z}_s$ could be neglected when it is compared to the load impedance $\vec{Z}_{Load}$. Suppose a fault occurs in one of the parallel feeders. Let $\vec{Z}_f$ represents the fault impedance. The voltage at PCC during fault could be approximately expressed as (since $|\vec{Z}_{Load}| \gg |\vec{Z}_f|$):

$$\bar{V}_{\text{sag}} \approx \frac{\bar{Z}_f}{\bar{Z}_s + \bar{Z}_f} |\bar{E}| \quad (2.3.1)$$

It is clear that the magnitude of the PCC voltage will change from $|\bar{E}|$ to $|\bar{V}_{\text{sag}}|$. From (2.3.1), one sees that the sag becomes deeper for faults electrically closer to the customer (when $|\bar{Z}_f|$ becomes smaller), and for systems with a smaller fault level (when $|\bar{Z}_s|$ becomes larger). The duration of the sag is mainly determined by the fault-clearing time.

Note that a single-phase model has been used here, whereas in reality the system is of three-phase. That means that equation (2.3.1) strictly speaking only holds for a balanced three-phase faults.

Furthermore, from the above analysis, one must realize the difference between a voltage sag and an interruption. When a fault occurs in the power system, the large fault current flows through the fault position and leads to the voltage sag on the PCC bus. Once the faulted feeder protection device operates and isolates the faulty feeder from the system, the voltage sag on PCC bus would disappear but the customers on the faulted feeder will experience an interruption. In conclusion, sag means there is still a remaining voltage on the supply and the duration will be determined by the fault-clearance time. However, an interruption is the complete disruption of supply to the loads and its recovery time will be much longer than the sag duration since the faulted feeder normally needs longer time to be repaired.

Sags due to starting of induction motor:

Another important cause of voltage sags, one which has actually been of much more concern to designers of industrial power systems in the past, is the starting of large induction motors [2, 57-62]. During start-up, an induction motor will draw a current typically five to six times higher than the motor current rated value. This current remains high until the motor reaches its nominal speed. Depending on the starting method and motor construction, typically the duration can be for between several seconds to a minute. The reduction in voltage depends very much on the system parameters. Consider the system shown in Figure 2-10. The difference between this figure and Figure 2-9 is that

the fault impedance $\bar{Z}_f$ shown early is replaced by the motor impedance $\bar{Z}_M$ during the motor start-up.

Figure 2-10 Equivalent circuit for voltage sag due to induction motor starting

During motor start-up, the PCC voltage is:

$$\bar{V}_{\text{sag}} = \frac{\bar{Z}_M}{\bar{Z}_s + \bar{Z}_M} \bar{E} \quad (2.3.2)$$

An estimation of the sag voltage during motor starting-up can be obtained in the following way. When a motor of rated $\bar{S}_{\text{motor}}$ is fed from a source with short-circuit power $\bar{S}_{\text{source}}$, the source impedance can be expressed as:

$$\bar{Z}_s = \frac{|\bar{V}_n|^2}{\bar{S}_{\text{source}}^*} \quad (2.3.4)$$

where $\bar{V}_n$ is the rated voltage of the motor and * indicates its complex conjugate. For the motor impedance during starting:

$$\bar{Z}_M = \frac{|\bar{V}_n|^2}{\beta \bar{S}_{\text{motor}}^*} \quad (2.3.5)$$

with β the ratio between the starting current and the nominal current.

Equation (2.3.2) can now be written as:

$$\bar{V}_{\text{sag}} = \frac{\bar{S}_{\text{source}}^*}{\bar{S}_{\text{source}}^* + \beta \bar{S}_{\text{motor}}^*} \bar{E} \quad (2.3.6)$$

Of course one needs to realize that this is only an approximation. The last expression can be used to estimate the sag due to the motor starting, however for an accurate assessment one needs to resort to a power system analysis package. The duration of the voltage sag depends on a number of motor parameters, of which the motor inertia is the main one.

2.3.2 Characteristics of Voltage Sags

Generally speaking, voltage sags could be divided into three-phase balanced sags and unbalanced sags[2].

Balanced Voltage Sags:

Balanced voltage sags mainly originate from three-phase faults. Equation (2.3.1) and Figure 2-9 could also be used to explain this kind of sag. As shown, that the magnitude of the sag will be determined by the source impedance $\bar{Z}_s$ and the fault impedance $\bar{Z}_f$. On the other hand, note that the short circuit in a system not only causes a reduction in the network voltage magnitude but also a jump in the phase angle of the voltage. Consider Figure 2-9, suppose $\bar{Z}_s = R_s + jX_s$ and $\bar{Z}_f = R_f + jX_f$. Substituting them into equation (2.3.1), the argument of the complex voltage sag is:

$$\theta = \arg(\bar{V}_{\text{sag}}) = \arctg\left(\frac{X_f}{R_f}\right) - \arctg\left(\frac{X_s + X_f}{R_s + R_f}\right) \quad (2.3.7)$$

In the unlikely event that $\frac{X_s}{R_s} = \frac{X_f}{R_f}$, θ is zero and there is no phase-angle jump. In most cases, the phase-angle jump will thus be present as the (X/R) ratios of $\bar{Z}_s$ and $\bar{Z}_f$ are different. Reference [2] provides a more detailed explanation of the phase jump and the numerical examples in the reference show that θ could be within the range $(-60^\circ, 20^\circ)$.

Unbalanced Voltage Sags:

The analysis of voltage sag presented in the previous Sections assume a three-phase fault on a balanced system. The voltage divider model in Figure 2.9 was used to describe the three-phase faults: the impedances used in that figure are the positive phase sequence values of the network configuration. However, most faults in power systems involve single phase or two phases [2, 15]. In such a situation, it is necessary to consider all three phases into account or use the symmetrical component theory to carry out the analysis.

For non-symmetrical faults the voltage divider in Figure 2.9 can still be used but it has to be split into its three phase sequence components: a positive-sequence network, a negative-sequence network, and a zero-sequence network. The three component networks are shown in Figure 2.11, where $\vec{V}_1$, $\vec{V}_2$ and $\vec{V}_0$ represent the positive-, negative-, and zero-sequence voltage respectively, at the PCC; $\vec{Z}_{s1}$, $\vec{Z}_{s2}$ and $\vec{Z}_{s0}$ are the respective source sequence impedance values and $\vec{Z}_{f1}$, $\vec{Z}_{f2}$ and $\vec{Z}_{f0}$ are that of the feeder. The three sequence components of the fault current are denoted by $\vec{I}_1$, $\vec{I}_2$ and $\vec{I}_0$. The positive phase sequence source voltage is denoted by the emf $\vec{E}$. It is assumed that there is no source voltage component in the negative and zero-sequence networks. The three components networks have to be connected into one equivalent circuit at the fault location. The connection of the component networks depends on the fault type.

- **Single-Phase Faults:**

For a single-phase fault, the three networks should be connected in series at the fault. The resulting circuit for a single-phase fault in phase “a”, is shown in Figure 2-11.

Figure 2-11 Equivalent circuit for a single-phase fault

Assume that $\vec{E} = 1\angle 0^\circ$, the following expressions are obtained for the sequence component voltages at the PCC:

$$\vec{V}_1 = \frac{\vec{Z}_{f1} + \vec{Z}_{s2} + \vec{Z}_{f2} + \vec{Z}_{s0} + \vec{Z}_{f0}}{(\vec{Z}_{f1} + \vec{Z}_{f2} + \vec{Z}_{f0}) + (\vec{Z}_{s1} + \vec{Z}_{s2} + \vec{Z}_{s0})} 1\angle 0^\circ \quad (2.3.8)$$

$$\vec{V}_2 = -\frac{\vec{Z}_{s2}}{(\vec{Z}_{f1} + \vec{Z}_{f2} + \vec{Z}_{f0}) + (\vec{Z}_{s1} + \vec{Z}_{s2} + \vec{Z}_{s0})} 1\angle 0^\circ \quad (2.3.9)$$

$$\vec{V}_0 = -\frac{\vec{Z}_{s0}}{(\vec{Z}_{f1} + \vec{Z}_{f2} + \vec{Z}_{f0}) + (\vec{Z}_{s1} + \vec{Z}_{s2} + \vec{Z}_{s0})} 1\angle 0^\circ \quad (2.3.10)$$

The voltages in the three phases at the PCC during the fault are obtained by transforming back from sequence domain to phase domain:

$$\begin{aligned} \vec{V}_a &= \vec{V}_1 + \vec{V}_2 + \vec{V}_0 \\ \vec{V}_b &= \alpha^2 \vec{V}_1 + \alpha \vec{V}_2 + \vec{V}_0 \\ \vec{V}_c &= \alpha \vec{V}_1 + \alpha^2 \vec{V}_2 + \vec{V}_0 \end{aligned} \quad (2.3.11)$$

where $\alpha = \frac{1}{2} + \frac{\sqrt{3}}{2}j$

From (2.3.8) through (2.3.11) calculate the voltages in each phase, as follows:

$$\begin{aligned} \vec{V}_a &= 1 - \frac{\vec{Z}_{s1} + \vec{Z}_{s2} + \vec{Z}_{s0}}{(\vec{Z}_{f1} + \vec{Z}_{f2} + \vec{Z}_{f0}) + (\vec{Z}_{s1} + \vec{Z}_{s2} + \vec{Z}_{s0})} \\ \vec{V}_b &= \alpha^2 - \frac{\alpha^2 \vec{Z}_{s1} + \alpha \vec{Z}_{s2} + \vec{Z}_{s0}}{(\vec{Z}_{f1} + \vec{Z}_{f2} + \vec{Z}_{f0}) + (\vec{Z}_{s1} + \vec{Z}_{s2} + \vec{Z}_{s0})} \\ \vec{V}_c &= \alpha - \frac{\alpha \vec{Z}_{s1} + \alpha^2 \vec{Z}_{s2} + \vec{Z}_{s0}}{(\vec{Z}_{f1} + \vec{Z}_{f2} + \vec{Z}_{f0}) + (\vec{Z}_{s1} + \vec{Z}_{s2} + \vec{Z}_{s0})} \end{aligned} \quad (2.3.12)$$

Now set $\Delta = (\vec{Z}_{f1} + \vec{Z}_{f2} + \vec{Z}_{f0}) + (\vec{Z}_{s1} + \vec{Z}_{s2} + \vec{Z}_{s0})$, then the voltage drop in the non-faulted phases consists of three terms:

- A voltage drops proportional to the positive-sequence source impedance, along the direction of the pre-fault voltage, $-\alpha^2 \frac{\vec{Z}_{s1}}{\Delta}$ or $-\alpha \frac{\vec{Z}_{s1}}{\Delta}$.
- A voltage drops proportional to the negative-sequence source impedance, along the direction of the pre-fault voltage in the other non-faulted phase, $-\alpha^2 \frac{\vec{Z}_{s2}}{\Delta}$ or $-\alpha \frac{\vec{Z}_{s2}}{\Delta}$.

- A voltage drops proportional to the zero-sequence source impedance, along the direction of the pre-fault voltage in the faulted phase, $-\frac{\bar{Z}_{s0}}{\Delta}$.

Under the assumption that the positive, negative and zero sequence impedances of the system as well as the fault have the same angle, these voltages can be shown by the phasor diagram in Figure 2-12.

Figure 2-12 Phase-to-ground voltages during a single-phase fault

• Phase-to-Phase Faults:

Suppose the fault is between the phases “b” and “c”. Thus “a” is the non-faulted phase. For a phase-to-phase fault, the positive- and negative-sequence networks are connected in parallel, as shown in Figure 2-13. The zero-sequence voltages and currents are zero for a phase-to-phase fault.

The sequence voltages at the PCC are (again with $\bar{E} = 1\angle 0^\circ$):

$$\begin{aligned}\bar{V}_1 &= 1 - \frac{\bar{Z}_{s1}}{(\bar{Z}_{f1} + \bar{Z}_{f2}) + (\bar{Z}_{s1} + \bar{Z}_{s2})} \\ \bar{V}_2 &= \frac{\bar{Z}_{s2}}{(\bar{Z}_{f1} + \bar{Z}_{f2}) + (\bar{Z}_{s1} + \bar{Z}_{s2})} \\ \bar{V}_0 &= 0\end{aligned}\tag{2.3.13}$$

Figure 2-13 Equivalent circuit for a phase-to-phase fault

The phase voltages can be found from (2.3.13) by using the transformation (2.3.11). This results in the following expressions:

$$\begin{aligned}\bar{V}_a &= 1 - \frac{\bar{Z}_{s1} - \bar{Z}_{s2}}{(\bar{Z}_{f1} + \bar{Z}_{f2}) + (\bar{Z}_{s1} + \bar{Z}_{s2})} \\ \bar{V}_b &= \alpha^2 - \frac{\alpha^2 \bar{Z}_{s1} - \alpha \bar{Z}_{s2}}{(\bar{Z}_{f1} + \bar{Z}_{f2}) + (\bar{Z}_{s1} + \bar{Z}_{s2})} \\ \bar{V}_c &= \alpha - \frac{\alpha \bar{Z}_{s1} - \alpha^2 \bar{Z}_{s2}}{(\bar{Z}_{f1} + \bar{Z}_{f2}) + (\bar{Z}_{s1} + \bar{Z}_{s2})}\end{aligned}\quad (2.3.14)$$

From (2.3.14), it is shown that the voltage drop in the non-faulted phase depends on the difference between the positive and negative-sequence source impedances $\bar{Z}_{s1}$ and $\bar{Z}_{s2}$. As these are usually equal, the voltage in the non-faulted phase will not be affected by the phase-to-phase fault. Under the assumption $\bar{Z}_{s1} = \bar{Z}_{s2}$, (2.3.14) becomes:

$$\begin{aligned}\bar{V}_a &= 1 \\ \bar{V}_b &= \alpha^2 - \frac{(\alpha^2 - \alpha) \bar{Z}_{s1}}{(\bar{Z}_{f1} + \bar{Z}_{f2}) + 2\bar{Z}_{s1}} \\ \bar{V}_c &= \alpha + \frac{(\alpha^2 - \alpha) \bar{Z}_{s1}}{(\bar{Z}_{f1} + \bar{Z}_{f2}) + 2\bar{Z}_{s1}}\end{aligned}\quad (2.3.15)$$

It is clear that the faulted phases have the same voltage changes, $\frac{(\alpha^2 - \alpha)\bar{Z}_{s1}}{(\bar{Z}_{f1} + \bar{Z}_{f2}) + 2\bar{Z}_{s1}}$, but opposite in direction. These voltages are shown in the phasor diagram Figure 2-14.

Figure 2-14 Phase-to-ground voltages during a phase-to-phase fault

- **Two-Phase-to-Ground Faults:**

Single-phase and phase-to-phase faults have been discussed in the two previous Sections. The only asymmetrical fault type remaining is the two-phase –to-ground fault, which is the most complicated one among them. Here only the equivalent circuit and the corresponding equations are given. Detailed analysis can be found from [2].

For a two-phase-to-ground fault the three sequence networks are connected in parallel, as shown in Figure 2-15. It is again possible to calculate component voltages and from these, calculate voltages in the three phases in the same way as has been done for the single-phase and phase-to-phase faults.

The sequence voltages at the PCC for a fault between phases b and c and ground are given by the following expressions:

$$\bar{V}_1 = 1 - \frac{\bar{Z}_{s1}(\bar{Z}_{f0} + \bar{Z}_{f2} + \bar{Z}_{s0} + \bar{Z}_{s2})}{D}$$

$$\bar{V}_2 = \frac{\bar{Z}_{s2}(\bar{Z}_{f0} + \bar{Z}_{s0})}{D}$$

$$\bar{V}_0 = \frac{\bar{Z}_{s0}(\bar{Z}_{f2} + \bar{Z}_{s2})}{D} \quad (2.3.16)$$

$$\text{with } D = (\bar{Z}_{f0} + \bar{Z}_{s0})(\bar{Z}_{f1} + \bar{Z}_{f2} + \bar{Z}_{s1} + \bar{Z}_{s2}) + (\bar{Z}_{f1} + \bar{Z}_{s1})(\bar{Z}_{f2} + \bar{Z}_{s2}) \quad (2.3.17)$$

From (2.3.16) it is possible to calculate the phase voltages in the three phases:

$$\begin{aligned} \bar{V}_a &= 1 + \frac{(\bar{Z}_{s2} - \bar{Z}_{s1})(\bar{Z}_{f0} + \bar{Z}_{s0})}{D} + \frac{(\bar{Z}_{s0} - \bar{Z}_{s1})(\bar{Z}_{f2} + \bar{Z}_{s2})}{D} \\ \bar{V}_b &= \alpha^2 + \frac{(\alpha \bar{Z}_{s1} - \alpha^2 \bar{Z}_{s2})(\bar{Z}_{f0} + \bar{Z}_{s0})}{D} + \frac{(\bar{Z}_{s0} - \alpha^2 \bar{Z}_{s1})(\bar{Z}_{f2} + \bar{Z}_{s2})}{D} \\ \bar{V}_c &= \alpha + \frac{(\alpha^2 \bar{Z}_{s2} - \alpha \bar{Z}_{s1})(\bar{Z}_{f0} + \bar{Z}_{s0})}{D} + \frac{(\bar{Z}_{s0} - \alpha \bar{Z}_{s1})(\bar{Z}_{f2} + \bar{Z}_{s2})}{D} \end{aligned} \quad (2.3.18)$$

Figure 2-15 Equivalent circuit for a two-phase-to-ground fault

The voltage sags due to the various types of faults have been discussed in the previous Sections. For each type of fault, expressions have been derived for the voltages at the PCC. It is quite clear that during fault, the voltage at the point of the short-circuit is dependent on the fault type and system conditions. Note that when a voltage sag occurs, it will not only cause a change in voltage magnitude but also a change in the phase angle of the voltages. The phase shift is again dependent of the power system and fault conditions. Mitigation of transients related to the phase-angle shift is therefore another aspect which need to be considered.

2.3.3 Equipment Behavior Under Voltage Sags

In this Section, a brief description on the impact of voltage sags on electrical equipment will be given. For further reading, interested readers may wish to refer to [63-69]. Three types of equipment will be discussed since they are perceived to be most sensitive to voltage sags. The equipment are:

- Computers, consumer electronics, and process-control equipment that will often be powered through a single-phase diode rectifiers.
- Adjustable-speed ac drives which are normally fed through three-phase rectifiers.
- Adjustable-speed dc drives which are fed through three-phase controlled rectifiers.

Computers and consumer electronics equipment [2]:

The most sensitive component in a computer to sags is its power supply system. Generally, the power supply consists of a diode rectifier along with an electronic voltage regulator (dc/dc converter). A simplified configuration of the power supply to a computer is shown in Figure 2-16. A capacitor is connected to the dc-bus side of the rectifier, which would reduce the voltage ripples at the input of the voltage regulator. The voltage regulator could supply a more stable dc voltage to the downstream electronic part of the computer by regulating the input un-adjusted dc voltage. When ac supply voltage reduces (sags), the dc-side voltage of the rectifier also decreases. Since the voltage regulator could only keep its output voltage constant over a certain range of input voltage, if the input voltage is too low and outside its control range, the voltage regulator will create a low voltage. The control system of the computer will detect this undervoltage and will automatically send a signal to shut down the computer. This may lead to the undesirable loss of data or interrupt of the flow or work of the computer users.

Similarly, process control equipments are widely used in industry. Their power supply also has similar structure to the desktop computers. Thus, they are also extremely sensitive to voltage sags; equipment are reported to trip when the voltage drops below 80% for a few cycles. The consequences of the tripping of process control equipment can be enormous. A desktop computer trip might only lead to the loss of 1 hour of work

(typically less), but the tripping of a process-control computer can easily lead to shut-down of the whole production line. A restarting time of more than several hours is also very common. It is clear that the former is merely inconvenience, whereas the latter should be avoided at any cost.

Other low-power electronic devices also have the similar power supply as the computers and the process control equipment. Consequently, they are also quite sensitive to sags. The difference is in the different consequences of the sag-induced trip.

Figure 2-16 A typical computer power supply

Adjustable-speed AC drives [63-66]:

The configuration of most ac drives is as shown in Figure 2-17. The three ac voltages are fed to a three-phase diode rectifier. The output voltage of the rectifier is smoothed by means of a capacitor connected to the dc bus. The inductance L presents in some drives aims at smoothing the dc link current and so reducing the harmonic distortion in the current taken from the supply. The dc voltage is inverted to an ac voltage of variable frequency and magnitude. This is the so-called voltage-source converter (VSC). The most commonly used method to achieve the convention is through pulse-width modulation (PWM). The motor speed is controlled through the magnitude and frequency of the output voltage of the VSC. For ac motors, the rotational speed is mainly determined by the frequency of the stator voltages. Thus, by changing the frequency an easy method of speed control is obtained.

Figure 2-17 Typical ac drive configuration

Adjustable-speed drives are equally sensitive to voltage sags just like process control equipment discussed in the previous section. Tripping of adjustable-speed drives can occur due to one or more of following reasons:

- The drive controller or protection will detect the sudden change in operating conditions and trip the drive to prevent damage to the power electronic components.
- The drop in dc bus voltage that results from the sag will cause maloperation or tripping of the drive controller or of the PWM inverter.
- The increased ac currents during the sag or the post-sag overcurrents charging the dc capacitor will cause an overcurrent trip or blowing of fuses which protect the power electronics components.
- The process driven by the motor will not be able to tolerate the drop in speed or the torque variations due to the sag.

After a trip some drives restart immediately when the voltage recovers; some restart after a certain delay and others only after a manual restart. The various automatic restart options are only relevant when the process tolerates a certain level of speed and torque variations.

Adjustable-speed dc drives [67]:

DC motors are used extensively in industrial variable-speed drive applications since they can provide a high starting torque and offer easy speed control over a wide range.

Figure 2-18 DC motor drive with separately excited armature and field winding

A typical configuration of a dc motor drive is as shown in Figure 2-18. The armature winding, which uses most of the power, is fed via a three-phase controlled rectifier. The armature voltage is controlled through the control of the firing angle of the thyristors. The longer the delay in the firing angle, the lower is the armature voltage. There is normally no capacitor connected to the dc bus. The torque produced by the dc motor is determined by the armature current, which shows almost no ripple due to the large inductance of the armature winding. The field winding is normally powered from one of the phase-to-phase voltages of the supply, and it draws only a small amount of power. Thus a single-phase rectifier is sufficient. To limit the field current, a resistance is placed in series with the field winding. A capacitor is used to limit the voltage (and torque) ripple.

DC motor drives are particularly susceptible to sag since they normally have no extra energy storage other than the motors' own inertia. Whenever a voltage sag occurs, the dc motor will slow down due to undervoltage or unbalance, while may cause the tripping of the drive. Furthermore, the field winding of most dc motor drives is supplied from an uncontrolled diode rectifier bridge. During a sag, the voltage applied to the field winding will also collapse, thus weakening the field. The time constant of the field winding is rather large compared to the sag duration, and when power is restored, the dc motor may experience an overcurrent in the armature circuit. This is accompanied by an overshoot in the motor speed during re-acceleration. Furthermore, phase-angle jumps following sags could affect the angle at which the thyristors are triggered. This could cause commutation

failures in the controlled rectifier. Since the firing angle of thyristors determine the dc output voltage, this could also affect the normal operation of dc motor drives.

2.3.4 Mitigation Methods

In the previous sections, voltage sags have been discussed in terms of their origin, characteristics, and their effects on equipment. In this section, existing and future possible ways of mitigating voltage sags will be described.

When a short-circuit fault occurs in the system, it will create a large fault current in the system. The effect of this short-circuit current at other nodes in the system is a voltage sag. The duration of sag will be determined by the fault-clearing time of the protection device. If the resulting voltage sag exceeds a certain severity, as discussed in section 2.1.4, it will cause an equipment trip. Admittedly, while not only short circuits lead to equipment trips, events such as capacitor switching or motor starting can also induce the sags. Nevertheless, the large majority of equipment trips will be due to short circuit faults. Therefore, based on the above analysis, the following five methods could be adopted as sag mitigation methods:

- Reducing the number of short circuit faults.
- Reducing the fault-clearing time.
- Changing the system such that short circuit faults result in less severe events at the equipment terminals or at the customer interface.
- Connecting mitigation equipment between the sensitive equipment and the supply.
- Improving the immunity of the equipment.

Among these mitigation methods, the mitigation equipment at the system-equipment interface will be discussed in details later. The other four mitigation methods will be discussed in this section.

Reducing the number of short circuit faults:

Reducing the number of short-circuit faults in a system not only reduces the sag frequency but also the frequency of sustained interruptions. Thus this is a very effective way of improving the quality of supply. Many customers suggest this as the obvious

solutions when a voltage sag or short interruption problem occurs. Unfortunately, the solution is rarely that simple to implement. A short circuit not only leads to a voltage sag or interruption of supply to customers but may also cause damage to utility equipment/plant. Therefore most utilities will already have reduced the fault frequency as far as economically feasible. In individual cases there could still be room for improvement, e.g., when the majority of trips is due to faults on one or two distribution lines. In which case, the following fault mitigation techniques are possible: replace overhead lines by underground cable, use covered wires for overhead line, install additional shielding wires, increase the insulation level. Operational measures such as to implement a strict policy of tree trimming, increase maintenance and inspection frequencies may also help to reduce the number of fault incidents.

Reducing the fault-clearing time

Faster fault clearing can significantly limit the sag duration. Utilities could consider the opportunity to install current limiting fuses or modern static circuit breakers. They are able to clear the fault well within half a cycle at the power frequency, thus preventing the voltage sag from lasting too long. Sometimes the situation is improved if there is enough energy stored in the load system to ride through the sag. Note that reducing the fault-clearing time does not reduce the number of sags but only their severity. It does not do anything to reduce the number or duration of interruptions.

Changing the system:

By implementing changes in the supply system, the severity of the event can be reduced. Here the costs can become very high, especially at the transmission and sub-transmission voltage levels. The main mitigation method against interruptions is the installation of redundant components. Some examples of mitigation methods especially directed toward voltage sags include installing a generator near the sensitive load. The generator will attempt to maintain the voltage during a sag if it is due to a remote fault. One other way is to use split bus arrangement in the supply path to limit the number of feeders in the exposed area. Install current-limiting coils at strategic places in the system to increase “electrical distance” to the fault is another possibility, although one should realize that

this can make the sag worse for other customers. Finally, one can connect the bus supplying sensitive equipment from two or more substations. A voltage sag in one substation will be mitigated by the infeed from the other substations.

Improving the immunity of the equipment:

Improvement of equipment immunity is probably the most effective solution against equipment trips due to voltage sags. Unfortunately, a customer often finds out about equipment immunity after the equipment has been installed. It may be very difficult for a customer to find out about the level of immunity of equipments to sags, as the customer has not had direct contact with manufacturer. Most adjustable-speed drives have become off-the-shelf type where the customer has little influence on the specifications. Only large industrial equipment is custom-made for a certain application, which enables the incorporation of voltage-tolerance requirements.

2.3.5 Mitigation Equipment

The most commonly applied method of mitigation is the installation of additional equipment at the system-equipment interface [2]. Recent developments point toward a continued interest in this way of mitigation. The popularity of mitigation equipment is because it is the only place where the customer has control over the situation. Under such a scheme, most of the mitigation techniques are based on the injection of active power to compensate for the loss of active power supplied by the system.

Dynamic Voltage Restorer – (DVR):

DVR is considered to be a very effective Customer Power Device to protect a critical load from supply voltage disturbances. The schematic diagram of a sensitive load protected by an ideal series compensator (DVR) is shown in Figure 2-19. In this diagram, the function of the DVR can be explained readily whereby the DVR is represented by an ideal voltage source that injects a voltage $\vec{V}_{inj}$ into the downstream system. When the point-of-common coupling bus (PCC) voltage ($\vec{V}_s$) is under voltage disturbance

condition, then the load-side voltage $\vec{V}_L$ could be restored to its nominal level through the following equation.

$$\vec{V}_L = \vec{V}_s + \vec{V}_{inj}$$

Since the DVR is the type of series compensator that is the main concern of this thesis, a more detailed description of the DVR will be given in the next Section.

Figure 2-19 Schematic diagram of a sensitive load protected by a DVR

Backup Power Source- SMES, BESS [70][68]:

One of the main disadvantages of a series compensator device is that it can not operate during an interruption. A shunt compensator could operate during an interruption, but its storage requirements are much higher. This leads to the proposal to use the shunt-connected backup power source, as shown in Figure 2-20. The configuration is very similar to the shunt compensator. The difference is the presence of the static switch between the system and the load bus. The moment the system voltage drops below a set rms value, the static switch opens and the load is supplied from the energy storage reservoir through the voltage-source converter (VSC). Various forms of energy storage have been proposed. A so-called Superconducting Magnetic Energy Storage (SMES) stores electrical energy in a superconducting coil. Alternatively, a Battery Energy Storage System (BESS) uses a large battery bank to store the energy. Notice that for small devices the energy storage is not a problem, but using a SMES, BESS, or any other way of storage at medium voltage will put severe strains on the storage. A backup power is only feasible if it can ride through a considerable fraction of short interruptions.

Figure 2-20 Shunt-connected backup power source

All backup power sources suggested in the literatures use shunt connection, but it is also feasible to use a series connection, as shown in Figure 2-21. This device could operate as a series controller for sags and as a backup power source for interruptions. The moment a deep sag is detected, Static Switch 1 opens and Static Switch 2 closes.

Figure 2-21 Series-connected backup power source

Cascade-Connected Voltage Controllers-UPS [71-74]:

The Uninterruptible Power Supply (UPS) is neither a shunt nor a series device, but what could be described as a cascade connected controller. The basic configuration of a typical UPS is shown in Figure 2-22. Its operation is somewhat similar to the converter part of an ac adjustable-speed drive. Compare this figure with Figure 2-17. A diode rectifier is connected back-to-back with an inverter. The main difference is the energy storage connected to the dc bus of a UPS. In current commercially available UPS, the energy storage is in the form of a battery block. Other forms of energy storage might also be suitable in the future.

Figure 2-22 Typical configuration of an uninterruptible power supply (UPS)

During normal operation, the UPS takes its power from the supply. The rectifier converts the ac voltage to dc and inverts it again to ac with the same frequency and rms value. The design of the UPS is such that the dc voltage during normal operation is slightly above the battery voltage so that that battery block remains in standby mode. All power comes from the source. The only purpose of the battery block in normal operation is to keep the dc bus voltage constant. The load is powered through the inverter that generates a sinusoidal voltage, typically using a PWM switching pattern. To prevent load interruptions due to inverter failure, a static transfer switch is used. In case the inverter output drops below a certain threshold, the static switch is opened then the load is switched back to the upstream supply.

During a voltage sag or interruption the battery block maintains the voltage at the dc bus for several minutes or even hours, depending on the battery size. The load will thus tolerate any voltage sag or short interruption without problem. For long interruptions, the UPS enables a controlled shutdown, or the starting of a backup generator.

Other solutions:

Some mitigation equipments which are not based on the voltage-source converter are discussed below.

- ***Motor-Generator Sets [75][77]***

A motor-generator set is an old solution for mitigating voltage sags. It consists of an induction motor and a synchronous generator. They are connected to a common axis together with a large flywheel. The motor would convert the incoming electrical energy to mechanical energy, and the mechanical energy would keep the generator rotate to offer

a supply to load. The flywheel is used as the energy storage device. When the source supply suffers a sag or interruption, the flywheel could maintain the synchronous generator to continue to rotate and thus protect the sensitive load. These kinds of systems are still in use (and new ones are still being installed) in industrial installations.

Figure 2-23 Principle of motor-generator set

- **Ferroresonant transformers**

Many voltage sag problems can be handled by ferroresonant transformers, which are also called constant voltage transformers (CVT). A typical ferroresonant transformer is shown in Figure 2-24. In this figure, PW, SW, NW and CW represent the primary winding, secondary winding, neutralizing winding and compensating winding respectively. Details can be found in [76].

Figure 2-24 Basic principle of the construction of a ferroresonant transformer

Ferroresonant transformers are basically 1:1 transformers. Since they are excited high on their saturation curves, this device could provide output voltage that is not significantly affected by input voltage variations. CVT are especially attractive for constant, low power loads. Variable loads, especially those with high inrush currents, present more of a problem for CVTs because of the tuned circuit on the output. In this case, the transformers should be sized at least two times larger than the load rating [78].

2.4 Dynamic Voltage Restorer

As the main focus of this thesis is on series compensation, a more detailed description of the technology will now be given. The Dynamic Voltage Restorer (DVR) is one of the most common types of series compensator [22-53]. Hence, attention will be directed toward this device.

2.4.1 Main Functional Components

A possible configuration of a three-phase DVR is shown in Figure 2-25. It consists of the following parts.

• *Voltage Source Converter (VSC)*

The PWM-controlled VSC is the most important component of DVR, which is used to synthesize the injection voltage by employing solid-state power-electronic switches (such as IGBT) in a pulse-width modulated (PWM) inverter. Generally in the DVR, three single-phase full-leg VSC are used, which consists of four pairs of IGBT and diode. Thus the phase and magnitude of injection voltage of each phase can be separately controlled. During the normal steady-state operation of the power system, the relevant pairs of the VSC are controlled to short-circuit the secondary side of the injection transformer. Hence the injected voltage $\vec{V}_{inj}$ would be zero.

• *DC-Link/Energy Storage Device*

The dc terminal of the VSC will be connected to an energy source or an energy storage device of appropriate capacity. The storage device would supply the necessary energy to the VSC and allows the generation of injected voltages. For a given voltage sag depth and as will be discussed in latter Chapters, the ride-through capability of the DVR is determined by the energy storage capacity provided. For most DVR applications, the energy storage device can be an electrolytic capacitor bank.

• Charging Circuitry

A prolonged and deep voltage sag might exhaust the energy stored in the energy storage device. Hence the charging circuitry is necessary to recover the energy to its nominal level. The main component of charging circuitry can be an uncontrolled rectifier that is connected to the ac power supply. Some additional electronic-control circuit is also used to limit the charging current and the voltage to prevent over-voltage on the dc-link. Presently, most of the charging circuits in the DVR are only active after the sag is over. There are some designs which try to charge the energy storage device during sag period by utilizing the remaining supply, but which needs more complicated control on the charging circuit and the DC-link voltage.

• Injection transformer

The purpose of including an injection transformer is to boost and couple the injection voltage generated by VSC into the feeder circuit. The winding ratio of the booster transformer has to be carefully selected to enable the VSC to compensate for the deepest voltage dips at the minimum DC link voltage. During the normal operation of the power system, the relevant switches of the VSC will be controlled in such a way that the secondary-side of the injection transformer will be short-circuited.

• Harmonic Filters

The output of PWM-controlled VSC will include a large amount of high-order harmonics due to the high switching frequency of the VSC. Therefore harmonics-filtering system is necessary to offer a clean injected voltage on the primary-side of the injection transformer. Generally, the filter system will consist of L-C section, the values of the filter components will depend on the switching frequency of VSC. The filtering system can be placed either on the primary side (line-side filter system) or on the secondary side (inverter-side filtering system), as illustrated in Figure 2-26 [26-27].

Figure 2-25 Structure of the dynamic voltage restorer

• By-pass Circuit Breaker & Isolators

The by-pass circuit breaker, installed on the primary-side of the injection transformer serves to offer continuous power supply to loads during the maintenance of the DVR. They can also be used to protect DVR from abnormally high downstream load or fault current. Isolators ISO1 and ISO2 on the incoming and outgoing sides of the DVR serve to perform isolation function.

• Control & Protection system

Although not shown in Figure2-25, a control and protection system is needed for the proper operation of DVR. Once a voltage sag is detected, the control system of DVR will decide how to inject a suitable voltage into the system. It requires the DVR to have a

quick response and guarantee a high quality waveform. Generally, a microcontroller or DSP will undertake the implementation of the designed control algorithm of injected voltage. Additionally, some protection features such as IGBT error signal will also be monitored and the control system will also response quickly on these warning signals to protect the restorer. Sometimes a PC is also needed: the PC will provide the users with an interface to monitor and modify the control and protection code to be executed by the digital controller.

2.4.2 Restoration Schemes

As explained early, when supply voltage disturbance occurs, not only the magnitude of supply voltage changes, a phase shift may also occur on the remaining supply voltage. Therefore, the design of restoration schemes of DVR becomes an interesting and challenging technical issue in the control of voltage injection. At present, there are three known strategies of restoration to compensate for voltage sag.

- ***Pre-sag compensation [30-31]:***

The upstream source-side voltage is continuously tracked and the load voltage is compensated match the wave form of that at the pre-sag condition. The method gives a nearly undisturbed load voltage, but the drawback is the control system of DVR must have a phase-locked-loop arrangement to trace and lock the phase angle of the pre-sag voltage. And in some sag cases, the pre-sag compensation method could lead to the DC-link voltage rise due to absorption of active power from the external system. This aspect of the problem will be examined in details in Chapter 3.

- ***In-phase compensation [31, 33]:***

The generated DVR voltage is always in phase with the measured supply voltage, regardless of the load current and the pre-sag voltage. This is the most simple restoration scheme. It is especially suitable for single-phase DVR, since it would ignore the phase shift of the supply voltage. For three-phase system, the situation is quite different. This is because three-phase sags could be unbalanced, particularly when the sags have different

phase shift on each phase. If the in-phase compensation is adopted, the three-phase compensated voltages are likely to remain unbalanced unless the phase shift has been taken into account. In which, the compensation method is no longer valid.

- **Energy-saving compensation [30]:**

To fully utilize the capability of the energy storage, this method uses information about the load current to minimize the depletion of the stored energy while maintaining the load voltage magnitude constant. This is the main advantage of using the energy-saving compensation method. The method is very attractive since it increases the ride-through ability of the DVR. Accordingly and with the assumption of constant load power factor, the mechanism of the energy-saving voltage restoration scheme can be explained using Figure 2-26. In this instance, the DVR is approximated as an ideal voltage source, connected in series with the load. The non-linear characteristics of the inverter are ignored for the purpose of the discussion.

Figure 2-26 Phasor Diagram under energy-saving voltage restoration scheme

In the above figure, the load current $\vec{I}$ phasor is used as the reference. $\vec{V}_{pre}$ and $\vec{I}_{pre}$ represent the pre-sag load-side voltage and current, ϕ is the load power factor angle. As

will become clearer shortly, θ is the phase angle jump in $\vec{V}_{\text{pre}}$ due to the occurrence of the sag. ϕ is the phase angle of the sustained source-side voltage ($\vec{V}_{\text{sag}}$) with respect to the load current $\vec{I}$ during the sag.

An essential requirement for voltage restoration is to ensure that the magnitude of the load terminal voltage is maintained. In fact, if the in-phase compensation method is used, the injection voltage ($\vec{V}_{\text{inj}}$) will be in phase with $\vec{V}_{\text{sag}}$ and the voltage magnitude remains unchanged from its pre-sag value. The active power injection from the DVR for in-phase compensation method will be $V_{\text{ip1}}I$. On the other hand, the energy-saving method requires a phase adjustment of $\vec{V}_{\text{inj}}$ such that the magnitude of the restored load-side voltage $\vec{V}$ equals that of its pre-sag value. $\vec{V}_{\text{inj}}$ will be at a phase angle α with respect to $\vec{V}_{\text{sag}}$, as shown in Figure 2-26. The active power injection for the energy-saving method is only $V_{\text{ip2}}I$. Obviously, by changing the angle α between $\vec{V}_{\text{inj}}$ and $\vec{V}_{\text{sag}}$, the real power during the injection has been greatly reduced. More detailed explanation of the scheme is contained in [25].

Notice that the first two restoration schemes are only suitable if the capacity of the energy storage device of the DVR is sufficiently large and no limit is placed on the injection voltage magnitude. On the other hand, the last scheme would necessitate the injected voltage to effect a phase-angle adjustment in the load-side voltage so as to maximize the use of the stored energy in the DVR. While this is a very desirable outcome, the energy-saving compensation scheme will however cause a voltage phase angle shift at the load terminals. It is well known that loads such as adjustable speed drives or induction motors are particularly sensitive to deviation in either voltage magnitude and/or phase-angle shift/jump. Thus Chapter 4 will focus on the dynamic interactions between the DRV and loads caused by the voltage phase shift.

2.5 Conclusions

An overview on the various types of power quality disturbances is given in this Chapter. The concept of Custom Power to mitigate power quality disturbances is also introduced. Since voltage sag is the most common voltage disturbance in practice, a detailed analysis concerning the origin, and characteristic of sags are also presented. Among the many possible methods to mitigate the problem, the Dynamic Voltage Restorer (DVR) is one series Custom Power device used to reduce the impacts of source-side voltage sags. Details of DVR structure and functions are also described in this chapter.

The application of the DVR is very limited since there are still a number of technical details to be studied on the design, and operational experience to be collected. The following Chapters will attempt to address the issues of the compensation strategy, and the possible interactions between the restorers with different kinds of loads.

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Chapter 3

A Generalized Voltage Compensation Strategy for Mitigating the Impacts of Voltage Sags/Swells

3.1 Introduction

As pointed out in Chapter 2, power quality problems have received increasing attention in recent years because of the significant economic impacts they can impinge on customers [2]. One method to enhance supply quality is based on the use of Series Custom Power compensation technique, an example of which is the Dynamic Voltage Restorer (DVR). As was already introduced in the previous Chapter, this device is shown to be particularly suitable for mitigating the undesirable impacts of the most common power quality disturbances known as voltage sags [2]. As shown in Section 2.3.5, essentially a DVR functions by injecting a voltage component in series with the load voltage so that the load-side voltage is maintained at its nominal level during a voltage disturbance. As will be shown clearly later, the load demand can be met fully or partially by the adjustments of power flow from an energy storage facility within DVR over the disturbance interval.

The Dynamic Voltage Restorer was first patented by Laszio Gyugyi et al [22] in 1995. Reference [22] provides a complete introduction of the components and compensation method of the restorer. It also describes a method on how to optimize the real power flow from the DVR to the utility system. This is realized by adjusting the insertion voltage based on a corrective error signal between the utility supply voltage and nominal ideal voltage signal.

In [5], a comprehensive analysis regarding the application of the DVR was given. It pointed out that the voltage sag definition based on the *rms* representation masks the information of the phase-shift and waveshape. When the supply-side voltage sag occurs, not only the magnitude in the voltage changed, phase shift might also be introduced. As the DVR is a series-connected compensation device, the phase-shift accompanying the

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voltage sag should be considered when sizing the DVR. Furthermore, the authors also pointed out that phase shift may have important impacts on some types of the load. On the other hand, regarding the function of the DVR, it was concluded that it is difficult for the restorer to mitigate the fast transients due to the finite response time of the DVR. The problems regarding the possible inrush current upon the recovery of the voltage sag are also analyzed in the paper. According to the authors, if the DVR cannot fully correct the voltage sag, an inrush current may occur upon the recovery which re-magnetizes transformer windings, accelerate motors and recharge capacitors. Therefore measures must be taken to remove the DVR from the system under this circumstance.

In [30], a complete description on energy saving injection method has given. It indicated that the energy saving method could be realized by injecting a voltage with phase advance with respect to the sustained source-side voltage. Therefore the reactive power could be used to help voltage restoration, and the consumption of real power from the perspective of energy storage device can be reduced. On the other hand, it also stated that the energy-saving injection method comes at the expense of increased voltage injection magnitude, load power swing, phase shift, and discontinuity of voltage wave-shape. For this reason, other compensation methods are also suggested in terms of satisfying Custom Power while taking into consideration the capacity of the energy storage device and the voltage injection constraint of the DVR.

In [23], the prototype of the first commercial Dynamic Voltage Restorer was introduced. It was built by Westinghouse for EPRI in 1996 and installed on the Duke Power Company 12.47-KV system to protect an automated yarn manufacturing and weaving factory. Since this paper is the first one to present the practical experience of DVR installation in the existing power system, it attracts great attention and later has been adopted as reference by many researchers. In this paper, the possibility of voltage restoration with optimized energy injection was also described but it did not give details.

In 1998, ABB also introduced their DVR in [36]. The main difference in the hardware between ABB and Westinghouse designs is the newly developed Integrated Gate Commutated Thyristors (IGCT) were selected for the converter used in the ABB installation. It was believed that the IGCT could provide the DVR with higher converter

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reliability, efficiency and inherent safety than the IGBT described in [23]. However, the optimization of energy for sag compensation was not mentioned in this scheme. The restoration scheme injects three single-phase AC voltages in series with the supply voltages to compensate for the supply-side voltage sag.

Presently, most studies on DVR compensation are focused on applying the technique in mitigating the effects of voltage sags. The main reason is because during a voltage sag, there is a decrease in power delivered to loads. Through the actions of protective devices, some of these loads are often automatically disconnected from the supply system. The disturbance in supply results in loss of production to customers. Hence there is considerable interest in finding ways to mitigate the effects of the sags.

In fact, voltage perturbation in the form of sag and swell both exists in power systems. Unfortunately, the impacts of voltage swell have not been so closely investigated. A prolonged and excessive voltage swell can be damaging to equipment. Power transformers in the load area could enter into magnetic flux saturation region during the swell and subject them to high electromagnetic stress. Often voltage sag and swell may occur consecutively in practice. For instance, the voltage recordings included in [54] show just such a system-wide voltage sag/swell occurrence. The recordings are pertaining to the Aug 2003 black-out of the Northeastern part of the US electricity network. A typical disturbance scenario could begin with a severe fault that causes deep voltage sag, and as was mentioned earlier, the automatic disconnection of load from the grid results in a partial load rejection. Upon fault clearance, however, generators and/or reactive power compensators in the power system require some time to adjust their outputs to control network voltages. The imbalance in reactive power will result in a momentary and/or temporary over-voltage or swell in the network. Hence it is most apparent that a DVR voltage injection strategy should be able to cater for both sag and swell.

Similar in principle for sag compensation, the DVR can also be used to restore load voltages during a swell. However, it will be shown in this Chapter that under most voltage swell condition, the DVR will absorb active power from the external system. This will then lead to a voltage rise in the energy storage device within the DVR. In some of the DVR currently available on the market, large capacitors are used as a medium for

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energy storage [2, 5, 6]. Excessive and prolonged over-voltage is undesirable, as this can damage the storage device. Therefore a compensation strategy suitable for dealing with voltage swell should also address the aforementioned problem.

This Chapter begins with a detailed analysis of three existing series compensation techniques described in Section 2.4.2. Based on the outcome of the analysis, a generalized compensation scheme for mitigating both voltage sag and swell is described. Central to the proposed technique is the adjustment of the injection voltage phase, with the objective of restoring the load voltage level while exercising some control on the voltage of the energy storage device within the SC. Numerical examples are used to illustrate the effectiveness of the proposed technique.

3.2 Comparison of Sag/Swell Compensation Methods

As stated earlier, one effective technique for restoring voltage sag is through the injection of a voltage component (and therefore energy) upstream of a protected load. Stripped of the peripheral equipment, Figure 3-1 shows the essential components of a series voltage compensator. In this figure, $\vec{V}_s$ and $\vec{V}$ represent the voltages on the source-side and load-side of the DVR respectively. These quantities are readily measurable.

With the DVR installed between the protected load and its source bus, the injection voltage $\vec{V}_{inj}$ from the DVR can be made to counteract a voltage sag or swell appearing in $\vec{V}_s$, to ensure a high quality load voltage $\vec{V}$. For the convenience of analysis, the non-linear characteristics of the DVR inverter are ignored. Also during the compensation process, the load power factor is assumed to be constant and lagging.

Within the DVR, $\vec{V}_{inj}$ is generated by the voltage-source converter (VSC) which is typically based on an SPWM scheme. This has been described in Section 2.4. The injected energy is obtained through the energy storage device connected to the VSI. In Figure 3-1, the energy storage system is shown as a capacitor with its terminal voltage V_d .

Figure 3-1 A typical Dynamic Voltage Restorer application scheme

The performance of such a scheme depends very much on the control of $\vec{V}_{inj}$. As described in Section 2.4, presently there are three known restoration techniques for sag compensation, denoted herewith as in-phase compensation, pre-fault compensation and energy-saving compensation. While detailed description of the three compensation techniques is included in [22, 30-33, 36], the most significant difference in these schemes is the selection of reference voltage for the control system of the DVR during the restoration process.

In-phase compensation requires the compensated load voltage $\vec{V}$ to be always in phase with the measured supply voltage $\vec{V}_s$ regardless of the pre-disturbance voltage. Therefore the reference voltage of the control system will always be in phase with $\vec{V}_s$. On the other hand, pre-fault compensation requires the continuous tracking of $\vec{V}_s$. Through the injection of $\vec{V}_{inj}$, $\vec{V}$ is made equal to that of its pre-fault value. The technique uses the pre-fault voltage of $\vec{V}_s$ as the reference signal for its control system. Finally, the energy-saving compensation method necessitates the introduction of $\vec{V}_{inj}$ to effect a phase angle adjustment in $\vec{V}$, while ensuring that the magnitude of $\vec{V}$ equals to that of its pre-fault value. Thus $\vec{V}$ will have a phase shift with respect to its pre-fault value. By changing this phase shift, the injected active power through the series compensator can be controlled so as to maximize the load ride-through capability afforded by the energy storage device.

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Figure 3-2 is a phasor diagram describing the electrical condition of Figure 3-1 where it is clear that $\vec{V} = \vec{V}_s + \vec{V}_{inj}$. The phasor diagram can be used to explain the three compensation schemes analytically, assuming that the voltage sag/swell is balanced. The general case of unbalanced sag/swell will be addressed later. The situation under which the phasor diagram applies is when the sag has entered into a steady state.

Figure 3-2 A general phasor diagram showing compensation of (a) a voltage sag and (b) a voltage swell.

In Figure 3-2, the compensated load current phasor $\vec{I}$ is used as the reference, $\vec{V}$ is the compensated load voltage, ϕ is the load power factor angle, and $\vec{V}_{pre}$ is the pre-fault phasor of $\vec{V}_s$. θ is the phase shift between $\vec{V}_{pre}$ and $\vec{V}_s$. It is defined such that if $\vec{V}_s$ lags $\vec{V}_{pre}$, θ is positive. As $\vec{V}_s$ is the consequence of an upstream fault or disturbance, its magnitude and phase are not controllable. Note that in arriving at Figure 3-2, the compensated load-side voltage magnitude $|\vec{V}|$ is restored to its pre-disturbance value $|\vec{V}_{pre}|$. ψ is the phase difference between $\vec{V}_{pre}$ and $\vec{V}$. It is defined thus: if $\vec{V}$ lags $\vec{V}_{pre}$, then ψ is positive.

Based on Figure 3-2, therefore it can be seen that the total injected active power from the DVR under such a balanced condition is:

$$P_{inj} = 3|\vec{V}_{inj}||\vec{I}|\cos\beta = 3[|\vec{V}|\cos\phi - |\vec{V}_s|\cos(\phi + \psi - \theta)]|\vec{I}| \quad (3.2.1)$$

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where β is the phase difference between $\vec{I}$ and $\vec{V}_{inj}$. Equation (3.2.1) will be applicable for any of the three compensation methods described earlier by just considering how ψ is oriented, as described next.

- **In-phase compensation method**

Since this method requires the compensated load voltage $\vec{V}$ to be in phase with $\vec{V}_s$, then ψ would be equal to θ . Using this method, the corresponding phasor diagrams for voltage sag and swell mitigation are as shown in Figure 3-3.

Figure 3-3 In-phase compensation method for mitigating (a) a voltage sag and (b) a voltage swell.

From the above figures, it is noticed that the injected active power from the DVR is:

$$P_{inj} = 3[|\vec{V}| \cos \phi - |\vec{V}_s| \cos(\phi)]|\vec{I}| \quad (3.2.2)$$

During a voltage sag, i.e. $|\vec{V}| < |\vec{V}_s|$, (3.2.2) indicates that this compensation method will cause a net injection of active power from the DVR into the external interconnected system. Conversely under a voltage swell, since $|\vec{V}| > |\vec{V}_s|$, P_{inj} will be negative. In other words, if in-phase compensation method is to be used to mitigate a voltage swell, the DVR will absorb active power from the external system. The absorbed power would have to be stored in the energy storage system and the result would be a rise of its terminal voltage V_d . The detailed analysis on the extent of the voltage rise is shown in the Appendix A. If the compensation process were to last for a relatively long time, V_d could

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exceed its design limit. Hence an alternative means has to be found to overcome this problem.

- ***Pre-fault compensation***

Pre-fault compensation has often been adopted in practice since it would result in an almost undisturbed load voltage. By continuously tracking $\vec{V}_s$, $\vec{V}_{inj}$ is introduced in such a way that $\vec{V}$ is compensated to its pre-fault condition $\vec{V}_{pre}$. Therefore, ψ would be 0 and the injected active power is:

$$P_{inj} = 3[\vec{V}|\cos\phi - |\vec{V}_s|\cos(\phi - \theta)]|\vec{I}| \quad (3.2.3)$$

From this equation, one observes that if the magnitude of $\vec{V}_s$ is such that

$$|\vec{V}_s| > \frac{|\vec{V}|\cos\phi}{\cos(\phi - \theta)}$$

then $P_{inj} < 0$. Under such a voltage disturbance condition, the DVR absorbs active power from the external system and similar to the in-phase compensation method for a voltage swell, it will cause a rise of V_d during the restoration process.

On the other hand, if

$$|\vec{V}_s| < \frac{|\vec{V}|\cos\phi}{\cos(\phi - \theta)}$$

this leads to $P_{inj} > 0$. The DVR will inject active power to the external system and V_d would decrease. This is the usual operational mode considered in the literature when pre-fault compensation method has been applied.

The above observation can also be elaborated using Figure 3-4. For practical and more likely scenario, $\vec{V}_s$ is assumed to lie within the first quadrant of the complex plane, with respect to the compensated load current $\vec{I}$ [2].

In Figure 3- 4, for any voltage sag/swell which results in the phasor $\vec{V}_s$ to lie to the left of the vertical line “ $|\vec{V}|\cos\phi$ ” (as for example $\vec{V}_{s2}$), the DVR will inject active power into

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the system. This can be seen from (3-2-1) for since $|\beta| = |\beta_2| < 90^\circ$, therefore $P_{inj} = 3|\vec{V}_{inj}||\vec{I}|\cos\beta_2 > 0$. This case will be similar with most of sag compensation.

Conversely, if a voltage disturbance results in $\vec{V}_s$ (such as $\vec{V}_{s1}$) lying any where within the shadowed area (i.e. to the right of the line $|\vec{V}|\cos\phi$), this will lead to a rise of V_d because the DVR absorbs active power from the external system since $|\beta_1| > 90^\circ$. Therefore, a rise in V_d level is also possible if one uses the pre-fault compensation method. Again, the pre-fault compensation method may have to be modified in order to control the voltage across the energy storage capacitor.

Figure 3-4 Phasor diagram illustrating Pre-fault compensation method

- **Energy-saving compensation:**

The purpose of energy-saving compensation is to introduce $\vec{V}_{inj}$ to effect a phase angle adjustment in the load-side voltage so as to maximize the load ride-through capability offered by the DVR energy storage device. The load-side compensated voltage $\vec{V}$ will have a phase shift ψ with respect to its pre-fault phasor $\vec{V}_{pre}$. By adjusting ψ through the control of the VSI, the injected active power through the DVR could be controlled. From (3.2.1) and when $\cos(\phi + \psi - \theta) = 1$, the injected active power would reach its minimum value:

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$$P_{\min} = 3(|\vec{V}| \cos \phi - |\vec{V}_s|) |\vec{I}| \quad (3.2.4)$$

i.e. when $\psi = \psi_{\min} = \theta - \phi$. Under such a minimum power injection strategy and when

$$|\vec{V}_s| < |\vec{V}| \cos \phi,$$

$P_{\text{inj}} = P_{\min} > 0$. It means that the DVR must inject active power into the system to compensate for this kind of disturbance, a situation depicted in Figure 3-5(a) and which inevitably means a voltage sag. The minimum power injection strategy would therefore allow the voltage sag to be restored for a longer interval before a given amount of stored energy in the DVR is depleted [25].

Conversely, this compensation strategy also means the maximum absorption scenario when $|\vec{V}_s| > |\vec{V}| \cos \phi$ and this will accelerate the dc-link voltage rise. Indeed, in cases when the dc-link voltage is lower than its rated voltage, one may adopt the minimum power injection strategy to absorb active power from the grid in order to increase V_d . This point will be explored further in Section IV. Notice that as $\cos \phi \leq 1$, then a voltage swell will automatically mean that $|\vec{V}_s| > |\vec{V}| \cos \phi$. Hence if the minimum power injection strategy is used to mitigate the swell, it will unavoidably cause a rise in V_d level. This is described by the phasor diagram of Figure 3-5(b).

Figure 3-5 Minimum power injection strategy

In order to control the rise in V_d , it is proposed that one applies a zero power injection strategy. From (3.2.1), if $P_{\text{inj}} = 0$, then $\psi = \psi_0$ where

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$$\psi_0 = \theta - \phi + \arccos\left(\frac{|\vec{V}| \cos \phi}{|\vec{V}_s|}\right) \quad (3.2.5)$$

Since for any voltage swell, $|\vec{V}_s| > |\vec{V}| \cos \phi$. Therefore, zero power injection can always be realized for any voltage swell condition. The corresponding phasor diagram for this injection strategy is as shown in Figure 3-6 where it is shown that $\vec{V}_{inj}$ will be perpendicular to the compensated load current $\vec{I}$. From trigonometry, it can also be readily shown that $\vec{V}$ will be displaced from $\vec{V}_{pre}$ by the angle ψ_0 , given by (3.2.5). Thus if one adopts $\vec{V}_{pre} \angle \psi_0$ as the voltage reference phasor for the compensation control system during a voltage swell, the injected voltage phasor $\vec{V}_{inj}$ will be perpendicular to the compensated load current $\vec{I}$. As the injected power is zero, the dc-link voltage rise can be avoided.

In summary, among the three compensation schemes, both the in-phase and pre-fault compensation methods could cause the rise of the dc-link voltage during the process of mitigating a voltage swell. Unless additional measures are taken to control the voltage rise, this may endanger the normal operation of the DVR. Only the energy-saving method is shown to be able to mitigate the effect of the voltage swell by adopting the zero-power injection strategy. It thus avoids the possibility of an excessive voltage rise across the dc-link.

Figure 3-6 Zero power injection strategy

3.3 Unbalanced Sag/Swell

The above analysis has been carried out for balanced voltage sag/swell. In practical design, the three phases of the DVR are likely to share a common energy storage system to reduce cost. Hence when the voltage restoration device is to be used to compensate for an unbalanced voltage disturbance, it is possible that not all the phases are injecting active power into the external system. Under such a condition, it is not only necessary to restore the magnitude of load-side phase voltage to its nominal level and maintain a balanced supply, but also V_d will have to be controlled over the compensation process. In this Section, the energy-saving compensation method will be extended to cater for such an unbalanced disturbance event.

Figure 3-7 depicts the consequence of an upstream unbalanced fault on the three phases of $\vec{V}_s$ where the pre-fault voltages are denoted as $\vec{V}_{Apre}$, $\vec{V}_{Bpre}$ and $\vec{V}_{Cpre}$ for the respective phases and where the standing voltages during sag/swell are $\vec{V}_A$, $\vec{V}_B$ and $\vec{V}_C$ respectively. The corresponding phase shifts are θ_A , θ_B and θ_C , as shown.

Figure 3-7 Source-side voltages due to an upstream unbalanced fault

As the load is assumed balanced, then the load current and the compensated load-side phase voltages must also be balanced. Therefore, there should be the same phase shift ψ between the load-side pre-disturbance and compensated voltages of each phase. The total active power flow between the DVR and the external system can be calculated by referring to Figure 3-8 where the compensated phase A load current has been adopted as

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the reference phasor. Rotate $\vec{V}_{Bpre}$ by $+120^\circ$ and $\vec{V}_{Cpre}$ by -120° so that these voltages coincide with $\vec{V}_{Apre}$. Then also rotate $\vec{V}_B$ by $+120^\circ$ and $\vec{V}_C$ by -120° and denote these voltages by $\vec{V}_B'$ and $\vec{V}_C'$ respectively. The phase difference between $\vec{V}_{Apre}$ and $\vec{V}_B'$ is therefore maintained at θ_B , and that between $\vec{V}_{Apre}$ and $\vec{V}_C'$ is θ_C . Using the same symbols as in Figure 3-2, the total injected power is given by:

$$P = 3|\vec{V}||\vec{I}|\cos\phi - [|\vec{V}_A|\cos(\phi + \psi - \theta_A) + |\vec{V}_B|\cos(\phi + \psi - \theta_B) + |\vec{V}_C|\cos(\phi + \psi - \theta_C)]|\vec{I}| \quad (3.3.1)$$

Expanding the last expression and after some manipulation, one obtains

$$P = 3|\vec{V}||\vec{I}|\cos\phi - \sqrt{M^2 + N^2}|\vec{I}|\cos(\phi + \psi - \arctg \frac{M}{N}) \quad (3.3.2)$$

where

$$M = |\vec{V}_A|\sin\theta_A + |\vec{V}_B|\sin\theta_B + |\vec{V}_C|\sin\theta_C \quad (3.3.3)$$

$$N = |\vec{V}_A|\cos\theta_A + |\vec{V}_B|\cos\theta_B + |\vec{V}_C|\cos\theta_C \quad (3.3.4)$$

When $\cos(\phi + \psi - \arctg \frac{M}{N}) = 1$, that is

$$\psi = \psi_{\min} = \arctg \frac{M}{N} - \phi \quad (3.3.5)$$

one arrives at the minimum injected power condition similar to that described earlier when deriving (3.2.4). The minimum injected power is

$$P_{\min} = (3|\vec{V}|\cos\phi - \sqrt{M^2 + N^2})|\vec{I}| \quad (3.3.6)$$

Figure 3-8 Phasor diagram showing the relationship between source-side and the compensated load-side voltages and current.

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From (3.3.2), note that if, $\frac{\sqrt{M^2 + N^2}}{3} \geq |\vec{V}| \cos \phi$, zero power injection can also be realized by the suitable setting of the voltage phase shift ψ . In which case, ψ between the pre-fault and compensated load-side voltage is

$$\psi_0 = \arctg \frac{M}{N} - \phi - \arccos\left(\frac{3|\vec{V}| \cos \phi}{\sqrt{M^2 + N^2}}\right) \quad (3.3.7)$$

It is interesting to note that one can arrive at (3.2.4) from (3.3.6) by equating

$$|\vec{V}_s| = \frac{\sqrt{M^2 + N^2}}{3}, \text{ and } \theta = \arctg \frac{M}{N}.$$

From the above, one can calculate the positive phase sequence component of $\vec{V}_s$,

$$\vec{V}_{s,pps} = \frac{1}{3}(N + jM) = \frac{1}{3}(\vec{V}_A + \vec{V}_B' + \vec{V}_C') = |\vec{V}_{s,pps}| \angle \theta \quad (3.3.8)$$

Thus, in the case of unbalanced voltage disturbance, one only needs to calculate $\vec{V}_{s,pps}$ so that it can be used to decide on the compensation method to be used. A generalized compensation strategy for the restoration of load-side voltage may now be obtained and this is given in the next section.

3.4 A Generalized Compensation Strategy

As discussed, unbalanced voltage disturbances can be considered in a similar manner as that for a balanced one by considering the positive phase sequence voltage of $\vec{V}_s$. Also from the previous section, one notes that for a balanced disturbance, if $|\vec{V}_{s,pps}| < |\vec{V}| \cos \phi$, the DVR has to inject active power into the external system and the dc-link voltage must decrease during the compensation process. Under this condition, if the intention is to reduce the rate of the decrease of V_d and therefore extend the capacity of the DVR to compensate for the sag, the minimum power injection method can be used for better utilization of the capacity of the energy storage system. On the other hand, if $|\vec{V}_{s,pps}| > |\vec{V}| \cos \phi$, and since the compensation process might cause the rise of dc-link

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voltage, one may wish to ensure the dc-link voltage remains close to its rated value. Therefore, the zero-power injection strategy can be used under this condition.

Furthermore, if the dc-link voltage is lower than its rated value when such a disturbance occurs, one may initially use the minimum power injection method (which is also the maximum power absorption condition if $|\vec{V}_{s,pps}| > |\vec{V}| \cos \phi$) to absorb active power from the external system. This helps to increase the dc-link voltage. Once the dc-link voltage has reached its rated value, zero-power injection strategy can then follow to keep the dc-link voltage constant.

Based on the above reasoning, let the load-side power factor angle be ϕ and the pre-disturbance source-side voltages as $\vec{V}_{Apre} = |\vec{V}_{pre}| \angle \phi_{pre}$, $\vec{V}_{Bpre} = |\vec{V}_{pre}| \angle (\phi_{pre} - 120^\circ)$, $\vec{V}_{Cpre} = |\vec{V}_{pre}| \angle (\phi_{pre} + 120^\circ)$. The pre-fault voltage will be adopted as the reference for the respective phase. The compensating strategy can be described as follows.

Step 1: As soon as an upstream voltage disturbance is detected, adopt the pre-fault compensation method to restore the load-side voltage to its nominal level as rapidly as possible. The injection voltage corresponds to that of Figure 3.4.

Step 2: Determine the source-side voltage magnitudes $|\vec{V}_A|$, $|\vec{V}_B|$, $|\vec{V}_C|$ and phase θ_A , θ_B , θ_C (with respect to their respective pre-fault quantities). Calculate M, N and then $\vec{V}_{s,pps}$ using (3.3.3), (3.3.4) and (3.3.8). Calculate ψ_{min} using (3.3.5).

Step 3: If $|\vec{V}_{s,pps}| < |\vec{V}_{pre}| \cos \phi$, go to **step 4**.

Otherwise, go to **step 5**.

Step 4: Minimum power injection strategy is to be applied. Set the new reference

voltages as $\vec{V}_{Aref} = |\vec{V}_{pre}| \angle (\phi_{pre} + \psi_{min})$, $\vec{V}_{Bref} = |\vec{V}_{pre}| \angle (\phi_{pre} - 120^\circ + \psi_{min})$,

$\vec{V}_{Cref} = |\vec{V}_{pre}| \angle (\phi_{pre} + 120^\circ + \psi_{min})$, where $\psi_{min} = \theta - \phi$.

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DVR injects voltages $\vec{V}_{Ainj} = \vec{V}_{Aref} - |\vec{V}_A| \angle \theta_A$,

$\vec{V}_{Binj} = \vec{V}_{Bref} - |\vec{V}_B| \angle (\theta_B - 120^\circ)$, $\vec{V}_{Cinj} = \vec{V}_{Cref} - |\vec{V}_C| \angle (\theta_C + 120^\circ)$.

Load-side voltage will then recover to its normal magnitude but with a phase shift ψ_{min} with respect to its pre-sag $\vec{V}_{pre}$ phasor. Go to Step 8.

Step 5: Check if V_d is less than its rated value. If so, go to **step 6**. Otherwise go to **Step 7**.

Step 6: Maximum power absorption strategy is to be applied. Set the new reference

voltages $\vec{V}_{Aref} = |\vec{V}_{pre}| \angle (\phi_{pre} + \psi_{min})$, $\vec{V}_{Bref} = |\vec{V}_{pre}| \angle (\phi_{pre} - 120^\circ + \psi_{min})$,

$\vec{V}_{Cref} = |\vec{V}_{pre}| \angle (\phi_{pre} + 120^\circ + \psi_{min})$, where $\psi_{min} = \theta - \phi$. Since $|\vec{V}_{s,pps}| \geq |\vec{V}_{pre}| \cos \phi$,

$P_{min} \leq 0$. DVR will absorb real power from the external system to increase the terminal voltage of the energy storage dc-link. Once V_d reaches its rated value, go to **step 7**. Otherwise go to Step 8.

Step 7: Set the new reference voltages $\vec{V}_{Aref} = |\vec{V}_{pre}| \angle (\phi_{pre} + \psi_0)$,

$\vec{V}_{Bref} = |\vec{V}_{pre}| \angle (\phi_{pre} - 120^\circ + \psi_0)$, $\vec{V}_{Cref} = |\vec{V}_{pre}| \angle (\phi_{pre} + 120^\circ + \psi_0)$, where ψ_0 is given

by (3.2.5). With this reference voltage set for the DVR control system, the zero-power injection strategy is realized and V_d can be kept constant.

Step 8: Check load-side voltage magnitudes to see if they are within tolerance of their pre-fault values. If they are not, go to Step 2. Otherwise maintain the injection level until the next voltage disturbance.

The above generalized compensation strategy provides a new method on mitigating voltage sag as well as for voltage swell. It is applicable for balanced and unbalanced voltage disturbances. In term of the practical implementation of the scheme, a digital micro controller can be used, through which the supply voltage parameter estimation and the proposed strategy can be implemented. A software phase locked loop (PLL) can be used to generate the reference (or pre-sag) load voltage. Once a sag is detected, the PLL records the pre-sag voltage phasor while the controller detects the sag voltage and phase

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shift. From the results of Section 3, the required phase reference is calculated. The reference voltage vector can be sent to the pulse width modulator of the micro controller. The modulator will then direct the gate signals to the switching devices in the VSI, according to the modulated signal to synthesize the reference voltage [85]. In this Chapter, the control strategy is illustrated by the simulation examples shown in the next section.

3.5 Illustrative Examples

The following examples will illustrate the application of the generalized compensation strategy proposed earlier. The network disturbance scenario considered begins with a balanced three-phase fault that causes a severe balanced voltage sag on $\vec{V}$, the downstream load bus voltage shown in Figure 3-9. Furthermore, due to the under-voltage release actions of the load protection relays, it is assumed that a proportion of the loads have been disconnected from the supply. The partial load rejection coupled with the delay in the voltage regulation actions of upstream generators has resulted in a voltage swell following the fault clearance.

In simulating the restoration process of the DVR, the injection strategy based on that described in Section 3.4 is used. In carrying out the study, the controller with a feed-forward loop intended for DVR and described in [77] is adopted. The controller is designed so as to offer good dynamic response characteristics against supply and load disturbances. The size of the capacitor used in the dc link is 0.01F.

Figure 3-9 shows the source-side bus voltage profile (only phase A is included) where it has been assumed that:

- At $t = 0.06$ s, the three phase fault occurs and leads to a voltage sag;
- At $t = 0.2$ s, the fault is cleared from the system, but due to the partial load rejection, a voltage swell follows;
- At $t = 0.4$ s, the voltage swell has subsided because the excitation system of the upstream generator has successfully regulated the network voltage.

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The rated bus voltage is assumed to be 220 V, the rated load power is 30 kVA, 0.8 pf lagging. The voltage sag introduced is 0.5 p.u., and with a phase shift of $+25^\circ$. The voltage swell is about 1.24 p.u. and the phase shift is also about $+25^\circ$.

Figure 3-9 Source-side Voltage (Phase A) Under Disturbance

For the purpose of comparison, suppose the pre-fault and the generalized compensation methods will be used to mitigate the disturbances. The results of the pre-fault compensation are as shown in Figure 3-10, and Figure 3-11 shows the results under the generalized compensation method.

For pre-fault compensation method, the pre-fault voltage is always adopted as the reference voltage. The phasor diagrams for the sag and swell stages are as shown in Figure 3-12.

From Figure 3-11(b) and 3-11(c), it is clear that the generalized compensation process can be divided into four intervals:

For $t = 0.06 \text{ s} - 0.08 \text{ s}$, the pre-fault compensation method is adopted to restore the load-side voltage to its nominal level as rapidly as possible. The source-side voltage, and their respective phase shifts are determined over this interval.

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Figure 3-10. DVR under pre-fault compensation: (a) Load-side voltage (phase A), (b) Voltage across DVR energy storage capacitor, (c) Active power (solid line) and reactive power (dashed line) from DVR.

For $t = 0.08 \text{ s} - 0.2 \text{ s}$, since $|\vec{V}_{\text{sag}}| < |\vec{V}| \cos \phi$, the minimum power injection strategy has been adopted for correcting the voltage sag. Based on the results of the analysis, the compensated load voltage is shown to have a phase shift of -12° with respect to its pre-fault level, as Figure 3-13(a) shows;

For $t = 0.2 \text{ s} \sim 0.31 \text{ s}$, over this interval, as the dc-link voltage is lower than its rated value, the minimum injection method is still applied to absorb active power from the external

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system. This helps the recovery of V_d . The corresponding phasor diagram is Figure 3-13(b).

Figure 3-11 DVR under generalized compensation scheme: (a) Load-side voltage (phase A), (b) Voltage across DVR energy storage capacitor, (c) Active power (solid line) and reactive power (dashed line) from DVR.

For $t = 0.31s - 0.4s$, since V_d has reached its rated value, the zero-power injection strategy is adopted to keep V_d constant until the end of compensation. Based on the results of the analysis, it can be readily shown that the compensated voltage would have a phase shift of 38° with respect to its pre-fault level. The phasor diagram is as shown in Figure 3-14.

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Figure 3-12. Phasor diagram showing pre-fault compensation method for: (a) the voltage sag (b) the voltage swell stages.

Figure 3-13 Minimum power injection: (a) sag, (b) swell

Figure 3-14 Zero power injection method

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Comparing the pre-fault compensation method with that of the generalized strategy, one notes that:

1. Pre-fault compensation method offers an almost undisturbed load voltage since the reference voltage under the pre-fault control method is kept constant. This can be verified by comparing Figure 3-10(a) and Figure 3-11(a). In Figure 3-11(a), at time $t = 0.31$ s, there is a clear disturbance on the load voltage waveform because the phase angle of the reference voltage under the generalized strategy has been shifted from -12° to 38° .
2. Under the pre-fault compensation method, if no limit is placed on the capacitor voltage, the voltage has exceeded its rated value (See Figure 3-10(b)). On the other hand, under the generalized compensation method, when the capacitor voltage reaches its rated value, the zero-power injection strategy replaces the minimum power injection scheme. The rise of the dc-link voltage is avoided, as Figure 3-11(b) shows.
3. Before $t = 0.3$ s, since the reference voltages of the pre-fault and the generalized compensation methods only results in a 12° phase difference in the restored load voltage, therefore the injected active powers under these two different compensation methods are almost the same. See Figure 3- 10(c) and Figure 3-11(c). This has resulted in nearly the same profile on the capacitor voltages for $t < 0.3$ s, as Figure 3-10(b) and Figure 3-11(b) show.

Although only balanced fault disturbance has been considered in this example, a certain amount of DC-component in the phase voltage/current waveforms are to be expected. However, the analysis in the next chapter shows that the rate of the decay of the DC-components depends on the L/R ratio of the static load. For the 0.8 p.f. (lagging) static load assumed in the study, the load will have a time constant of approximately 2.4 ms. Hence, the duration of the DC-component is less than 10 ms in this example. It explains why such voltage/current offset is not discernible in the study.

3.6 Conclusions

The restoration of load voltage through a VSC-based series compensator has been considered. It is shown that during voltage sag, the restoration process almost inevitably requires energy injection from the DVR to the external system. However, maintaining the load-side voltage during a voltage swell could lead to a rise in the VSC energy storage voltage V_d . This is because the DVR device will have to absorb active power from the external network. It is shown that only the energy-saving method, incorporating the zero-power injection strategy, is capable of maintaining V_d constant. The method is based on the adjustment of the phase angle of the injected voltage.

For unbalanced voltage disturbances, the analysis shows that it is only necessary to consider the positive phase sequence component of the voltages. The same injection strategy as that for balanced disturbances can be applied. From the analytical results obtained, a generalized compensation strategy for voltage sag/swell disturbances has been proposed. With this method, the load voltage can be restored for sag and swell while V_d can also be controlled.

Despite the generalized compensation strategy is a very attractive method to improve the performance of DVR, one also notices that during the compensation, phase shift on the supply voltage is also unavoidable if the optimized energy injection is to be applied. Some types of loads such as induction motors might be very sensitive to this phase shift. Therefore, the response of various kinds of loads under such phase shift will need to be studied. This is addressed in the next Chapter.

Chapter 4

Load Response Under the Voltage Phase Shift

From Chapter 3, it is clear that the generalized voltage compensation strategy is considered advantageous as it can increase the protected load ride-through capability, given the limited capacity of the energy storage device of the restorer [30]. However, by injecting a controllable voltage onto the load, it can also cause a phase shift on load terminal voltage. While detailed analysis will be given later, it is inevitable that the phase shift will introduce a certain level of dynamic transients in the loads. And as the DVR is in series connection with the load, the power module of the DVR inverter may also be subject to the transients. Therefore, implementation of the generalized voltage compensation control of the DVR has to be considered in conjunction with load responses.

In the previous reported works on series compensation, often it had been assumed that the transient response would diminish quickly and the corresponding effect could be ignored. No attempt has been made to study the load response under the voltage phase shift and the impacts of the load transient on the DVR device. The research work to be described in this Chapter is an attempt to obtain some qualitative measures of the interactions between the DVR and loads.

In the following part of this Chapter, the load responses under the voltage phase shift will be analyzed. Static linear load and induction motor are selected as typical load models because they are most commonly found in practice. Furthermore, since some induction motors are connected to the power system through adjustable speed-drives, the effect of voltage phase shift on the adjustable speed-drive systems will also be studied.

4.1 dq Transformation and Space Phasor

A rotating motor load will be considered in detail, the use of d-q transformation technique will be reviewed first. This is because analysis of the rotating motor load behavior is greatly simplified by the use of the transformation. It refers all voltage/current quantities to a reference frame rotating at a speed, defined as ω_x [78].

Figure 4-1 Orientation of three phase and d-q systems in a complex plane

The dq variables are related to the three-phase machine variables by equation (4-1-1).

The phasor diagram relationship between the abc axis and dq axis is as shown in Figure 4-1. As γ_0 , the initial position of the axis of phase “a” voltage with respect to the d axis, is arbitrary for a balanced system, one may choose $\gamma_0 = 0$.

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos(\omega_x t + \gamma_0) & \cos(\omega_x t + \gamma_0 - \frac{2}{3}\pi) & \cos(\omega_x t + \gamma_0 + \frac{2}{3}\pi) \\ -\sin(\omega_x t + \gamma_0) & -\sin(\omega_x t + \gamma_0 - \frac{2}{3}\pi) & -\sin(\omega_x t + \gamma_0 + \frac{2}{3}\pi) \end{bmatrix} \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix} \quad (4-1-1)$$

In the equation, v_a, v_b, v_c are the phase voltages of phases a, b, c respectively; ω_x is the rotating speed of the reference frame; γ_0 is the phase shift between the stator magnetic field and the axis of phase a of the stator of the three-phase machine.

For facilitating the formulation and solution of electric machine equations, the concept of space phasors is introduced. In fact, the space phasor is an alternative to the orthogonal-axis model [78]. Denote:

$$\bar{V} = \bar{V}_d + j\bar{V}_q \quad (4-1-2)$$

The quantity of $\vec{V}$ is called the direct space phasor of the voltage. The direct space phasor describes completely the machine terminal voltages if the terminal voltages are symmetric during both steady state and transients.

Then, from equations (4-1-1) and (4-1-2) with $\gamma_0 = 0$:

$$\vec{V} = \sqrt{\frac{2}{3}}(\vec{V}_a + \vec{V}_b e^{j\frac{2}{3}\pi} + \vec{V}_c e^{-j\frac{2}{3}\pi})e^{-j\omega_s t} \quad (4-1-3)$$

If it is a balanced system, the supply voltages are of the forms:

$$V_{a,b,c}(t) = \sqrt{2}V \sin(\omega_1 t + \gamma - \frac{2}{3}\pi(k-1)) \quad (4-1-4)$$

where $k = 1, 2, 3$. ω_1 is the synchronous speed (corresponds to the power supply frequency), and γ corresponds to the initial angle of the phase voltage. In the d-q frame, γ could also be considered to correspond to the instant when the voltage phase shift occurs.

Substituting equation (4-1-4) into (4-1-3), the space phasor of voltage can be written as:

$$\vec{V} = \sqrt{\frac{2}{3}}(\vec{V}_a + \vec{V}_b e^{j\frac{2}{3}\pi} + \vec{V}_c e^{-j\frac{2}{3}\pi})e^{-j\omega_s t} = -j\sqrt{3}Ve^{j(\omega_1 - \omega_s)t + j\gamma} \quad (4-1-5)$$

Similar expressions can also be written for the current quantities.

With the transformation of the reference frame, the phase-angle jump of the load voltage under abc reference frame is seen as the changes in space phasor in the dq frame.

4.2 Load Modeling

Load modeling is a difficult problem because power system loads are aggregates of many different devices. However, the proposed energy-saving control strategy described in Chapter 3 for the DVR can simplify the analysis somewhat since the DVR action would only result in the load experiencing possible changes in the voltage phase angle. There shall be no change in the voltage magnitude.

4.2.1 Linear Static Load

In its simplest form, it is reasonable to represent loads by linear series resistive-inductive (RL) circuits [25]. For a general three-phase RL circuit, the relationship between the current and voltage of the load is:

$$L \frac{di_k}{dt} + Ri_k = v_k(t), \quad k = a, b, c \quad (4-2-1)$$

It also can be written in the form,

$$(R + L \frac{d}{dt}) \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} = \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix} \quad (4-2-2)$$

Or transform into the dq frame

$$\begin{bmatrix} R + L \frac{d}{dt} & -\omega_1 L \\ \omega_1 L & R + L \frac{d}{dt} \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} = \begin{bmatrix} v_d \\ v_q \end{bmatrix} \quad (4-2-3)$$

where ω_1 is the synchronous speed (corresponds to the power supply frequency).

Adopting the concept of space phasor, set $p = \frac{d}{dt}$, $\vec{I} = I_d + jI_q$ and $\vec{V} = V_d + jV_q$, then

equation (4-2-3) can be rewritten as:

$$\vec{I} = \frac{\vec{V}}{R + pL + j\omega_1 L} \quad (4-2-4)$$

The above equation represents the RL circuits and is applicable for the analysis of both the steady state and transient state behavior of the load.

4.2.2 Induction Motor Load

Induction motors have been widely used in industry. Comparing to other rotating machinery, the existence of the large inductances in the induction motors could weaken their ride-through capability. Motors are thought to be particularly vulnerable to voltage dips. Therefore, in this part, it is reasonable to regard the induction motors as typical example of industry loads.

Generally, transients of the induction motors will include two parts: electromagnetic transients and electromechanical transients [78-80]. Notice that voltage sag phenomenon

is usually associated with fault and its subsequent clearance; its effect is only observed over an interval of a few cycles of the mains frequency. Thus for such a short time, it is the electromagnetic transients of the DVR-motor system which are dominant. Over this period, it is also reasonable to assume that the motor mechanical speed (ω_m) is constant [78]. In other words, only the electromagnetic transient of the induction motor is considered and the electromechanical transient is ignored.

From [78], it can be seen that the direct space phasor equations of an induction motor in a reference frame rotating at some constant speed ω_x are given by:

$$\vec{V}_s = R_s \vec{I}_s + p\vec{\lambda}_s + j\omega_x \vec{\lambda}_s \quad (4-2-5)$$

$$\vec{V}_r = R_r \vec{I}_r + p\vec{\lambda}_r + j(\omega_x - \omega_m) \vec{\lambda}_r \quad (4-2-6)$$

$$\vec{\lambda}_s = L_s \vec{I}_s + L_m (\vec{I}_s + \vec{I}_r) \quad (4-2-7)$$

$$\vec{\lambda}_r = L_r \vec{I}_r + L_m (\vec{I}_s + \vec{I}_r) \quad (4-2-8)$$

where $\vec{V}_s, \vec{V}_r$: the space phasors of stator and rotor voltages; $\vec{I}_s, \vec{I}_r$: the space phasors of stator and rotor currents; $\vec{\lambda}_s, \vec{\lambda}_r$: the space phasors of stator and rotor flux linkages; R_s, R_r : the stator and rotor resistances; L_s, L_r : the stator and rotor leakage inductances; L_m : the magnetizing inductance; ω_m : the mechanical speed of the rotor; ω_x : the rotated speed of the reference frame. The corresponding induction motor circuit is as shown in Figure 4-2.

Figure 4-2 Induction motor circuit

Notice that the model is highly non-linear and is not amenable to simple analysis. Hence the intention is to obtain a simplified relationship between the stator voltage and stator current in order to facilitate the analysis.

As the rotor windings are always short-circuited, it means that $\vec{V}_r = 0$. From (4-2-6), one obtains:

$$\vec{\lambda}_r = \frac{-R_r \vec{I}_r}{p + j(\omega_x - \omega_m)} \quad (4-2-9)$$

Substituting (4-2-9) into (4-2-8), then

$$L_r \vec{I}_r + L_m (\vec{I}_s + \vec{I}_r) = \frac{-R_r \vec{I}_r}{p + j(\omega_x - \omega_m)}$$

That is

$$\vec{I}_r = - \frac{L_m \vec{I}_s}{L_r + L_m + \frac{R_r}{p + j(\omega_x - \omega_m)}} \quad (4-2-10)$$

Substituting (4-2-10) into (4-2-7),

$$\vec{\lambda}_s = (L_s + L_m) \vec{I}_s - \frac{L_m^2 \vec{I}_s}{L_r + L_m + \frac{R_r}{p + j(\omega_x - \omega_m)}} = L(p) \vec{I}_s \quad (4-2-11)$$

$$\text{where } L(p) = (L_s + L_m) - \frac{L_m^2}{L_r + L_m + \frac{R_r}{p + j(\omega_x - \omega_m)}}$$

During the transient process, it means that at the initial instant of step voltage change, $t \Rightarrow 0$, then $p = \frac{d}{dt} \Rightarrow \infty$. Generally, for a motor, it can be assumed that the reference frame to be attached on the motor rotor, that is $\omega_x = \omega_m$. In the more likely situation that several motors with different rotating speeds are connected in parallel, a situation that will be discussed later, in order to reduce the error of the calculation, one could adopt the average rotating speed of these motors as the reference-frame speed. Compared to the large value of p at the initial time of step voltage change, the difference between ω_x and ω_m could be ignored. Hence, let L_{tr} denotes the approximate transient inductance of the motor, i.e.,

$$L_{tr} = \lim_{p \rightarrow \infty} L(p) = (L_s + L_m) - \frac{L_m^2}{L_r + L_m} = L_s + \frac{L_r}{\frac{L_r}{L_m} + 1} \quad (4-2-12)$$

Note that in practical motors, $L_r \ll L_m$, therefore

$$L_{tr} \approx L_s + L_r \quad (4-2-13)$$

On the other hand, if $t \Rightarrow \infty$, that means the motor is operating on steady-state, then $p \Rightarrow 0$, therefore, equation (4-2-12) is changed to:

$$L_{ss} = \lim_{p \rightarrow 0} L(p) = L_s + L_m - \frac{L_m^2}{L_r + L_m + \frac{R_r}{j(\omega_x - \omega_m)}} \quad (4-2-14)$$

Under such a condition, assuming the reference frame is rotating at the synchronous speed ω_1 , then

$$\omega_x - \omega_m = \omega_1 - \omega_m = s\omega_1$$

where s is the slip of the motor. Then, from (4-2-14):

$$L_{ss} = L_s + L_m \frac{\left(\frac{R_r}{s} + j\omega_1 L_r\right)}{j\omega_1 L_m + \left(\frac{R_r}{s} + j\omega_1 L_r\right)} \quad (4-2-15)$$

As the slip is usually small, $s \Rightarrow 0$, this also means that $\omega_1 \approx \omega_m$, $L_{ss} \approx L_s + L_m$

L_{ss} represents the approximate steady-state inductance of the induction motor.

Substitute (4-2-11) into (4-2-5), one obtains:

$$\vec{V}_s = R_s \vec{I}_s + p\vec{\lambda}_s + j\omega_x \vec{\lambda} = R_s \vec{I}_s + (p + j\omega_x)L(p)\vec{I}_s \quad (4-2-16)$$

Based on the above analysis of the motor under steady state and transient conditions, it is reasonable to re-write the last equation as:

$$\vec{I}_s = \frac{\vec{V}_s}{R + pL + j\omega_x L} \quad (4-2-17)$$

where, $R = R_s$ is the stator resistance and

$$L = \begin{cases} L_s + L_r, & \text{for transient} \\ L_s + L_m \frac{\left(\frac{R_r}{s} + j\omega_1 L_r\right)}{j\omega_1 L_m + \left(\frac{R_r}{s} + j\omega_1 L_r\right)} \approx L_s + L_m, & \text{for steady-state} \end{cases}$$

The corresponding simplified induction motor circuit is shown in Figure 4-3:

Figure 4-3 Simplified induction motor circuit

Equation (4-2-17) and Figure 4-3 represent the motor simplified circuits, applicable for the analysis of the motor transient and steady-state behavior.

It must be pointed out that by comparing the motor stator currents at the transient and steady-state, since L_{tr} would be much larger than L_{ss} , the transient current will also be much larger than the steady-state current. This is the key reason that explains why the stator current under the step voltage phase shift can be several times that under the steady state. This point will be examined in detailed later.

Comparing equations (4-2-4) and (4-2-17), it is obvious that the simplified motor model has a similar structure but not the response characteristics with the static linear loads. Therefore, it is reasonable to analyze these two kinds of loads in the same way during the transient process. However, the transient response of the induction motor due to the stator voltage phase angle shift will be dependent only on the motor electrical parameters and is independent of the mechanical load attached on its rotor shaft. In contrast, the response of the static linear load will be determined by the load demand.

4.2.3 Adjustable speed drive

Adjustable speed motor drives are among the most common power electronic-based industrial equipment [2]. Chapter 2 provides a brief description of its structural frame. A typical motor drive has a three-stage topology: a diode rectifier, dc link or bus, and a

PWM inverter as shown in Figure 2-17. The drive normally supplies variable frequency ac power to a three-phase induction motor.

In such arrangement, the motor drive has some energy stored in the dc link capacitor. It also often uses passive diodes at the “front-end”. Due to the existence of the energy storage, the motor drive will have better ride-through capability when a voltage phase shift occurs. This will be explained in Section 4.3.3.

4.3 Load Response Under Step Voltage Phase Shift

Having established a suitable load model, the response under the voltage phase shift could be examined in the following way. The analysis will begin by examining the response of a single load connected downstream of the DVR, and extend to the more practical situation where parallel loads are connected together with some impedances between the DVR and the loads. The response of adjustable speed drive under voltage phase shift will also be described. Some illustrative examples and experimental results will be given to verify the accuracy of the analysis.

4.3.1 Single Aggregate Load Response

To mitigate the impacts of the voltage sags on the downstream loads, one of the most fundamental requirements of the DVR device is to restore the load voltage to its nominal level as quickly as possible once the voltage sag is detected. Hence the compensating process is very short and one could approximate the voltage change as a step change in phase.

From (4-2-17) and transform it into the s-domain, thus:

$$\Delta \vec{I}(s) = \frac{\Delta \vec{V}}{s} \cdot \frac{1}{L(R/L + j\omega_x + s)} \quad (4-3-1)$$

$\Delta \vec{I}$ and $\Delta \vec{V}$ represent the respective stator current and voltage changes under the step voltage phase shift. In the space vector notation, from (4-1-5), therefore:

$$\Delta \vec{V} = \vec{V} - \vec{V}_{\text{pre}} = -j\sqrt{3}V e^{j(\omega_1 - \omega_x)t} (e^{j(\psi + \gamma)} - e^{j\gamma}) = -j\sqrt{3}V 2 \sin \frac{\theta}{2} e^{j(\omega_1 - \omega_x)t + j(\frac{\pi}{2} + \frac{\psi}{2} + \gamma)} \quad (4-3-2)$$

In arriving at this expression, one can refer to the Figure 4-4, ψ is the voltage phase shift, γ corresponds to the instant when the shift occurs, and ϕ is load power factor angle.

Figure 4-4 Phasor diagram under energy-saving compensation

Transforming (4-3-1) into the time function and substituting (4-3-2) into it, thus:

$$\Delta \vec{I}(t) = -j\sqrt{3}V_2 \sin \frac{\Psi}{2} \left\{ e^{j((\omega_1 - \omega_x)t + \frac{\pi}{2} + \frac{\Psi}{2} + \gamma - \arctg \frac{\omega_x L}{R})} \cdot \frac{1}{\sqrt{R^2 + \omega_x^2 L^2}} \right. \\ \left. - e^{-j((2\omega_1 - \omega_x)t - \frac{\pi}{2} - \frac{\Psi}{2} - (\gamma - \arctg \frac{\omega_x L}{R}))} \cdot \frac{e^{-R/L t}}{\sqrt{R^2 + \omega_x^2 L^2}} \right\}$$

Note that $\arctg(\omega_1 L / R)$ is the load power factor angle ϕ , and from Figure 4.4, it is clear that $\gamma = \phi - \psi$, then the above equation could be rewritten as:

$$\Delta \vec{I}(t) = -j\sqrt{3}V_2 \sin \frac{\Psi}{2} \left\{ e^{j((\omega_1 - \omega_x)t + \frac{\pi}{2} + \frac{\Psi}{2})} \cdot \frac{1}{\sqrt{R^2 + \omega_x^2 L^2}} \right. \\ \left. - e^{-j((2\omega_x - \omega_1)t - \frac{\pi}{2} + \frac{\Psi}{2})} \cdot \frac{e^{-R/L t}}{\sqrt{R^2 + \omega_x^2 L^2}} \right\} \quad (4-3-3)$$

From (4.1.1), one can obtain the phase currents in the actual machine [16]:

$$\Delta \vec{I}_a(t) = \sqrt{\frac{2}{3}} (\Delta \vec{I}_d \cos \omega_x - \Delta \vec{I}_q \sin \omega_x) = \sqrt{\frac{2}{3}} \text{Re}(\Delta \vec{I}(t) e^{j\omega_x t}) \quad (4-3-4)$$

Substituting (4-3-3) into the last equation, phase “a” current change is therefore:

$$\Delta \bar{I}_a(t) = 2\sqrt{2}V \sin \frac{\Psi}{2} \left\{ \frac{\sin(\omega_1 t - \frac{\Psi}{2} + \frac{\pi}{2})}{\sqrt{R^2 + \omega_x^2 L^2}} - \sin[(\omega_1 - \omega_x)t - \frac{\Psi}{2} + \frac{\pi}{2}] \frac{e^{-\frac{R}{L}t}}{\sqrt{R^2 + \omega_x^2 L^2}} \right\} \quad (4-3-5)$$

As this is a three-phase symmetrical system, the phases B and C must have similar transient response as the phase A.

Equation (4-3-5) is consisted of two parts: the first represents the steady-state current response under the instantaneous voltage phase shift, and the later corresponds to the transient current component. From the above analysis of the simplified induction motor model, it is noticed that the inductance “L” of the motor under the steady state and the transient process is different. Therefore, if the load is a induction motor, (4-3-5) should be changed to:

$$\Delta \bar{I}_a(t) = 2\sqrt{2}V \sin \frac{\Psi}{2} \left\{ \frac{\sin(\omega_1 t - \frac{\Psi}{2} + \frac{\pi}{2})}{\sqrt{R^2 + \omega_x^2 L_{ss}^2}} - \sin[(\omega_1 - \omega_x)t - \frac{\Psi}{2} + \frac{\pi}{2}] \frac{e^{-\frac{R}{L_{tr}}t}}{\sqrt{R^2 + \omega_x^2 L_{tr}^2}} \right\} \quad (4-3-6)$$

where L_{ss} , L_{tr} are given in (4-2-12) and (4-2-15) respectively

Next consider the problem of how to estimate the peak value of the current. As the time interval from the voltage phase shift to the peak value of stator current appears is very short, therefore the rotor speed ω_x will be very close to ω_1 . It is reasonable that to assume the term “ $\sin((\omega_1 - \omega_x)t - \frac{\Psi}{2} + \frac{\pi}{2})$ ” in (4.3.6) as constant.

Then, (4-3-6) can be rewritten as:

$$\Delta \bar{I}_a(t) = K \left(\frac{\sin(\omega_1 t - \frac{\Psi}{2} + \frac{\pi}{2})}{\sqrt{R^2 + \omega_x^2 L_{ss}^2}} - K_1 \frac{e^{-\frac{R}{L_{tr}}t}}{\sqrt{R^2 + \omega_x^2 L_{tr}^2}} \right) \quad (4-3-7)$$

where the parameters $K = 2\sqrt{2}V \sin(\frac{\Psi}{2})$ and $K_1 = \sin(\frac{\pi}{2} - \frac{\Psi}{2}) = \cos(\frac{\Psi}{2})$, $(\omega_1 - \omega_x)t \approx 0$

The first term in (4-3-7) will be at its maximum value at time t when $\left| \sin(\omega_1 t - \frac{\Psi}{2} + \frac{\pi}{2}) \right| = 1$. Similarly, the 2nd term will reach its maximum value where

$e^{-\frac{R}{L_{tr}}t} = 1$. Therefore, the largest current peak value that can occur is the sum of the two maximum values of the two terms in (4-3-7). That is:

$$|\Delta \bar{I}_{a-\text{peak}}| \approx \left| \frac{K}{\sqrt{R^2 + \omega_x^2 L_{ss}^2}} \right| + \left| \frac{KK_1}{\sqrt{R^2 + \omega_x^2 L_{tr}^2}} \right| \quad (4-3-8)$$

If $R \ll \omega_x L_{ss}$ and $R \ll \omega_x L_{tr}$, then the above equation could also be simplified as:

$$|\Delta \bar{I}_{a-\text{peak}}| \approx \left| \frac{K}{\omega_x L_{ss}} \right| + \left| \frac{KK_1}{\omega_x L_{tr}} \right| \quad (4-3-9)$$

Generally, one can utilize (4-3-8) or (4-3-9) to estimate the peak value of the transient current. The above equation also show that for a given phase shift ψ , the peak value of dynamic current will be decided by the mechanical speed ω_x and the inductances of the motor. Since the term $\left| \frac{KK_1}{\omega_x L_{tr}} \right|$ will dominate in the above equation (for generally $\frac{\psi}{2}$ is less than 30° and therefore, $K_1 = \cos(\frac{\psi}{2})$ is close to 1, and $L_{ss} \gg L_{tr}$), it means that motors with smaller stator and rotor inductances will produce the larger peak value in the stator current.

Furthermore, the decay time of the transient current component will only be governed by the term " $e^{-\frac{R}{L_{tr}}t}$ ". From (4-2-17), it is clear that the decay time of the stator current will be determined by R_s, L_s, L_r , and is independent of the rotor resistance R_r and the mechanical load with which the motor is connected.

Also notice that the phase shift ψ would directly influent the magnitude of the current change. As the term KK_1 is proportional to $\sin \psi$, an increase of the phase shift ψ will cause a larger transient load response.

Comparing the static linear load and induction motor, the change of inductance in the induction motor means that the transient response of the induction motor will be more

drastic and complicated than that of the static linear load. Hence, the following analysis will only consider the induction motor load.

4.3.2 Parallel Loads Response

The more likely situation in industry is that many motor loads are connected in parallel downstream of the DVR. The impedance between the DVR and loads also may come into play. Figure 4-5 shows the equivalent circuit. The impedance between the restorer and loads is represent by a R-L circuit, and the loads are assumed to consist of two motors (denoted as motor1 and motor2). In the distribution system, it is reasonable to regard the power line impedance as a R-L circuit where R and L are regarded as the sum of power line and source resistance and inductances respectively.

Figure 4-5 Equivalent circuit for supply system with parallel loads model

Assume that the reference frame will rotate at the speed of ω_x , which could be considered as the average speed of the two motors, for the purpose of reducing the calculation error. $\vec{V}_o$ represents the stator voltage.

Note that

$$\Delta \vec{I} = \Delta \vec{I}_1 + \Delta \vec{I}_2$$

That is:

$$\frac{\Delta \vec{V} - \Delta \vec{V}_o}{R + Lp + j\omega_x L} = \Delta \vec{V}_o \left(\frac{1}{R_{m(1)} + L_{m(1)}p + j\omega_x L_{m(1)}} + \frac{1}{R_{m(2)} + L_{m(2)}p + j\omega_x L_{m(2)}} \right) \quad (4-3-10)$$

Where,

$R_{m(1)}, R_{m(2)}$: the stator resistances of motor1 and motor2 respectively

$L_{m(1)}$: the inductance of motor1 (see equation (4-2-17)).

$L_{m(2)}$: the inductance of motor2 (see equations (4-2-17)).

Note that $L_{m(1)}$ and $L_{m(2)}$ have been approximated by two inductance values under the steady state and the transient states.

Let:

$$R + Lp + j\omega_x L = Z$$

$$R_{m(1)} + L_{m(1)}p + j\omega_x L_{m(1)} = Z_1$$

$$R_{m(2)} + L_{m(2)}p + j\omega_x L_{m(2)} = Z_2$$

Then,

$$\Delta \vec{V}_o = \Delta \vec{V} \frac{Z_1 Z_2}{Z Z_1 + Z Z_2 + Z_1 Z_2}$$

Set $\Delta Z = Z_1 // Z_2 = \frac{Z_1 Z_2}{Z_1 + Z_2}$, then the last equation becomes:

$$\Delta \vec{V}_o = \Delta \vec{V} \frac{Z_1 Z_2}{(Z + \Delta Z)(Z_1 + Z_2)} \quad (4-3-11)$$

As $\vec{I}_1 Z_1 = \vec{I}_2 Z_2 = \vec{V}_o$, therefore:

$$\Delta \vec{I}_1 = \Delta \vec{V} \frac{Z_2}{(Z + \Delta Z)(Z_1 + Z_2)} \quad (4-3-12)$$

$$\Delta \vec{I}_2 = \Delta \vec{V} \frac{Z_1}{(Z + \Delta Z)(Z_1 + Z_2)} \quad (4-3-13)$$

$$\Delta \vec{I} = \Delta \vec{I}_1 + \Delta \vec{I}_2 = \Delta \vec{V} \frac{Z_1 + Z_2}{(Z + \Delta Z)(Z_1 + Z_2)} = \frac{\Delta \vec{V}}{Z + \Delta Z} \quad (4-3-14)$$

From the above equations it is clear that if the motor load parameter ΔZ is fixed, then as the source impedance Z increases, $\Delta \vec{I}_1, \Delta \vec{I}_2, \Delta \vec{I}$ would reduce. This means that as the electrical distance between the loads and the DVR becomes larger, the effect on the loads caused by the injected voltage phase shift will be smaller.

Now consider a special case when the DVR is connected next to the load terminals. Then $Z \ll \Delta Z$, and Z could be ignored. As the voltage step change occurs at the terminals of the motors' stator, this will lead to:

$$\Delta \vec{I} = \Delta \vec{I}_1 + \Delta \vec{I}_2 = \frac{\Delta \vec{V}}{R_{m(1)} + L_{m(1)}p + j\omega_x L_{m(1)}} + \frac{\Delta \vec{V}}{R_{m(2)} + L_{m(2)}p + j\omega_x L_{m(2)}} \quad (4-3-15)$$

The total current response is the sum of motor responses. In the previous analysis, it has assumed that the rotating speed ω_x of reference is equal to the motor rotor speed ω_m . In this case, assuming the two motors rotate at the speeds of $\omega_{m(1)}$ and $\omega_{m(2)}$ respectively.

Therefore substituting (4-3-5) into (4-3-15), the total line current change is:

$$\begin{aligned} \Delta \vec{I}_a(t) = 2\sqrt{2}V \sin \frac{\Psi}{2} \left\{ \sin(\omega_1 t - \frac{\Psi}{2} + \frac{\pi}{2}) \frac{1}{\sqrt{R_{m(1)}^2 + \omega_{m(1)}^2 L_{tr(1)}^2}} + \frac{1}{\sqrt{R_{m(2)}^2 + \omega_{m(2)}^2 L_{tr(2)}^2}} \right. \\ \left. - \sin[(\omega_1 - \omega_{m(1)})t - \frac{\Psi}{2} + \frac{\pi}{2}] \frac{e^{-R_{m(1)}/L_{tr(1)} t}}{\sqrt{R_{m(1)}^2 + \omega_{m(1)}^2 L_{tr(1)}^2}} \right. \\ \left. - \sin[(\omega_1 - \omega_{m(2)})t - \frac{\Psi}{2} + \frac{\pi}{2}] \frac{e^{-R_{m(2)}/L_{tr(2)} t}}{\sqrt{R_{m(2)}^2 + \omega_{m(2)}^2 L_{tr(2)}^2}} \right\} \quad (4-3-16) \end{aligned}$$

where, $L_{tr(1)}$ and $L_{tr(2)}$ denotes the approximate transient inductance of motor1 and motor2, $L_{ss(1)}$ and $L_{ss(2)}$ represents the approximate steady-state inductance of motor1 and motor2.

The above equation can also be extended TO the situation when “n” motors are connected in parallel. In practice, it is most likely that each machine will have an interposing impedance between it and the DVR, and the voltage phase shift occurs on the load-side of the DVR terminals, as shown in Figure 4-6. The phase shift in $\vec{V}$ is then denoted as $\Delta \vec{V}$.

Figure 4-6 Dynamic Voltage Restorer with multi-machine load

In this case, since the interposing impedance Z_i could be represented by a R-L circuit, it can be combined with the load impedance. The total change in the current is therefore given by:

$$\Delta \vec{I} = \sum_{i=1}^n \frac{\Delta \vec{V}}{(R_{(i)} + R_{m(i)}) + j\omega_{m(i)}(L_{(i)} + L_{m(i)})} \quad (4-3-17)$$

Where $R_{(i)}$ and $L_{(i)}$ are the resistance and the inductance of the i^{th} branch respectively, $R_{m(i)}$ and $L_{m(i)}$ are the stator resistance and the inductance of motor i . Note that $L_{m(i)}$ has been approximated by two inductance values under the steady state and the transient states.

One may obtain an estimate of the peak value of the dynamic current response based on practical conditions. In industry, the range of slip of different motors under rated load is between 0.01 to 0.05. This means that the rotating speeds of the motors under rated load conditions do not vary greatly. If the motors have similar parameters, especially if the ratio $\frac{R_s}{L_s + L_r}$ is almost the same, their respective transient currents will reach their peak

value almost simultaneously. That is to say, the peak of the total current could be the sum of these motors' peak values. This is also the most strenuous case. Therefore one could assume this largest possible value as the peak of the current when n motors are supplied through the DVR, that is:

$$\vec{I}_{a_peak}(t) \approx \sum_{i=1}^n \vec{I}_{a_peak(i)} \quad (4-3-18)$$

When Z_i is too small to be considered and if the phase angle shift and loads are fixed, the most drastic dynamic current response could be estimated. This will bring the largest impact on the DVR. A more detailed analysis of the phenomenon on the restorer would be given in the next Chapter.

4.3.3 Adjustable speed drive response

Figure 2-17 shows the configuration of a typical ac driver system. The first stage consists of a three-phase diode bridge rectifier followed by a dc-link capacitor and inductor. The LC section filters the incoming full-wave rectifier voltage to produce a low-ripple dc voltage. The inverter section converts the dc voltage back to a variable frequency and variable magnitude ac voltage using a pulse width modulated (PWM) control scheme.

Figure 4-7 shows a typical dc link voltage waveform under the steady state. When a pair of diodes is conducting, the output dc link voltage is at the highest level of the instantaneous line-to-line voltage. The voltage will not only charge the dc link capacitor, but also supply power to the loads. When there is no diode conduction, the capacitor will discharge its energy to the load, and the dc link voltage will reduce in an exponential manner.

Figure 4-7 DC link voltage waveform for three phase bridge rectifier

As was explained earlier, once a voltage sag occurs, the series compensator will restore the load voltage to its nominal level and the compensating process could be considered as

a step change in the voltage phase angle. The corresponding dc link voltage will also experience a drop in magnitude, but since the capacitor of DC-link is energy storage device, it could discharge the energy to maintain the DC link voltage stable. As the voltage phase shift is instantaneous, the output of ASD is usually almost not affected. This conclusion would be verified by simulation results in Section 4.4.

4.4 Illustrative Examples

The following examples will illustrate the results of the analysis of different types of load response under the step voltage phase shift. As stated early, the dynamic response of induction motor will be more drastic and complicated than that of the static linear load, Therefore only motor load responses are explained here. Case 1 to 4 show the motor response under the different load situations. Case 5 shows an ASD response under voltage phase shift. The loads are assumed to be connected downstream of the DVR. The frequency of power system supply is 50 HZ. The simulation software Matlab was used to obtain the simulation results.

Case 1

The parameters for the motor1 [78] used in the study are shown in Table 4-1.

The supply base voltage is 220V and the voltage disturbance is that due to the DVR injecting a series voltage at $t=3s$, which causes the phase angle of the load terminal voltage to change suddenly. The induction motor is operating at the rated load prior to the disturbance.

The responses of stator current (phase A) under the different phase angle shift (ψ) of 30° and 20° are shown in Figure 4-8. It is observed that the stator current exhibits drastic oscillations. As has been deduced from the analysis, current response magnitude will be proportional to $\sin \psi$: the extent of the dynamic oscillation under the 30° phase shift is seen to be larger than that under 20° phase shift.

Table 4-1: Parameters of the induction motor1 used in study

Frequency	$f = 50 \text{ Hz}$
No. of Poles	$P = 4$
Motor stator and rotor parameters	$R_s = 0.063 \text{ ohms}, R_r = 0.083 \text{ ohms}$ $L_s = L_r = 1.4 \text{ mH}$ $L_m = 29 \text{ mH}$
	$J/P = 0.06 \text{ kg m}^2$
Stator Voltage (phase)	$V = 127 \text{ V}$
Rated output Power	2.5 KW
Rated Torque	15.63 Nm
Rated mechanical angular velocity	156.3 rad/sec
Rated Stator Current	14.9 A

On the other hand, from (4-3-6), the decay time of the transient will be decided by the motor time constant $\frac{L_s + L_r}{R_s}$ and is independent of any other factors. In this case, from the R_s and L_s, L_r values given in Table 4.2, the value of the motor time constant is determined to be 0.04sec. Generally, the transient should be over some 3 to 5 times of the time constant, which is about 0.13-0.22 seconds in this example. Indeed, this conclusion can be seen from the above figures: despite different angle shifts, about 0.2 seconds after the phase shift occurs, the deviations from the steady-state phase A current in both situations diminish to negligible values.

(a)

(b)

Figure 4-8(a) Response of Motor1 current (Phase-A) under different voltage phase angle shift (a) $\psi = 30^\circ$, (b) $\psi = 20^\circ$

Based on (4-3-7), the estimated results for the peak value of dynamic current under the 30° and 20° phase shifts are 127A and 90A respectively. The simulation values using Matlab indicate 122A and 80A respectively. Considering the calculation is based on the simplified motor model and under the worse case condition (assuming the first part and the second part of equation (4-3-6) reach their maximums simultaneously), the estimated result is deemed acceptable.

Case 2

Next consider a different induction motor (motor2). The parameters for motor2 [78] used in the study are shown in Table 4-2.

Assume that the DVR is connected very close to the load, and then the impedance between the DVR and load is ignored. After injecting a compensated voltage at $t=3s$ at the terminals of the induction motor, the current response under the 30° and 20° phase shift are shown in Figure 4-9.

Table 4-2 Parameters of motor2

Frequency	$f = 50 \text{ Hz}$
No. of Poles	$P = 4$
Motor stator and rotor parameters	$R_s = 0.435 \text{ ohms}, R_r = 0.816 \text{ ohms}$ $L_s = L_r = 2 \text{ mH}$ $L_m = 69.31 \text{ mH}$
	$J/P = 0.089 \text{ kg m}^2$
Stator Voltage (phase)	$V = 127 \text{ V}$
Rated output Power	2 KW
Rated Torque	12 Nm
Rated mechanical angular velocity	151.6 rad/sec
Rated Stator Current	7.66 A

It is clear that similar to Case 1, the larger phase shift produces the more drastic oscillation in the stator current. And in this case, the motor time constant ($\frac{L_s + L_r}{R_s}$) is 0.0092 sec, which is much smaller than that in Case 1. Therefore, the decay time of the transient has been reduced greatly. Only about 0.05 seconds after the phase shift occurs, the current will recover to its steady state value.

The estimated peak values by using equation (4-3-7) are 75.4A and 51.6A. These are much larger than that obtained from the simulation because of the point of wave switching effect. Nevertheless, the transient current are at least 3-5 times that of the rated value.

Figure 4-9 Response of Motor2 current (Phase-A) under the different voltage phase angle shift (a) $\psi = 30^\circ$, (b) $\psi = 20^\circ$

Case 3:

In this case, motor1 and motor2 are parallel connected downstream of the DVR, with different impedances between the DVR and loads being examined. The equivalent circuit is showed in Figure 4-5. The system impedance is varied to simulate a change in the power line impedance and supply transformer connection between the source and loads. The voltage phase angle shift is 30° in each instance. The following figures show the simulation results.

As shown in the discussion in section (4.3.2), it is concluded that with the increase of the system impedance, the extent of the current oscillation will be reduced. This conclusion can be readily seen in the above figures.

Figure 4-10 Response of parallel loads with the different impedances between the loads and the DVR: Voltage phase shift $\psi = 30^\circ$, (a) System impedance $Z_s = 0.1 + j10\Omega$
(b) System impedance $Z_s = 0.2 + j20\Omega$

Case 4:

When the impedance between loads and DVR device is negligible, the most drastic current response could be obtained for a given phase shift. The following figures show the simulation results. It is clear that the two motors reach their peak values almost at the same time, thus producing the largest transient currents at the terminals. As the DVR is in series connection with the loads, similar current will also flow through the hardware of the DVR. The components of the DVR would need to endure such transients. In the next Chapter, a detailed analysis of this problem will be given.

Figure 4-11 Response of parallel loads with no impedances between the loads and the DVR: Voltage phase shift $\psi = 30^\circ$

Case 5:

Assuming a voltage phase shift occurs at the terminal of an adjustable speed drive. The following figures serve to illustrate the situation when an adjustable speed drive responds under the step voltage phase shift.

The parameters about ASD as follows:

Supply voltage = 450V, ASD: Capacitor of dc-link = 5 mF, inductance = 0.2 mH, switching frequency of inverter = 2000HZ. Induction Motor: Rated power = 3KVA, Rated voltage = 380V. The above parameters are taken from [81].

In this simulation, a voltage phase shift of 30° is assumed at $t = 0.2$ s. From Figure 4-12(a), notice that there is an oscillation in the dc link voltage because of the input voltage phase shift. However, this oscillation lasts for a very short time. About one cycle later, the dc link voltage recovers to its rated value. Figure 4-12(b) shows the output voltage of the adjustable speed drive: obviously, since the dc-link voltage oscillation is small and lasts for a very short time, there is almost no effect on the output voltage.

(a)

(b)

Figure 4-12 (a) DC-link voltage response under the instantaneous voltage phase shift
 (b) Output voltage response of adjustable speed drive under the instantaneous voltage phase shift

4.5 Experimental verification

In order to verify the analytical results obtained in the previous Section, an experimental set-up shown in Appendix B and represented by the schematic diagram of Figure 4-13 has been arranged. The experiment permits the response of an induction motor due to a voltage phase shift to be measured.

Figure 4-13 Schematic of the experiment set-up to measure induction motor response.

In the above figure, the controllable voltage source is used as the power supply of the motor. The source was programmed to initiate a specific phase shift at a pre-assigned time. The induction motor under test was installed downstream of the controllable voltage source. To reduce the possibility of the motor from being damaged during the test, the supply voltage was intentionally reduced to 40V, as compared to the motor rated voltage of 220 V. A circuit current protection was also installed between the source and the motor. A dynamometer, acting as a controlled electro-mechanical load device, was connected to the motor so that it would allow different rotor torque to be applied. The motor terminal voltage and current were monitored by the oscilloscope. It was also connected to the controlled voltage source by a trigger line. The oscilloscope would be able to capture the motor voltage and current from the pre-disturbance steady-state right through to when the phase shift has occurred. The parameters of the induction motor are given in Appendix B. Recognizing the effect of point-on-wave switching on the dynamic current magnitude, the phase angles were introduced at the same initial angle ψ for each set of tests. This was for the purpose of comparison.

Figure 4-14 shows the induction motor stator current (phase A) under two terminal voltage phase shifts and with the same load torque of 0.2 Nm.

(a)

(b)

Figure 4-14. Motor stator current (phase A) under (a) 30 ° and (b) 40° phase shift: load torque 0.2Nm, voltage scale: 50V/div, current scale: 2A/div.

The figure shows that subsequent to the step voltage phase shift, the transient current is some 2.5-3.5 times that of the steady-state value. An increase of the phase shift has resulted in a larger transient current.

Furthermore, based on the motor parameters, the time constant of the motor is $\tau = (L_s + L_r)/R_s = 0.007$ s. From Figure 4-14, one observes that the motor transient is essentially over within 1.5 cycles of the mains frequency, or a time interval of some 4 times of the motor τ . It shows that the transient interval is independent of the change in magnitude of the phase shift.

Figure 4-15 shows another set of the test results when the induction motor was under different load conditions when subjected to the same voltage phase shift. It shows that despite the two different load levels (load torque of 0.1 N.m and 0.2 N.m), the magnitude of the transient (i.e. peak-peak) current relative to its steady-state value has remained relatively unchanged. Furthermore, the transient process appears to last for about 1.5 cycles in both cases. These observations are in agreement with the conclusions drawn

from the theoretical analysis: the decay time of transient process is only determined by the motor stator resistance and inductance, rotor inductance but is independent of other factors.

(a)

(b)

Figure 4-15 Motor stator current (phase A) under (a) load torque 0.1N.m and (b) load torque 0.2N.m: $\psi=40^\circ$, voltage scale: 50V/div, current scale: 2A/div

4.6 Load Response Under Exponential Phase Shift

As illustrated in the last Section, the step-type phase shift could result in large oscillations of motor load currents. In order to alleviate this problem, one solution based on “an exponential phase shift method” will now be proposed. The detailed theoretical analysis will be given to show that the phase shift scheme can be realized to achieve energy saving while the load current oscillation can be reduced at the same time.

The instantaneous phase shift method forces the voltage changes ($\Delta \vec{V}$) from zero to its phasor value. The stator current change is given by (4-3-1). $\Delta \vec{V}$ can be resolved into its d-q components. The method requires $\Delta \vec{V}_d$ and $\Delta \vec{V}_q$ to be introduced in the manner depicted in Figure 4-16(a). The proposed exponential phase shift method, shown in Figure 4-16(b), allows the phase adjustment be introduced in an exponential manner for

the d-q components. In adopting this latter injection strategy, the purpose is to ensure that the perturbations in the load currents would be significantly reduced compared to that resulting from the instantaneous phase shift method.

In the manner shown in Figure 4-16(b), T is the time constant of the exponential voltage change ($\Delta \vec{V}$). Clearly $\Delta \vec{V}$ would have (essentially) reached its final value in some $3 \sim 5$ T when the energy-saving objective referred to in Section 2.4.2 or [25] would have been realized.

Figure 4-16: (a) Instantaneous phase shift, (b) Exponential phase shift method

Expressed in the s-domain, one can represent the voltage change under the instantaneous voltage injection and the exponential phase shift methods as $\Delta \vec{V}_{in}(s) = \frac{\Delta \vec{V}}{s}$, and

$\Delta \vec{V}_{pr}(s) = \frac{\Delta \vec{V}}{s(Ts + 1)}$ respectively. The results of Section 4.3 show that the expression for

the motor load current under the instantaneous voltage injection ($\Delta \vec{V}$) method is given by (4-3-5). The corresponding load current under the proposed exponential phase shift method is:

$$\Delta \vec{I}_{pr}(s) = \frac{\Delta \vec{V}}{s} \cdot \frac{1}{Ts + 1} \cdot \frac{1}{L(R/L + j\omega_x + s)} \quad (4-6-1)$$

Since $\Delta \vec{V}$ is a space vector and its magnitude could be expressed as $|\Delta \vec{V}|$, then the magnitudes of the load current under the instantaneous injection and exponential phase shift methods can be obtained:

$$\left| \Delta \vec{I}_{in}(s) \right| = \frac{|\Delta \vec{V}|}{s} \cdot \frac{1}{\sqrt{(R + Ls)^2 + (\omega_x L)^2}} \quad (4-6-2)$$

$$\left| \Delta \vec{I}_{pr}(s) \right| = \frac{|\Delta \vec{V}|}{s} \cdot \frac{1}{Ts + 1} \cdot \frac{1}{\sqrt{(R + Ls)^2 + (\omega_x L)^2}} \quad (4-6-3)$$

Comparing (4-6-2) and (4-6-3), it is clear that when $T=0$, $\left| \Delta \vec{I}_{in}(s) \right| = \left| \Delta \vec{I}_{pr}(s) \right|$, and by increasing the injection voltage time-constant T , the current transients $\left| \Delta \vec{I}_{pr}(s) \right|$ in the load and in the DVR will decrease. Thus by reducing the rate by which $\Delta \vec{V}$ is introduced, a simple way to reduce and smooth out the dynamics of the load current oscillation can be realized. On the other hand, as the voltage injection has to be introduced as rapidly as possible so as to maximize the amount of energy saved, T has to be kept as low a value as possible.

In view of the above, T is an important design parameter on the DVR control. One may observe from (4-6-3) that the load current dynamics include two transient oscillations: one is pertaining to the rate by which the load terminal voltage is adjusted and is governed by the time constant T , the other current component is decided by the load time constant τ described in Section 4.3 where $\tau = (L_s + L_r) / R_s$ for a motor load. Under the exponential phase shift method, the load terminal voltage will almost reach its steady-state in about $3T \sim 5T$. Similarly, the load transient process will essentially be over in about $3\tau \sim 5\tau$. From the results of Section 4.2, one also observes that the load time constant τ is independent of its loading level. Therefore, to limit the transient current to within an acceptable range and to ensure that the load voltage transient duration is similar to that of the load current, one possible way would be to select the exponential phase shift time constant T such that $T=\tau$. In one word, if the load is a static linear type, $T = \tau = L/R$, T could be selected to be less than 10 ms as τ is typically about 3 ms. Instantaneous phase shift method can therefore be acceptable for use in series voltage injection if the loads considered are of static type, as was considered in [25]. If the load is an induction motor, T could be selected to be equaled to the motor time constant τ so that, $T = \tau = (L_s + L_r) / R_s$.

In conclusion, for a given static or induction motor load when the load electromagnetic time constant can be readily evaluated, it is necessary to set T such that it can balance the desire for achieving acceptable load swings while minimizing the amount of the injected energy taken from the DVR. As part of system design, it will be possible to effect the energy-saving objective by selecting the appropriate value of T such that the above consideration can be satisfied.

Numerical examples:

The DVR-load system (Figure 3-1) under voltage phase-angle disturbance will now be studied using some numerical examples. Since the impact of the phase shift on the motor load is more pronounced than that on linear static load, only the induction motor load is considered. The motor is connected downstream of an ideal voltage source and the line impedance between the source and motor is assumed negligible. The simulation software MATLAB was used. The induction motor model used in the simulation is a more detailed one compared to that used in the analysis: the electrical circuit of the machine is represented by a fourth-order state-space model and the mechanical part by a second-order system. The motor under study has the same parametric values as given in Case 1, Section 4.3. The frequency of the power supply is still 50 Hz and the base voltage is 220 V.

With reference to Figure 3-1, the voltage disturbance is assumed to be that due to the DVR injecting $\vec{V}_{inj}$ which causes the phase angle of the load terminal voltage $\vec{V}$ to shift. The induction motor is operating at rated load prior to the disturbance. The phase shift θ , necessary for achieving energy saving, is assumed to be 30° .

Figure 4-17 shows the load terminal voltage when $\vec{V}_{inj}$ is introduced instantaneously and that when it is introduced in an exponential manner. Figure 4-18(a) and Figure 4-18(b) are the load currents under the two injection methods. The time constant T used in the exponential phase shift method is set equal to the load time constant τ of 0.04s.

It is clear that with the instantaneous voltage phase shift, the motor stator current exhibits large and prolonged oscillation, the peak value of the current reaches almost 7 times of its nominal value. Under the exponential phase shift, the voltage phase shift is adjusted gradually so that energy-saving can be achieved while the load current dynamic has also been reduced effectively, as expected. The peak value of the load current under this latter scheme has been reduced to about 2.5 times of the pre-disturbance level, a much smaller value compared to that obtained under the instantaneous injection method.

Figure 4-17 Motor load terminal voltage under the instantaneous and exponential phase shift methods

Figure 4-18: Motor load current (Phase A) under (a) the instantaneous phase shift method and (b) the exponential phase shift method

Notice that despite the different injection methods, the load current transients in both cases have essentially subsided after a similar time interval. This means that the exponential phase shift has not prolonged the load current transients.

From this example, it shows conclusively the effectiveness of the exponential phase shift method in alleviating the load current transient under the energy-saving restoration strategy. The method minimizes the occurrence of large transients that may cause potential damages to the power system.

4.7 Conclusions

In this Chapter, the effects of voltage injection on loads have been examined. From the above analysis and verified by the simulation and hardware experimental results, it can be concluded that static linear load and induction motor both exhibit dissimilar dynamic response under the step voltage phase shift ψ . For the same load level, the response of motor will be more drastic than that of static linear load because of the latter's larger inductance. The magnitude of the current transient is proportional to $\sin \psi$. Therefore, a greater phase shift change will produce more drastic oscillations.

For a single motor, the decay time of the dynamic response will be decided by $\frac{L_s + L_r}{R_s}$

which is independent of other factors. The peak value of transient current could be several times that of its rated value because of the large difference in the inductances under the steady state and transient state. This Chapter also offers an effective way on how to estimate the peak value of load current under the phase shift.

For parallel loads, the impedance between the source and loads could mitigate the effect of the step voltage phase shift. When the DVR device is installed very close to the loads, the response of loads under the step voltage will impact more on the DVR system for the motor transient currents might reach their peak value almost at the same time. A method to estimate the peak value has been given.

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For adjustable speed-drive system, simulation results also show that because the dc-link voltage will only experience a much-reduced disturbance under the phase shift, the effect on the output voltage of an ac speed drive could be relatively small compared to that of a motor.

Finally, in an attempt to reduce the load transients, the exponential voltage injection scheme is proposed. It is shown to be effective in reducing the magnitudes of the load transients. Selection of the injection time constant T is seen to be a compromise between the desire to reduce load current swings and in maximizing the benefit that can be obtained from the stored energy of the DVR, in enhancing the supply quality of the power system.

Chapter 5

Impact of load transient current on DVR including thermal effects

As stated in Chapter 4, phase shift of supply voltage might induce drastic response on its downstream loads. The suggested exponential voltage injection scheme could greatly mitigate the magnitudes of the load transients. However, in most cases, the step voltage injection is still adopted to realize the voltage injection due to its convenience and simplicity. Therefore, large load transient current will still be introduced into the power system. Since the DVR is series-connected to the load, it must also experience similar load transient current caused by the step voltage phase shift. The voltage source converter is the main component of the DVR. It consists of IGBT and freewheeling diodes. As there are finite resistances in IGBT and diodes [81], the large transient load current will cause a temperature rise in the switching devices. If the converter temperature exceeds its design limit, the DVR could be damaged [81, 82]. This Chapter will describe a technique on how to evaluate the temperature rise following the voltage shift. The calculation allows an assessment on whether the load transient current could endanger the DVR and helps the design of the restorer so as to alleviate the heating problem.

5.1 Pulse-Width Modulation

The main component in the DVR is the voltage source converter (VSC). Generally the Pulse-Width Modulation switching pattern is used to control the output voltage of VSC. The following section contains an explanation of the basic operating theory of PWM pattern [83-85]. Figure 5-1 shows the single-phase (phase a) schematics of the full-bridge PWM converter series-connected to the load:

In the figure, I_a represent the load transient current in phase “a” caused by the step voltage shift and I_L is the corresponding transient current in the DVR converter.

Figure 5-1 Single-Phase (phase A) Schematics of the Full-bridge PWM converter Within the DVR

In converter circuits, in order to produce a sinusoidal output voltage waveform at a desired frequency, a sinusoidal control signal at the desired frequency is compared with a triangular waveform, as shown in Figure 5-2. The triangular waveform V_{tri} in Figure 5-2, which is also called the carrier wave $V_{carrier}$, is at a switching frequency f_c . It establishes the frequency with which the converter switches are turned on/off. The control signal $V_{control}$ is used to control the modulation ratio m_a and has a frequency f_l . f_l is the desired fundamental frequency of the converter voltage output, and would be equal to the power supply frequency. The amplitude modulation ratio m_a is defined as $m_a = \frac{\hat{V}_{control}}{\hat{V}_{tri}}$, where $\hat{V}_{control}$ is the peak amplitude of the control signal. The amplitude $\hat{V}_{tri}$ of the triangular signal is generally kept constant.

Figure 5-2 Pulse-Width Modulation (PWM)

In the DVR, V_{control} corresponds to the injected voltage $\vec{V}_{\text{inj}}$, which would be obtained from the reference voltage $\vec{V}_{\text{ref}}$ of DVR control system and supply sag voltage $\vec{V}_s$, as governed by the equation: $\vec{V}_{\text{inj}} = \vec{V}_{\text{ref}} - \vec{V}_s$. $\vec{V}_{\text{ref}}$ is generated through one of the injection scheme described in Chapter 3. Note that in Figure 5-2, it assumed the initial phase angle

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of the injected voltage is zero, therefore sinusoidal curve of V_{control} starts from zero point. If there is phase angle of injected voltage, the curve should be moved towards right or left along the time axis.

In the single-phase converter showed in Figure 5-1, the switches should be controlled based on the following rules. Note that the load is assumed to be the highly inductive load such as an induction motor.

When $V_{\text{control}} < V_{\text{tri}}$, the IGBT G1A and G4A should be given a signal to be turned on. If at that time the load current I_L is also larger than zero, then G1A and G4A would be conducting. If $I_L < 0$, despite G1A and G4A are given a signal to be turned on, they will not conducted, for the direction of the current with the inductive load can not be changed instantaneously. Then $V_A = V_d$ and $V_B = 0$, the output voltage across the DVR converter terminal should be approximately equal to V_d .

When $V_{\text{control}} > V_{\text{tri}}$ and if the load current I_L is still larger than zero, G1A and G4A should be off and G2A and G3A should be given a signal to be turned on. However as $I_L > 0$ and the load current can not be changed instantaneously, G2A and G3A will not conduct at once. The load current will flow through the freewheeling diodes D2A and D3A. Then $V_A = 0$ and $V_B = V_d$, the output voltage should be equal to $-V_d$.

The situation that G2A and G3A is given a signal to be turned on and G1A and G4A is off is almost similar.

From the above analysis it can seen that: If $I_L > 0$, G1A, G4A and D2A, D3A will be conducting. Conversely if $I_L < 0$, G2A, G3A and D1A, D4A will be conducting.

Figures 5-3 shows the current conducting wave in IGBT and Diode under the sinusoidal pulse-width modulation [82].

Figure 5-3 Current Conducting Wave in IGBT and Diode Under Sinusoidal Pulse-Width Modulation

5.2 Power Loss in the Converter

As shown in Figure 5-3, the converter consists of IGBT and freewheeling diodes, and they will conduct alternatively. Therefore, the power loss in the converter should also include two parts: IGBT loss and Diode loss.

5.2.1 IGBT Loss

The energy loss in IGBT will consist of three parts: Conducting loss, Turn-on loss and Turn-off loss [86-93].

5.2.1.1 Conducting Loss of IGBT

When one IGBT is conducting, the conducting energy loss is given by:

$$W_{\text{con}-T}(i) = \int_{t1(i)}^{t2(i)} V_{\text{ce(sat)}} I_L(t) dt$$

$$\text{Or, } W_{\text{con}-T}(i) = \int_{t1(i)}^{t2(i)} (V_{\text{ce(TO)}} I_L(t) + r_{\text{int}} I_L^2(t)) dt \quad (5-2-1)$$

where $V_{\text{ce(sat)}} = V_{\text{ce(TO)}} + r_{\text{int}} I_L$;

$V_{\text{ce(sat)}}$: the saturation value of collector-emitter(on-state voltage drop of the active IGBT);

$V_{\text{ce(TO)}}$:the static collector-emitter threshold voltage;

r_{int} : the internal resistance of IGBT, which could be obtained from the datasheet;

I_L : the current through the IGBT, it should be equal to the load current in this case.

$t1(i)$ and $t2(i)$ represent the switching on and off time while the IGBT conducts for the i^{th} time interval.

Now, the key is how to obtain $t1(i)$ and $t2(i)$. One could adopt an approximate method to determine them [83]. In practical conditions, the midpoint of each output pulse in the PWM wave will not coincident with the midpoint of the triangular carrier wave. However, if assuming the midpoint of each output pulse be symmetrical with the midpoint of the relevant triangular wave, this would simplify the calculation greatly. The method is shown in Figure 5-4[83]:

Figure 5-4 Simplified Calculation to determine the Switching Time of IGBT and Diode

As shown in Figure 5-4, at the time $t_{peak(i)}$, construct a vertical line that intersects with the modulation wave at point D, and then draw a horizontal line through D that intersect with the triangular wave at point A and B. The interval of the time $t1(i)$ and $t2(i)$, which correspond to A and B respectively, could be thought as the conducting time of IGBT for the i^{th} time. From the above figure one could also see that the pulse width $t_p(i)$ is very close to its actual value.

Assuming the peak value of the triangular wave is equal to 1:

$$\frac{1 - m_a \sin \omega_1 t_{peak(i)}}{t_p(i)/2} = \frac{2}{T_c/2} \quad (5-2-2)$$

Where, $t_{peak(i)} = (4i - 3) \frac{T_c}{2}$, $i = 1, 2, 3, \dots$, $T_c = 1/f_c$

Therefore, from equation (5-2-2), the pulse width $t_p(i)$ can be expressed as:

$$t_p(i) = \frac{T_c}{2} [1 - m_a \sin \omega_1 t_{peak(i)}] \quad (5-2-3)$$

Then,

$$t1(i) = t_{peak(i)} - \frac{t_p(i)}{2} \quad (5-2-4)$$

$$t2(i) = t_{peak(i)} + \frac{t_p(i)}{2} \quad (5-2-5)$$

Similar approximation can be applied for the negative wave.

5.2.1.2 Turn-on loss of IGBT

The turn-on and turn-off losses of IGBT can not be ignored [81, 82]. Since the duration of turn-on process is very short, one could assume that the current through the converter is constant. The same assumption is available in the analysis for the turn-off process [81].

Figure 5-5 IGBT Turn-On for an Inductive Load

Figure 5-5 shows the turn-on response if the power-circuit load is inductive with a time constant long in comparison with the switching time of the IGBT. While the IGBT is off, the freewheeling diode provides the path for the load current, and, since the diode is conducting, the voltage across the load terminals is close to zero. Application of a gating signal V_G initiates the IGBT turn-on process. After the gate-emitter voltage V_{GS} reaches its threshold voltage $V_{GS(TH)}$, the current I_C through the IGBT begins to rise and the current in the diode falls.

Since the recovery of a diode from the on-state to the blocking state takes a finite time, so, for a short time after the current I_D in the diode drops to zero while the IGBT turns on, the diode is on with a positive voltage applied to the cathode terminal of the diode, This allows conduction of reverse current until the diode recovers its blocking capability and this is reflected as a spike of current in the IGBT, as shown in Figure 5-5. The extra loss

in the switch due to the current spike can not be determined analytically, nor does a value appear in normal manufacturers' data sheets.

While the diode recovers its blocking state, the collector-emitter voltage across the IGBT begins to fall. And at that point, if by ignoring the current spike caused by the diode, one could consider the current across the IGBT has reached its steady state. Comparing the energy loss during the voltage fall time with it during the current rise time, the former is much smaller than the later. And as the fall time of voltage is not given in the datasheet, the energy loss during the voltage fall time could be ignored.

Generally, data sheets give typical delay time $t_{d(on)}$ and rise time t_r . By definition, the delay time $t_{d(on)}$ is the interval between the moment when the gate-emitter voltage V_{GE} has reached 10% of its end value, and the collector current I_C (that is also I_L) has increased to 10% of the load current, and the rise time t_r is from $0.1 I_L$ to $0.9 I_L$.

In summary, the turn-on energy loss is usually a measure of the dissipation in the IGBT during the interval from $0.1 V_{GE}$ to $0.9 I_L$. Therefore, an approximate calculation of the turn-on loss could be made:

$$W_{on-T} = \int_0^{t_{d(on)} + t_r} V_{CE} I_C dt$$

Where, $V_{GE} = V_d$, which is the dc link output voltage of the DVR;

Now set $t' = t - t_{d(on)}$. Since $I_C = \frac{I_L}{t_{d(on)} + t_r} t'$, where $I_L = I_L(tl(i))$, which is the load current value at the time of $tl(i)$. Then, the above equation could be rewritten as:

$$W_{on-T} = \int_0^{t_{d(on)} + t_r} V_d \frac{I_L}{t_{d(on)} + t_r} t' dt' \quad (5-2-6)$$

5.2.1.3 Turn-off Loss of IGBT

Figure 5-6 shows the turn-off response with the highly inductive loads. After sudden turn-off of the positive control, the gate-source voltage starts to decline based on the input capacitance of the IGBT and the gate resistance. IGBT will be changed until the gate-

source voltage drops to V'_{GS} , which maintains the operation between the saturated and active regions and is equal to 90% of the gate source voltage. Then the collector-emitter voltage V_{CE} of the IGBT begins to rise. The current I_C through the IGBT cannot drop considerably at that time, since the free-wheeling diode is poled in reverse direction as long as V_d is higher than V_{CE} and, therefore, is not able to take over load current I_L .

As soon as the collector-emitter voltage V_{CE} has exceeded the supply voltage V_d , the load current may commute to the free-wheeling diode, which is poled in forward direction at that time and the collector current I_C will drop.

Figure 5-6 IGBT Turn-Off for an Inductive Load

From the datasheets offered by the manufactures, one can check the turn-off delay time $t_{d(off)}$ and the fall time t_f . And $t_{d(off)}$ is defined as the time interval between the moment when the gate-emitter voltage V_{GE} has dropped to 90% of its turn-on value and the collector current has declined to 90% of the load current value. Finally t_f is defined as the time interval, where the collector current I_C drops from 90% to 10% of the load current I_L .

Then, from the above it can be seen that the energy loss due to the turn-off process would consist of two parts, as expressed below:

$$W_{\text{off-T}} = \int_0^{t_{d(\text{off})}} I_L \frac{V_{ds}}{t_{d(\text{off})}} t' dt' + \int_0^{t_f} V_d (I_{\text{load}} - \frac{I_{\text{load}}}{t_f} t'') dt'' \quad (5-2-7)$$

where $t' = t - t_{d(\text{off})}$, $t'' = t - (t_{d(\text{off})} + t_f)$, and $I_L = I_L(t_2(i))$, which is the load current value at the time $t_2(i)$.

5.2.1.4 Total power Loss of IGBT

Based on the above analysis, the total energy loss in the IGBT will be:

$$W_{\text{Tloss}}(i) = W_{\text{con-T}}(i) + W_{\text{on-T}}(i) + W_{\text{off-T}}(i) \quad (5-2-8)$$

and, the average power loss during the whole conducting process will be:

$$P_T(i) = \frac{W_{\text{Tloss}}(i)}{t_2(i) - t_1(i)} \quad (5-2-9)$$

where, $t_2(i) - t_1(i)$ is the conducting duration of IGBT for the i^{th} interval. They could be obtained from equation (5-2-4) and (5-2-5).

5.2.2 Diode Loss

Energy loss in the diode should include two parts: conducting loss and turn-off loss [86-93].

5.2.2.1 Conducting Loss of Diode

When one diode conducts, the conducting energy loss is:

$$W_{\text{con-D}}(i) = \int_{t_2(i)}^{t_1(i+1)} V_F I_L dt = \int_{t_2(i)}^{t_1(i+1)} (V_{FO} I_L + r_d I_L^2(t)) dt \quad (5-2-10)$$

where $V_F = V_{FO} + r_d I_L$, V_F is the forward on-state voltage (on-state voltage drop of the active diode); V_{FO} is the threshold voltage of the diode; r_d represents on-state resistance of the diode. The values of these parameters are usually provided for in the device's

datasheet. I_L is the current flow through the diode. It should be equal to the load current in this case. $t_{2(i)}$ and $t_{1(i+1)}$ represent the switching-on and switching-off times while the diodes are conducting for the i^{th} interval, as shown in Figure 5-4.

5.2.2.2 Turn-off Loss of Diode:

The waveform of the diode turn-off process is illustrated in the following figure:

Figure 5-7 Freewheeling Diode Turn-Off for an Inductive Load

From the above figure, the diode turn-off energy loss can be expressed as [81, 91]:

$$W_{off-D} = \int_{t_0}^{t_3} I_D V_D dt = \int_{t_0}^{t_2} I_D V_D dt + \int_{t_2}^{t_3} I_D V_D dt$$

During the period t_0 to t_2 , the diode forward voltage V_D changes little. Since generally the diode forward voltage is very small and could be negligible, therefore the last equation is approximated by:

$$W_{\text{off-D}} = \int_{t_2}^{t_3} I_D V_D dt \quad (5-2-11)$$

Then, the energy loss could be approximately calculated as:

$$W_{\text{off-D}} = \frac{1}{2} V_{\text{RRM}} I_{\text{RRM}} \left(\frac{di}{dt} \right)_{\text{diode}} \quad (5-2-12)$$

Where, $V_{\text{RRM}} = V_d + L_{\text{st}} \left(\frac{di}{dt} \right)_d$, L_{st} is the stray inductances of the converter, $\left(\frac{di}{dt} \right)_d$ is current gradient of diode. The last parameter depends mainly on how fast the diode turns off. Unfortunately, it has been recognized in the literature that it is difficult to determine the exact value of this parameter. It is therefore assumed that it is fixed and is independent of the load current. It also implies that V_{RRM} is a constant and is independent of the load current. I_{RRM} is the peak reverse recovery current. Generally it will depend on the load current and other parameters. However, from typical manufacturers' datasheets, it is observed that following the increase in the load current, I_{RRM} changes little. Furthermore, since during the transient process, the transient current should be much larger than the normal load current, it is reasonable to assume that I_{RRM} remains constant.

Thus during the transient process, when the transient current through the diode is much larger than its rated value, V_{RRM} , $\left(\frac{di}{dt} \right)_d$ and I_{RRM} could be assumed as constants. The diode turn-off energy loss can also be considered as a constant and is independent of the magnitude of the load current. This conclusion can be verified by referring to the data offered by the manufacturers [82, 94].

5.2.2.3 Total Power Loss of Diode

Based on the above analysis, the total energy loss in the diode will be:

$$W_{\text{Dloss}}(i) = W_{\text{con-D}}(i) + W_{\text{off-D}}(i) \quad (5-2-12)$$

Then the average power loss during the whole conducting process will be:

$$P_{\text{Dloss}}(i) = \frac{W_{\text{Dloss}}(i)}{t1(i+1) - t2(i)} \quad (5-2-13)$$

where, $t1(i+1) - t2(i)$ is the conducting duration of Diode for the i^{th} interval.

5.3 Temperature Rise In Converter

From the last section, it is seen that any current flows through the converter will cause electric energy to be converted to thermal energy losses. The consequence is for the temperature of the converter to rise. The heat potential due to forward, switching and blocking losses in the power modules has to be dissipated from the chip via the case and heat sink to the ambient [81].

In order to study the temperature rise within the module, it is convenient to make use of the thermal circuit model similar to that shown in Figure 5-8 [81, 82]:

Figure 5-8 Simplified Thermal Equivalent Circuit of A Power Module (Transient)

In the figure, r_{jc} , C_{jc} , Z_{jc} are the thermal resistor, thermal capacitor and thermal impedance between the junction and case respectively; r_{ch} , C_{ch} , Z_{ch} are the thermal resistor, thermal capacitive impedance and thermal impedance between the case and heatsink respectively; r_{ha} , C_{ha} , Z_{ha} are the thermal resistor, thermal capacitive impedance and thermal impedance between the heatsink and ambient respectively.

The thermal model of a power module will be considered to be linear. That is, it is considered to be a point source of power generation at the junction. This power, in the form of heat, flows from the junction, through the case and heat sink to ambient and creates a temperature gradient from the junction to ambient due to the equivalent thermal impedances.

Two conditions in this aspect have to be considered. One is the consideration of steady-state condition, where the converter has been conducting for a long time with the current either as a train of pulses or continuous. In this situation, from knowledge of the power

losses under the rated current condition that must be dissipated for a given device and ambient temperature, an appropriate heatsink could be selected [82].

The other is the consideration of transient current through the converter. In the case of short-duration and discontinuous operation, higher current values in the converter are possible compared to the values defined in the data sheets for the steady state operation. However, with this kind of operation, the heat capacity of the device should be considered. It must be ensured that the chip temperature does not exceed the maximum permissible limit (eg.150 °C) [88, 93]. Therefore, it is very necessary to deduce the temperature rise in the chip during the transient state caused by the voltage phase shift. This will be studied in the following section.

With refer to Figure 5-8 during the transient process, if one just consider the temperature rise in the chip, then the chip temperature as a function of the time t can be calculated by means of the power dissipation P and the transient thermal resistance between chip and case Z_{jc} [88, 92, 93].

$$T_j(t) = P(t) * Z_{jc}(t) + T_{case} \quad (5-3-1)$$

To simplify the calculation, it could assume the case temperature is constant during the transient interval. This is reasonable for the thermal response of case and heatsink is much slower than that of the chip [88, 92, 93].

5.3.1 Thermal Impedance

Since in the inverter, the IGBT and diode are mounted on the same basement together, they will conduct current alternatively. The values for Z_{jc} have to be entered separately for the IGBT and diodes.

For the purpose of this calculation the curves of the transient thermal resistance shown in the data sheets have to be transformed into an analytical function [92-93]. In other words, the equivalent thermal impedance between the junction and case would have distributed

components of thermal resistance and heat capacity. This leads to the following equations:

$$Z_{Tjc}(s) = \frac{r_{T1}}{1 + s\tau_{T1}} + \frac{r_{T2}}{1 + s\tau_{T2}} + \frac{r_{T3}}{1 + s\tau_{T3}} \dots \quad (5-3-2)$$

$$Z_{Djc}(s) = \frac{r_{D1}}{1 + s\tau_{D1}} + \frac{r_{D2}}{1 + s\tau_{D2}} + \frac{r_{D3}}{1 + s\tau_{D3}} \dots \quad (5-3-3)$$

Where r_{Tk} and r_{Dk} are the steady-state thermal resistances of the IGBT and diode respectively; τ_{Tk} and τ_{Dk} are the respective thermal time constants of the IGBT and diode. In this case, $\tau_k = r_k C_k$ where C_{Tk} and C_{Dk} are the steady-state thermal capacities of the IGBT and diode respectively. Usually device manufacturers can provide the parametric values of r_{Tk} , r_{Dk} and τ_{Tk} , τ_{Dk} . Hence C_{Tk} , C_{Dk} are readily known. Based on the above analysis, the thermal equivalent circuit of a power module is shown in Figure 5-9, typically for k of up to 5.

Figure 5-9 Thermal Equivalent Circuit of Power Module (Transient)

In the steady state, the equivalent capacitors in Figure 5-9 would be fully charged and would therefore not be considered in the analysis. The temperature gradient would be constant. The sums of the series-connected distributed thermal resistances r_{Tk} , r_{Dk} are the parametric values of the total IGBT and diode thermal resistances. These are denoted as r_{Tjc} , r_{Djc} and are available from the respective device datasheet. Hence, the steady state,

the temperature gradient would be fixed and the series sum of the distributed thermal resistances would be r_{Tjc} and r_{Djc} . They are the sum of the resistors respectively:

$$r_{Tjc} = \sum_{i=1}^n r_{Tk}, \quad r_{Djc} = \sum_{i=1}^n r_{Dk}$$

5.3.2 Thermal Effects due The Pulsed-Type Losses

As described in Section 5.1, the IGBTs and diodes conduct alternately. Therefore, the power loss in the converter can be approximated as a pulse-train, as shown in Figure 5-10. $P_{T(i)}$ and $P_{D(i)}$ are the IGBT and diode losses at the i^{th} interval, as given in (5-2-9) and (5-2-13)

Figure 5-10: Multiple Pulse of Power Loss in the Inverter

Now consider how to quantify the thermal effects of the pulsed type losses. Assuming one power pulse starts from time $t1$ and ends at $t2$. The effect can be calculated by considering it as the sum of a positive power pulse which starts from the time $t1$ and a negative power pulse which starts from the time $t2$, as plotted in Figure 5-11(a). The corresponding temperature rise is also shown in Figure 5-11(b). Under the effect of positive power, the temperature rise is ΔT_1 , and the temperature decrease caused by the negative power is shown as ΔT_2 . Then in the s-domain, the total temperature rise at time t should be expressed as:

$$\Delta T(t) = \Delta T_1(t) - \Delta T_2(t) = \frac{P}{s} e^{-t1s} Z_{jc}(s) - \frac{P}{s} e^{-t2s} Z_{jc}(s) \quad (5-3-4)$$

Where, Z_{jc} is the transient thermal resistance between chip junction and case.

Figure 5-11: Thermal Effects of the Single Pulsed Type Power

Then, in the s-domain, for the multiple power pulses in the inverter, one can write the following equation:

$$P(s) = P_T(s) + P_D(s)$$

$$P_T(s) = \frac{P_{T(1)}}{s} - \frac{P_{T(1)}}{s} e^{-t_2(1)s} + \frac{P_{T(2)}}{s} e^{-t_1(2)s} - \frac{P_{T(2)}}{s} e^{-t_2(2)s} \dots \quad (5-3-5)$$

$$P_D(s) = \frac{P_{D(1)}}{s} e^{-t_2(1)s} - \frac{P_{D(1)}}{s} e^{-t_1(2)s} + \frac{P_{D(2)}}{s} e^{-t_2(2)s} - \frac{P_{D(2)}}{s} e^{-t_1(3)s} \dots \quad (5-3-6)$$

where $e^{-t_1(i)s}$ and $e^{-t_2(i)s}$ are the time-shift operators. $i = 1, 2, 3 \dots$

5.3.3 Evaluation of Temperature Rise

Following the thermal equivalent block diagram (Figure 5-9), the characteristics of the chip temperatures of IGBT and diode versus time can be calculated according to the following equations:

$$T_{j_T}(s) = T_{case} + T_{coup/D} + P_T(s)Z_{Tjc}(s) \quad (5-3-7)$$

$$T_{j_D}(s) = T_{case} + T_{coup/T} + P_D(s)Z_{Djc}(s) \quad (5-3-8)$$

As the IGBTs and diodes are usually soldered onto a common copper plate in a power module, the elements $T_{coup/D}$ and $T_{coup/T}$ represent the thermal coupling of IGBTs and diodes with their corresponding anti-parallel elements. However, as these coupling effects only become significant at the supply frequency, therefore the effect of the coupling can be ignored in this calculation [82]. T_{case} is the temperature of the case that can be assumed to be constant during the transient interval. This is reasonable as the

thermal response of case and heatsink is much slower than that of the device [82][94]. With these points in mind, the temperature rise in the IGBTs and diodes can now be calculated using the following expressions:

$$\begin{aligned}\Delta T_{j_T}(s) &= P_T(s)Z_{Tjc}(s) \\ &= \left(\frac{P_{T1}}{s} - \frac{P_{T1}}{s}e^{-t2(1)s}\right)Z_{Tjc}(s) + \left(\frac{P_{T2}}{s}e^{-t1(2)s} - \frac{P_{T2}}{s}e^{-t2(2)s}\right)Z_{Tjc}(s) \dots\end{aligned}\quad (5-3-9)$$

$$\begin{aligned}\Delta T_{j_D}(s) &= P_D(s)Z_{Djc}(s) \\ &= \left(\frac{P_{D1}}{s}e^{-t2(1)s} - \frac{P_{D2}}{s}e^{-t1(2)s}\right)Z_{Djc}(s) + \left(\frac{P_{D1}}{s}e^{-t2(2)s} - \frac{P_{D2}}{s}e^{-t1(3)s}\right)Z_{Djc}(s) \dots\end{aligned}\quad (5-3-10)$$

After inverse-Laplace transformation one have,

$$\begin{aligned}\Delta T_{j_T}(t) &= \left(P_{T1}r_{T1}(1 - e^{-t/\tau_{T1}}) + P_{T1}r_{T2}(1 - e^{-t/\tau_{T2}}) + P_{T1}r_{T3}(1 - e^{-t/\tau_{T3}}) + \dots\right) \\ &\quad - \left(P_{T1}r_{T1}(1 - e^{-(t-t2(1))/\tau_{T1}}) + P_{T1}r_{T2}(1 - e^{-(t-t2(1))/\tau_{T2}}) + P_{T1}r_{T3}(1 - e^{-(t-t2(1))/\tau_{T3}}) + \dots\right) \\ &\quad + \left(P_{T2}r_{T1}(1 - e^{-(t-t1(2))/\tau_{T1}}) + P_{T2}r_{T2}(1 - e^{-(t-t1(2))/\tau_{T2}}) + P_{T2}r_{T3}(1 - e^{-(t-t1(2))/\tau_{T3}}) + \dots\right) \\ &\quad - \left(P_{T2}r_{T1}(1 - e^{-(t-t2(2))/\tau_{T1}}) + P_{T2}r_{T2}(1 - e^{-(t-t2(2))/\tau_{T2}}) + P_{T2}r_{T3}(1 - e^{-(t-t2(2))/\tau_{T3}}) + \dots\right) \\ &\quad \dots\end{aligned}\quad (5-3-11)$$

$$\begin{aligned}\Delta T_{j_D}(t) &= \left(P_{D1}r_{D1}(1 - e^{-(t-t2(1))/\tau_{D1}}) + P_{D1}r_{D2}(1 - e^{-(t-t2(1))/\tau_{D2}}) + P_{D1}r_{D3}(1 - e^{-(t-t2(1))/\tau_{D3}}) + \dots\right) \\ &\quad - \left(P_{D1}r_{D1}(1 - e^{-(t-t1(2))/\tau_{D1}}) + P_{D1}r_{D2}(1 - e^{-(t-t1(2))/\tau_{D2}}) + P_{D1}r_{D3}(1 - e^{-(t-t1(2))/\tau_{D3}}) + \dots\right) \\ &\quad + \left(P_{D2}r_{D1}(1 - e^{-(t-t2(2))/\tau_{D1}}) + P_{D2}r_{D2}(1 - e^{-(t-t2(2))/\tau_{D2}}) + P_{D2}r_{D3}(1 - e^{-(t-t2(2))/\tau_{D3}}) + \dots\right) \\ &\quad - \left(P_{D2}r_{D1}(1 - e^{-(t-t1(3))/\tau_{D1}}) + P_{D2}r_{D2}(1 - e^{-(t-t1(3))/\tau_{D2}}) + P_{D2}r_{D3}(1 - e^{-(t-t1(3))/\tau_{D3}}) + \dots\right) \\ &\quad \dots\end{aligned}\quad (5-3-12)$$

Since at the high switching frequency, the thermal coupling effect of IGBT and diodes could be ignored [88]. That also means IGBT and diodes will have different temperature rise. Therefore, the maximum temperature rise in the inverter should be given by:

$$\Delta T_j(t) = \max(\Delta T_{j_T}(t), \Delta T_{j_D}(t)) \quad (5-3-13)$$

The above analysis gives a detailed explanation on how to calculate the temperature rise in the DVR converter, and the whole process can be illustrated by a numerical example shown in the next Section.

5.4 Illustrative Examples

The following example illustrates the impact of motor load transient current on the DVR based on the above analysis. In this example, it is assumed that two groups of induction motors are connected electrically close to and downstream from a DVR. The first group of motors is consisted of 5 identical units, each rated 14.9 A, 15.93 Nm torque. The second group is also made up of 5 motors, each rated 7.78 A, 12 Nm torque. The parametric values of the induction motors are same with that in the simulation of Chapter 4, Case1 and Case 2. Using these values, one can readily verify that the two motor-group time-constants are 0.044 s and 0.009 s respectively. The IGBT power module selected for the DVR has the structure shown in Figure 5-1 and has the detailed data obtained from a manufacturer's datasheet [94]. This data is also given in the Appendix C.

Assume that a step voltage phase shift of 30° occurs at the terminals of the motors. Suppose the frequency of the carrier wave of the DVR is 5 kHz, and the amplitude modulation ratio m_a is 0.5. The voltage of the DC link is 300 V. It also means that the injected voltage (rms) is about 53 V ($= (V_d m_a) / (2\sqrt{2})$). Based on the analysis of Chapter 4, the transient process should be essentially over in about 5 times of the larger of the two motor time-constants, i.e. about 0.2 s.

Based on (4-3-6) and (4-3-18), the line current following the voltage phase shift can be readily evaluated. The power losses in the IGBT and diode are evaluated using the expressions derived in Section 5.3, the corresponding junction temperature rises of the IGBTs and diodes in the power module are calculated and shown in Figure 5-12 – Figure 5-15. In this example, the voltage phase shift is assumed to begin at $t = 0$ s. When this occurs, the PWM control of the DVR converter permits only IGBT T2 and diode D1 to conduct for the initial 0.06 s (approx) before T1 and D2 conduct.

Figure 5-12 Power loss and junction temperature rise in IGBT T1: ΔT_{j-T1}

Figure 5-13. Power loss and junction temperature rise in diode D2: ΔT_{j-D2}

Figure 5-14. Power loss and junction temperature rise in IGBT T2: ΔT_{j-T2}

Figure 5-15 Power loss and junction temperature rise in diode D1: ΔT_{j-D1}

The maximum (hot-spot) temperature rise in the device would be the maximum among the four temperature rises, i.e.,

$$\Delta T_j(t) = \max(\Delta T_{j-T1}(t), \Delta T_{j-D1}(t), \Delta T_{j-T2}(t), \Delta T_{j-D2}(t))$$

The “hot-spot” temperature rise is shown in Figure 5-16. It is seen that the highest device temperature rise is about 55°C . Generally, the case temperature is around 40°C - 60°C . Therefore, the maximum device junction temperature is up to 115°C . This is below the rated maximum temperature of 150°C for the power module, as given in the device datasheet. Hence the transient is unlikely to cause damage to the power module due to the thermal effect.

In this example, the normal load current (RMS) is only about 110 A for the 10 motors compared to the power module rated current of 400 A, and for a voltage phase shift of 30° . With greater load current, larger phase shift or other parametric changes, it is possible for the maximum device junction temperature to exceed the 150°C limit.

Figure 5-16. Maximum (hot-spot) Temperature Rise in the Power Module During the Transient Process: ΔT_j

5.5 Conclusions

As shown in previous Chapter, the transient load current due to voltage phase shift can be considerable. As the DVR is series connected to the load, large current would be induced

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in the restorer. By considering the power losses of the switching elements in a series-connected Custom Power device, a method to calculate the hot-spot temperature rise within the inverter has been proposed. From the analysis and verified by the simulation results, it is shown that the transient load currents can cause the device junction temperature to rise in a very short time. Hence the design of the Custom Power device would need to take into consideration the thermal effect on the inverter power module under the voltage phase shift condition.

Chapter 6

Dynamic Voltage Restorer in Mitigating Voltage Sag Due To Induction-Motor Starting

6.1 Introduction

In Chapter 2, it has been stated that voltage sags are mainly caused by the two disturbance events: short-circuit fault or the switchings of large loads such as the starting of large induction motors. The corresponding figures describing the situations are as shown in Figure 2-10 and Figure 2-11. Currently, most of research works on DVR applications are directed towards sags due to faults. The mitigation of voltage sag due to the switching on of large loads is seldom considered. In Section 2.3.1, a brief description on the characteristics of the sag due to motor-start has been given. In this Chapter, a more detailed study pertaining to this type of sag event will be described, and the compensation mechanism will also be considered. During start-up, an induction motor might take typically five to seven times larger current than normal. This will also cause a voltage depression on motor supply terminals, the magnitude of the voltage change will be determined by system parameters.

If the starting of induction motor occurs at the same position as the fault locations shown in Figure 2-10 and Figure 2-11, the DVR could easily restore the sensitive load terminal voltage to its nominal level since generally, the sag due to motor starting will be shallower than that due to a short-circuit fault.

However, the compensation process would be quite different if the induction motor is in parallel connection with sensitive loads, as Figure 6-1 shows. This is probably a situation commonly found in practice because induction motors represent the most extensive electrical power loads, and consume more than 60% of total load in the industrialized countries [2].

Figure 6-1 DVR-Motor starting model

In this case, the DVR is likely to be installed upstream of the PCC bus to protect the sensitive loads. Since the DVR is in series connection with the loads, it must experience the same motor-starting current during the compensation process. If the DVR is used to compensate for sag due to motor starting, one has to consider whether the maximum current during the motor start-up might endanger the DVR. Furthermore, the large starting current through the DVR also means that much more energy is required to be injected into the downstream loads. Depending on the sag magnitude and duration, this might deplete the energy storage device of the restorer.

With the DVR connected upstream of the motor, the large transient current could be damaging to switching devices in the DVR. An analysis for estimating the peak current and the required energy during motor-start is given. Based on the results of the analysis, suitable switching devices for the DVR can be selected. Illustrative examples verify that the analysis is effective, and the DVR can be relied on to maintain a good supply quality for sensitive loads during the motor-start.

6.2 Voltage sag due to motor starting

In the design of DVR, sag magnitude and duration would determine the DVR voltage rating and the energy capacity of the energy storage device. Therefore it is very necessary to study the characteristics of sag due to motor-start.

Induction motors draw up to 500% to 700% of its rated current during starting if it is a direct-start. This current remains high until the motor reaches its nominal speed. The depression in voltage also depends on the supply system impedance. Furthermore, the

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duration of this kind of sag depends on motor parameters; the complete process could take up to a few seconds for large motors.

Consider the system model shown in Figure 6-1, the voltage sag experienced by the sensitive loads is obtained from the voltage divider equation:

$$\vec{V}_{\text{sag}} = \frac{\vec{Z}_m}{\vec{Z}_s + \vec{Z}_m} \text{ p.u.} \quad (6-2-1)$$

where a source voltage of 1 p.u has been assumed. $\vec{Z}_s$ is the equivalent source impedance. $\vec{Z}_m$ is the motor impedance during its start-up and could be expressed as:

$$\vec{Z}_m(\sigma) = j\omega L_m // [R_s + \frac{R_r}{\sigma} + j(\omega L_s + \omega L_r)] \quad (6-2-2)$$

In arriving at (6-2-2), the corresponding equivalent circuit of the motor shown in Figure 6-2 has been assumed. It is a commonly used circuit to study a motor-start event, and it ignores the initial DC-offset component. This is because the DC component usually will last for a relatively short interval, compared to the whole starting process.

Figure 6-2 Commonly used equivalent circuit for induction motor

In this figure, $\vec{V}_s, \vec{I}_s$: the motor stator voltage and current; $\vec{I}_r, \vec{I}_m$ are the currents through the motor rotor and magnetizing field respectively, R_s, R_r : the stator and rotor resistances; L_s, L_r : the stator and rotor leakage inductances; L_m : the magnetizing inductance; σ is the motor slip.

Note that while the motor model shown in Figure 4-2 is suitable for transient analysis, the model shown in Figure 6-2 has the characteristic feature that it contains the slip-dependent resistance R_r / σ [95]. The motor impedance $\vec{Z}_m = \vec{V}_s / \vec{I}_s$ is thus a function

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of the slip. Therefore, the sag due to the motor starting produces voltage variation following the changes in the starting current.

From the point-of-view of the DVR design, the most severe sag event should be used. Obviously the deepest sag will be incurred by the largest possible starting current. Generally, the maximum starting current will occur at the initial time during the motor start-up. At that time, the motor slip σ is close to 1. Therefore, as an estimation, one could ignore the motor magnetizing inductance L_m and (6-2-2) becomes:

$$\vec{Z}_{\min} = R_s + R_r + j(\omega L_s + \omega L_r)$$

Substituting it into (6-2-1), the most severe sag during the motor start can be obtained:

$$\vec{V}_{\text{sag-min}} = \frac{\vec{Z}_{\min}}{\vec{Z}_s + \vec{Z}_{\min}} \quad (6-2-3)$$

$\vec{V}_{\text{sag-min}}$ would provide an estimate of the lowest voltage that can occur at the PCC due to the load switching event.

6.3 Restorer design considerations

Since the DVR is in series-connection with the motor load, the starting current will also flow through the DVR and this might exceed the current ratings of the power electronic switching components in the restorer. This is similar to the phenomenon described in Chapter 5. Therefore, it is necessary to estimate the peak current during the motor-start to ensure that the design of the DVR is adequate to deal with the large current. Furthermore, the starting will also draw energy from the DVR into the system. This might deplete the energy stored in the restorer. In order to guarantee that the DVR has enough energy during compensation, the energy required for the motor starting process must be estimated. A suitably sized energy storage device can then be incorporated.

6.3.1 Estimation of the peak starting current

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In this section, a time domain solution is derived to estimate the motor starting peak current [96, 97]. Since the peak current will appear at the initial period of the motor-start, the same assumption used in Section 2 will also be made use of herewith. Hence, set $\sigma = 1$. Ignoring the motor magnetizing inductance L_m , the per phase equivalent motor circuit in time domain is shown as Figure 6-3:

Figure 6-3 Per-phase equivalent motor circuit

In the figure, assuming at time $t=0$, a sinusoidal voltage is applied to the motor terminal, γ is the switching angle. The circuit equation is:

$$V_m \sin(\omega t + \gamma) = (R_s + R_r)I_s(t) + (L_s + L_r) \frac{dI_s(t)}{dt} \quad (6-3-1-1)$$

Set $R = R_s + R_r$, and $L = L_s + L_r$. By Laplace transformation, (6-3-1-1) can then be expressed in the s-domain:

$$I_s(s) = \frac{V_m (\sin \gamma \cdot s + \cos \gamma \cdot \omega)}{L(s + \frac{R}{L})(s + j\omega)(s - j\omega)} \quad (6-3-1-2)$$

Inverse Laplace of (6-3-1-2) and the time function of the stator current is:

$$I_s(t) = \frac{V_m}{\sqrt{R^2 + \omega^2 L^2}} \left\{ \sin\left(\arctg \frac{\omega L}{R} - \gamma\right) \cdot e^{-(R/L)t} + \sin\left[\omega t - \left(\arctg \frac{\omega L}{R} - \gamma\right)\right] \right\} \quad (6-3-1-3)$$

Detailed derivation is given in Appendix D.

From (6-3-1-3), it is clear that the peak current magnitude depends on the switching angle γ . The maximum current during the motor-start could be estimated in the following way: the first term on the RHS of (6-3-1-3) will be at its maximum when $e^{-(R/L)t} = 1$ and $\sin(\arctg \frac{\omega L}{R} - \gamma) = 1$. Similarly, the second term will be at its maximum value at time t

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when $\sin[\omega t - (\arctg \frac{\omega L}{R} - \gamma)] = 1$. Therefore, the largest current that can occur is the sum of the two maximum values of the two terms in (6-3-1-3). That is:

$$I_{\max} = \frac{2V_m}{\sqrt{(R_s + R_r)^2 + (\omega L_s + \omega L_r)^2}} \quad (6-3-1-4)$$

This is a conservative estimate but it would be appropriate to use it for the sizing of the restorer.

6.3.2 Estimation of required energy

The power used in starting the motor will include active power and reactive power. The reactive power exchanged between the DVR and the system is internally generated by the restorer. Active power exchange is between the restorer ac terminals and the restorer energy storage device. Therefore one must consider whether the energy storage device could provide sufficient active energy during the starting process.

From Figure 6-3, it is clear that the energy required for starting the motor can be calculated using the following equation:

$$E_a = \int_0^t 3I_s^2 (R_s + \frac{R_r}{\sigma}) dt \quad (6-3-2-1)$$

Notice that in (6-3-2-1), σ is a variable with time t . Also, the output electro-mechanical torque T_e of the motor is expressed as:

$$T_e \Omega_1 = 3I_s^2 \frac{R_r}{\sigma} \quad (6-3-2-2)$$

where Ω_1 is the synchronous mechanical speed of the motor, $\Omega_1 = \omega_1 / P$, P is the pairs of motor poles. ω_1 is the synchronous speed (corresponds to the power supply frequency).

The motor-load mechanical equation is:

$$T_e - T_L = J \frac{d\Omega_m}{dt} \quad (6-3-2-3)$$

where J is the inertia time constant of motor load, Ω_m is the mechanical speed of the motor, T_L is the load torque that the motor is connected. For ease of analyze, assuming the load torque is constant. Comparing to no-load starting, constant starting load requires

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more energy and is therefore a more stringent starting condition. Since $\Omega_m = \Omega_1(1 - \sigma)$, then (6-3-2-3) becomes:

$$T_e - T_L = J \frac{d\Omega_1(1 - \sigma)}{dt} = -J\Omega_1 \frac{d\sigma}{dt}$$

That is:

$$dt = -\frac{J\Omega_1}{T_e - T_L} d\sigma \quad (6-3-2-4)$$

Furthermore, from the assumption that $R_s \approx R_r$, T_e may be approximated as:

$$T_e = \frac{2T_m(1 + \sigma_m)}{\frac{\sigma}{\sigma_m} + \frac{\sigma_m}{\sigma} + 2\sigma_m} = \frac{2T_m(1 + \sigma_m)\sigma_m\sigma}{\sigma^2 + 2\sigma_m^2\sigma + \sigma_m^2} \quad (6-3-2-5)$$

where T_m is the maximum electro-magnetic torque, and σ_m is the slip when the maximum torque is reached [95]. Also from [95], the maximum torque and the corresponding slip are:

$$T_m = \frac{3PV^2}{4\pi f[R_s + \sqrt{R_s^2 + (\omega_1 L_s + \omega_1 L_r)^2}]} \quad (6-3-2-6)$$

$$\sigma_m = \frac{R_r}{\sqrt{R_s^2 + (\omega_1 L_s + \omega_1 L_r)^2}} \quad (6-3-2-7)$$

Note that (6-3-2-5), (6-3-2-6) and (6-3-2-7) are well-known expressions which can be found in references such as [95].

Substituting (6-3-2-2), (6-3-2-4) into (6-3-2-1):

$$E_a = J\Omega_1^2 \left\{ -\frac{R_s}{R_r} \int_1^{\sigma_r} \left(\frac{T_e}{T_e - T_L} \right) \sigma d\sigma - \int_1^{\sigma_r} \left(\frac{T_e}{T_e - T_L} \right) d\sigma \right\} \quad (6-3-2-8)$$

At $t=0$, $\sigma = 1$. Assume that the motor reaches its steady state when $\sigma = \sigma_r$. Following the derivation shown in Appendix E, the energy for the motor-start can then be calculated using the following equation:

$$E_a = J\Omega_1^2 \{M + N\} \quad (6-3-2-9)$$

where,

$$M = \frac{R_s}{R_r} A \{ (\sigma_r - 1) + A_1 \ln \frac{\sigma_r - \sigma_1}{1 - \sigma_1} + A_2 \ln \frac{\sigma_r - \sigma_2}{1 - \sigma_2} \},$$

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$$N = A[A_3 \ln(\sigma - \sigma_1) \Big|_1^{\sigma_r} + A_4 \ln(\sigma - \sigma_2) \Big|_1^{\sigma_r}].$$

All the constants A, A_i, σ_r and σ_i can be calculated readily, as shown in the Appendix.

The motor starting time can also be calculated:

$$t = \frac{J\Omega_1}{T_L} \{(\sigma_r - 1) + N\} \quad (6-3-2-10)$$

Details could be found in Appendix E.

6.3.3 Energy storage device capacity

The energy storage device is used to supply the necessary energy for the generation of the injected voltage. For most applications, the energy storage device can be an electrolytic capacitor bank. Thus based on the results of the previous Section, a suitable capacitor bank could be selected to cater for the sag compensation during motor-start.

From Figure 6-1, assume that the in-phase compensation technique described in Section 3.2 has been adopted for this application. Under this compensation scheme, the injected energy E_{inj} from DVR can be estimated:

$$E_{inj} = \left(\frac{|\vec{V}_{inj}|}{|\vec{V}_s|} \right) E \quad (6-3-3-1)$$

where $|\vec{V}_s|$ is the magnitude of the PCC voltage before motor-start, $|\vec{V}_{inj}|$ is the magnitude of DVR injected voltage. The most severe condition is when the voltage sag is at its lowest level, i.e.

$$|\vec{V}_{inj}| = |\vec{V}_s| - |\vec{V}_{sag-min}| \quad (6-3-3-2)$$

$|\vec{V}_{sag-min}|$ is as described by (6-2-3). E is the sum of motor starting energy E_a and the energy drawn by the parallel loads over the complete starting process. E can be calculated from the following equation:

$$E = P_s t + E_a \quad (6-3-3-3)$$

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In (6-3-3-3), P_s is the active power drawn by the parallel-loads, t is the motor starting time determined using (6-3-2-10). P_s can be assumed constant because the PCC voltage is maintained constant due to the compensation actions of the DVR.

Next consider the dynamics within the restorer. In the time domain, the injected voltage from the DVR during the motor-start can be expressed as:

$$V_{inj}(t) = V_d(t)m_a(t)\sin \omega_1 t = \sqrt{2}|\vec{V}_{inj}|\sin \omega_1 t \quad (6-3-3-4)$$

where $V_d(t)$ is the instantaneous dc-side voltage of the restorer, $m_a(t)$ is the modulation index of the PWM converter. Notice that the maximum value of $m_a(t)$ is 1. It means that if $V_d(t)$ is decreased to less than $\sqrt{2}|\vec{V}_{inj}|$, the injected voltage from the DVR will be too low as to be able to maintain the PCC bus voltage constant. Thus, there is a lower limit on the stored energy of the capacitor in order to ensure that this failure can be avoided. Let the capacitance of the storage device be C . The minimum value of C can be obtained from the following equation:

$$\frac{1}{2}CV_{dc}^2 - \frac{1}{2}CV_{min}^2 = E_{inj} \quad (6-3-3-5)$$

where, $V_{min} = \sqrt{2}|\vec{V}_{inj}|$. V_{dc} is the initial dc-side voltage of DVR. Substituting (6-3-3-1) into (6-3-3-5), one obtains:

$$C = \frac{2\left(\frac{|\vec{V}_{inj}|}{|\vec{V}_s|}\right)E}{V_{dc}^2 - 2|\vec{V}_{inj}|^2} \quad (6-3-3-6)$$

The value of V_{inj} is as given by (6-3-3-2). The above analysis provides a method on how to rate the storage capacity of the DVR in order to mitigate sag due to motor starting. The results are reasonable, as is illustrated by the numerical examples shown in the next Section.

6.4 Illustrative Examples

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The following examples will illustrate the analysis results shown earlier. The network configuration is similar to Figure 6-1, with the detailed parameters shown in Figure 6-4.

Figure 6-4 Simulation model

The induction motor under test is parallel connected with the sensitive loads. The DVR is installed upstream of the motor. In carrying out the study, the DVR controller with a feed-forward loop described in [77] is adopted. The DVR controller is designed so as to offer good dynamic response characteristics against supply and load disturbances.

The parameters of the motor are as follows. It is the same machine used in the hardware experiment of Chapter 4:

$$L_s = L_r = 0.01\text{H}, R_s = R_r = 2.803\Omega, L_m = 0.16\text{H}, V_{\text{phase-ground}} = 220\text{V}, f = 50\text{HZ}, \\ J = 0.006 \text{ kg.m}^2, P = 2, T_L = 12\text{N.m}$$

Next, consider the design of the DVR for mitigating voltage sag due to the starting of the motor. From (6-3-1-4), the possible maximum starting current for the motor is 67.4A. This means that the current ratings of the switching devices and other series-connected devices must be at least of this value.

Furthermore, based on the above parameters, from (6-3-2-9) and (6-3-2-10), the starting energy and starting time for the motor could be estimated:

$$E_a = 430.3\text{J}, t = 0.06\text{s};$$

The sum of motor starting energy and the energy consumed in the parallel loads calculated from (6-3-3-3) is

$$E = 3 * \left(\frac{220^2}{20} \right) * 0.06 + 430.3 = 865.9\text{J}$$

Also, the magnitude of sag due to motor starting, estimated using (6-2-3) is:

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$$|\vec{V}_{\text{sag-min}}| = 0.84 \text{ p.u.}$$

Then, the largest possible injected voltage from the DVR is:

$$|\vec{V}_{\text{inj}}| = 220 * (1 - 0.84) = 35 \text{ V}$$

The dc-link voltage V_{dc} is designed as 450V, which is the same as that used in the examples in Chapter 3. Thus from the above data and (6-3-3-6), the minimum required capacitance for the DVR energy storage device should be:

$$C = \frac{2 * (35 / 220) * 865.9}{450^2 - 2 * 35^2} = 0.0014 \text{ F}$$

Notice that this is only an estimated value. In order to guarantee that the DVR could provide enough energy to mitigate the sag, the capacitor could be selected slightly larger than the estimated value. Accordingly in the simulation to follow, a capacitance of 0.002F is adopted.

Next, the simulation results obtained will be shown. For the purpose of comparison, the simulation will be carried out under two conditions: (1) The induction motor will be started under no-DVR condition. There will be a voltage sag at the PCC-bus; (2) The induction motor will be started with the DVR in service. The PCC-bus voltage will be compensated to rated value because of the actions of the restorer. The results under no-DVR condition are shown in Figure 6-5, and Figure 6-6 shows the results with the DVR.

From Figure 6-5, it is clear that a voltage sag of about 0.16 p.u occurs at the PCC bus, the sag lasts for about 0.08s when the motor has reached its steady state. The maximum starting current is about 38A. Figure 6-5(c) shows the active power drawn by the motor. The energy can be calculated from the following equation:

$$E_a = \int_0^t P(t) dt$$

Using the results of Figure 6-5(c), E_a is obtained numerically and it is 160.05 J. Notice that this is only for a single-phase. Hence the total energy for the motor start should be: $3 * 160.05 \text{ J} = 480.2 \text{ J}$.

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Figure 6-5 Network response under motor-start without DVR: (a) PCC bus voltage; (b) Motor starting current; (c) Active power drawn by motor from the grid (single phase); (d) Mechanical speed of motor.

Figure 6-6(a) shows that because of the DVR compensation action, the PCC voltage is maintained at the rated value during the motor-start. This proves that the DVR is effective in mitigating the sag and the selection of the DVR energy storage device based on the previous analysis is suitable. The maximum motor peak current reaches 46A under this condition. The motor starting energy calculated from Figure 6-6(c) is about 457.1 J.

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Figure 6-6 Network response under motor-start with DVR: (a) PCC bus voltage; (b) Motor starting current; (c) Active power drawn by motor from the grid (single phase); (d) Mechanical speed of motor.

Comparing the with/without DVR conditions, one notes that:

1. Under no-DVR condition, voltage sag will appear at the PCC bus due to the starting of the motor. This also implies that the motor will be started under a lower than normal voltage. The result is that the starting process will be prolonged, but the magnitude of starting current could be reduced. Compare Figure 6- 5(b) and Figure 6-6(b): The maximum current is 38A, and the starting process lasts for about 0.1s under no-DVR condition (as Figure 6-5(b) shows). On the other hand, with the DVR, the maximum current is now 46A, which is higher than that under no-DVR condition. However, the motor-starting time is reduced, the motor reaches its steady-state in 0.06s (as Figure 6-6(b) shows). This duration is also quite close to the estimated motor starting time based on (6-3-2-10).

2. Despite under different starting condition, the required energy for the motor starting is almost the same. From Figure 6-5(c) and Figure 6-6(c), the amount of energy calculated are 480.2 J and 457.1 J respectively. This observation can be readily explained: from the point-of-view of energy balance, the total amount of mechanical energy drawn by the motor remains constant under both situations as the motor will need to rotate from rest to the same speed. Therefore the energy required should be almost the same. The difference would be due to the difference in internal power loss in the motor under different supply

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current. The energy calculated using (6-2-3-9) is 430.3 J, which is quite close to the values obtained in the simulation. The proposed method to calculate the energy is therefore reasonable.

3. Fig. 6(a) shows a perfect voltage waveform at the PCC bus. This illustrates that the DVR can provide enough energy for sag compensation due to the motor start. It also means the selection for DVR energy storage device is suitable.

6.5 Conclusions

Voltage sag due to the starting of motor has been studied. If the DVR is installed upstream of the motor, the large starting current will also flow through the DVR device and might damage DVR components. Furthermore, the large starting current will also require more energy to be drawn from the upstream system. In order to guarantee the restorer would offer a good supply quality for its downstream loads, a method to estimate the peak current and required energy for the motor starting process is presented. The proposed method can be adopted in the design of the DVR, with the view to provide the current rating of switching device and the energy storage device of the DVR.

Chapter 7

Conclusions and Recommendations

7.1 Conclusions

This thesis is concern with the application of series compensation technique for the purpose to provide customers with a good electrical supply quality. The likely causes of poor power quality considered are faults or load switchings. Compensation is achieved by injecting/absorbing energy from the external power system through the adjustment of a voltage component in series with the protected load terminal voltage. The specific series compensator considered in the thesis is called the Dynamic Voltage Restorer (DVR). It is a VSC-based Custom Power equipment.

Currently most of research works on DVR are focused on how to compensate for voltage sag, considered to be very harmful for customer devices. However, voltage swell, another kind of voltage disturbances, also exists in a power system and could damage equipment. The main difference between sag compensation and swell compensation through DVR is that under most of swell compensation, the DC-link voltage in the DVR energy storage device could rise. This is because the DVR would absorb active power from the external system. The research results shown in the thesis indicates that in order to avoid an excessive rise in the DC-link voltage, only the energy-saving and zero-power injection strategies could be used. The strategies are based on the adjustments of the injection voltage phase. Compensation of unbalanced voltage disturbances has also been analyzed. The analysis shows that it is only necessary to consider the positive phase sequence component of the voltages and the same injection strategy as that for balanced disturbance can be applied. Based on the analytical results obtained, a generalized compensation strategy for voltage sag/swell disturbances has been proposed in Chapter 3. With the proposed injection scheme, load terminal voltage can be restored for sag and swell while the DC-link voltage can also be controlled.

A voltage phase shift is inevitable and is associated with the generalized compensation scheme. The effect of the phase shift to the loads has been analyzed in Chapter 4. From the electrical circuit analysis, it is shown that the linear static load and the induction motor load will exhibit quite significantly different dynamic responses under the phase shift. One major factor which governs the responses is the phase shift magnitude. Furthermore, the response of induction motor to the phase shift will be more drastic than that on linear static load. This is because of the large difference in the motor inductance under the steady state and transient state. The peak value of the transient current in the motor could be several times of the rated value. For a single induction motor, the decay rate of the dynamic response will be dependent of the ratio of the motor stator reactance to rotor leakage inductance. It is independent of the loading condition. For parallel loads, the interposing impedance between the source and loads could mitigate the effect of the step voltage phase shift. The analysis in Chapter 4 also shows that if the DVR device is installed close to the loads, the response of the loads under the phase shift may cause all the motor transient currents to reach their peak values almost at the same time. The effect of phase shift on adjustable speed-drive system has also been studied. The results show that because of the buffering effect provided by the energy storage device in the drive system, the dc-link voltage will experience a much-reduced oscillation. Hence the impact of the phase shift on the output voltage of the speed drive is much less pronounced.

In order to alleviate the large oscillations of load currents due to step-type phase shift, one solution based on “an exponential phase shift method” has been proposed in Chapter 4. The results show that using such an injection strategy, the voltage phase shift is adjusted gradually to realize the energy-saving potential, while at the same time the load current dynamic can be reduced significantly compared to that obtained under the instantaneous injection method.

Since the large load transient currents will also flow through the series-connected DVR, the impact of this transient current on the DVR device has also been investigated. Chapter 5 provides a detailed analysis on how to assess the temperature rise in the power module under the voltage transient. By considering the power losses of the switching elements in a series-connected Custom Power device, a method to calculate the hot-spot temperature rise within the converter has been proposed. From the simulation results, it is shown that

the transient load currents can cause the device junction temperature to rise in a very short time. The power module within the DVR would be damaged if the temperature rise exceeds the device thermal limit. Therefore, the thermal impact of the transient load current on the series compensator must be considered.

It is also well-recognized that voltage sags are mainly caused by short-circuit faults and the switching in of large loads such as the starting of large motors. Presently, most of DVR compensation studies only consider sags due to faults. In Chapter 6, the mitigation of voltage sag due to motor starting is analyzed. This is for the case when the series compensator is installed upstream of the motor. The large starting current that flows through the DVR device may again induce large transient currents in the DVR switching components or deplete the stored energy in the energy storage device. To prevent this from happening, a method to estimate the peak current and the required energy for assisting with the motor-start is given. The role of the restorer is then to maintain the downstream load-bus voltage so that other sensitive loads would not be affected by the motor starting event. The proposed method can then be utilized by the restorer designers or power system planners, so as to guarantee that the DVR can effectively mitigate such kind of sag, without leading to any potential damage to the equipment.

7.2 Recommendations

In view of the progress described in the previous Chapters, the following directions are suggested as possible areas for further investigation.

In Chapter 3, the algorithms for the generalized voltage compensation strategy have been derived based on the assumption that the magnitude and phase angle of sag voltage remain constant during the whole process. Based on this, the control system of DVR can evaluate the required reference voltage for the generalized compensation strategy. In some cases, this assumption may not be valid. For example, if the sag is caused by the starting of a motor, the magnitude and phase angle of sag will change as the motor speed rises. Therefore it is difficult for the control system to find a fixed reference. Further study should be carried out on how to realize the suggested generalized compensation strategy under such case. Furthermore, the suggested generalized compensation strategy

assumes that there is sufficient range within which the DC-link voltage can vary. In practical design, there would be a limit placed on the magnitude of the DC-link voltage. If the DC-link voltage is too low to cater for the required injection voltage, the load-side voltage cannot be restored to its nominal level. The impact of this limit on the effectiveness of the restoration scheme needs to be investigated.

In Chapter 4, the study of the impact of phase shift on adjustable speed drive is based on simulation studies. A more detailed analysis including the drive electromagnetic dynamics and energy storage would be more conclusive. The proposed exponential phase shift injection technique described in Chapter 4 could be implemented on a proto-type DVR to study the complete restorer-drive system. It would form a useful platform to demonstrate the practical potential of the scheme.

Chapter 5 describes a technique by which the temperature-rise in the inverter due to voltage phase shift can be calculated. Unfortunately, the calculation process appears to be rather involved. One notices that during the transient process, it is only the maximum (hot-spot) temperature-rise within the inverter that is the main concern. Therefore, if one could simplify the calculation and find an effective way to directly estimate the hot-spot temperature-rise, this will be much more useful for the DVR designers.

Chapter 6 gives a detailed analysis on the mitigation of voltage sag due to motor starting. It assumes that only one single motor is connected to the downstream DVR. In practical situations, there may be many parallel-connected motors. Under such a case, the estimate of the peak load current and the design of the appropriate DVR will be a challenging task.

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APPENDICES

APPENDIX A

DC-link Voltage In The Restorer

This appendix provides an analysis of the relationship between the energy storage terminal voltage and the injected active power of the DVR.

Figure A-1 Simplified DVR model with its external system

In the above figure, $v_{inj}(t)$ is the DVR injected voltage and it can be expressed as

$$v_{inj}(t) = v_d(t) m_a \sin \omega_1 t = \sqrt{2} V_{inj} \sin \omega_1 t \quad (A-1)$$

where $v_d(t)$ is the energy storage voltage at time t and m_a is the modulation index of DVR inverter. $i_L(t)$ is the external load current. Suppose $i_L(t)$ lags $v_{inj}(t)$ by the angle β . Then it can be expressed as:

$$i_L(t) = \sqrt{2} I_L \sin(\omega_1 t - \beta) \quad (A-2)$$

Assuming energy stored in the inverter and filter could be ignored, then the VSC dc-side instantaneous power is equaled to the ac-side instantaneous power, that is:

$$v_d(t) \cdot i_d(t) = v_{inj}(t) \cdot i_L(t) \quad (A-3)$$

$i_d(t)$ is dc-side current and it can also be expressed as:

$$i_d(t) = C \frac{dv_d(t)}{dt} \quad (A-4)$$

C is capacitance of the dc link.

After time t , the injected (or absorbed) energy from the DVR to the external system will be:

$$\frac{1}{2} C V_{ds}^2 - \frac{1}{2} C v_d^2(t) = \int_0^t v_{inj}(t) \cdot i_L(t) dt \quad (A-5)$$

where V_{ds} is the initial value of $v_d(t)$.

Since

$$v_{inj}(t) \cdot i_L(t) = V_{inj} I_L \cos \beta - V_{inj} I_L \cos(2\omega_1 t - \beta) \quad (A-6)$$

Substitute (A-6) into (A-5), one obtains:

$$v_d(t) = \sqrt{\left(V_{ds}^2 + \frac{V_{inj} I_L}{\omega_1 C} \sin \beta \right) + \frac{V_{inj} I_L}{\omega_1 C} \sin(2\omega_1 t - \beta) - \frac{2 V_{inj} I_L \cos \beta}{C} t} \quad (A-7)$$

One notices that the fourth term on the RHS of (A-7) dominates the RHS of the equation as t increases. The third term is an oscillatory function and its contribution to $V_d(t)$ change is small. If $|\beta| < 90^\circ$, SC will inject active power into the system and the capacitor voltage will decrease. Conversely, if $90^\circ < |\beta| < 180^\circ$, SC will absorb active power from the system, the capacitor voltage will rise.

APPENDIX B

Experiment Configuration in Chapter 4

Motor parameters used in the experiment (Chapter 4):

The parametric values for the induction motor used in the experiment are as follows:

No. of poles (P) = 4, $R_s = 2.803\Omega$, $R_r = 2.803\Omega$, $L_s = L_r = 0.01H$, $L_m = 0.16H$, stator voltage (phase) = 220V. the rated power: 3KVA.

Due to hardware limitations, the applied voltage was applied at a much lower value than the rated value. In the experiment, the supply voltage (phase-ground) $V = 40V$ (RMS). With load torque = 0.1 N.m, stator current $I = 0.612A$, or peak current $I_p = 0.865A$. Mechanical speed $n = 1481$ rpm; With load torque = 0.2 N.m, stator current $I = 0.764A$, peak current $I_p = 1.08A$, mechanical speed $n = 1467$ rpm. The experiment system set-up is as shown in Figure B-1.

Figure B-1 Experiment set-up to measure motor current transients

APPENDIX C

Power Module Parameters Used in Chapter 5

SIEMENS IGBT Power Module BSM-400-GB-60-DN2 (Half-Bridge) is used in the example.

Maximum Ratings:

$$V_{CE} = 600 \text{ V}; I_C = 400 \text{ A}; T_C = 125^\circ \text{C}; T_{j\max} = 150^\circ \text{C}$$

Forward Characteristics:

$$V_{CE(\text{sat})} = 2.2 \text{ V}, T_j = 125^\circ \text{C}, V_{GE} = 15 \text{ V}.$$

$$V_F = 1.7 \text{ V}, T_j = 125^\circ \text{C}.$$

Switching Characteristics of IGBTs:

Inductive Load at $T_j = 125^\circ \text{C}$

$$t_{d(\text{on})} = 200 \text{ ns}, t_r = 190 \text{ ns}$$

$$(R_{Gon} = 4.7 \Omega, V_{GE} = 15 \text{ V}, V_{CC} = 300 \text{ V})$$

$$t_{d(\text{off})} = 680 \text{ ns}, t_f = 510 \text{ ns}$$

$$(R_{Gon} = 4.7 \Omega, V_{GE} = -15 \text{ V}, V_{CC} = 300 \text{ V})$$

Thermal impedances of module:

$$\text{IGBT: } R_{thjc} = 0.09 \text{ (k / w)}$$

$$\text{Diode: } R_{thjc} = 0.18 \text{ (k / w)}$$

k	IGBT		Diode	
	r_{Tk}	τ_{Tk}	r_{Dk}	τ_{Dk}
1	0.008653	2.047526	0.013709	2.640979
2	0.044893	0.064431	0.01236	0.445453
3	0.03566	0.012271	0.016866	0.169646
4	0	0	0.132465	0.022463
5	0	0	0.006449	0.000403

The circuit diagram of module should be:

Figure C-1 The Circuit Diagram of Power Module

Appendix D:

Calculation of the Motor Starting Peak Current

By inverse-Laplace transformation, (6-3-1-2) can be transformed into the time domain, that is:

$$\begin{aligned} I_s(t) &= A_1 e^{-(R/L)t} + A_2 e^{-j\omega t} + A_3 e^{j\omega t} \\ &= A_1 e^{-(R/L)t} + (A_2 + A_3) \cos \omega t - j(A_2 - A_3) \sin \omega t \end{aligned} \quad (D-1)$$

where,

$$A_1 = F(s) \cdot (s + R/L) \Big|_{s=-R/L} = \frac{V_m (-R/L \cdot \sin \gamma + \cos \gamma \cdot \omega)}{L(-R/L + j\omega)(-R/L - j\omega)} = \frac{V_m (\cos \gamma \cdot \omega - R/L \cdot \sin \gamma)}{L(R^2/L^2 + \omega^2)}$$

$$A_2 = F(s) \cdot (s + j\omega) \Big|_{s=-j\omega} = \frac{V_m (j\omega \cdot \sin \gamma - \cos \gamma \cdot \omega)}{L \cdot (-2j\omega) \cdot (R/L - j\omega)} = \frac{V_m (\cos \gamma - j \sin \gamma)}{L(-2jR/L - 2\omega)}$$

$$A_3 = F(s) \cdot (s - j\omega) \Big|_{s=j\omega} = \frac{V_m (j\omega \cdot \sin \gamma + \cos \gamma \cdot \omega)}{L \cdot (2j\omega) \cdot (R/L + j\omega)} = \frac{V_m (\cos \gamma + j \sin \gamma)}{L(2jR/L - 2\omega)}$$

$$A_2 + A_3 = \frac{V_m (\cos \gamma - j \sin \gamma)}{L(-2jR/L - 2\omega)} + \frac{V_m (\cos \gamma + j \sin \gamma)}{L(2jR/L - 2\omega)} = \frac{V_m (-\omega \cos \gamma + R/L \sin \gamma)}{L(R^2/L^2 + \omega^2)}$$

$$A_2 - A_3 = \frac{V_m (\cos \gamma - j \sin \gamma)}{L(-2jR/L - 2\omega)} - \frac{V_m (\cos \gamma + j \sin \gamma)}{L(2jR/L - 2\omega)} = \frac{V_m (j\omega \sin \gamma + jR/L \cos \gamma)}{L(R^2/L^2 + \omega^2)}$$

$$F(s) = L(s + R/L)(s + j\omega)(s - j\omega)$$

Substituting them into (D-1), one obtains:

$$\begin{aligned} I_s(t) &= \frac{V_m}{L(R^2/L^2 + \omega^2)} \{ (\omega \cos \gamma - R/L \sin \gamma) e^{-(R/L)t} + (-\omega \cos \gamma + R/L \sin \gamma) \cos \omega t \\ &\quad + (\omega \sin \gamma + R/L \cos \gamma) \sin \omega t \} \end{aligned} \quad (D-2)$$

Since: $(-\omega \cos \gamma + R/L \sin \gamma)^2 + (\omega \sin \gamma + R/L \cos \gamma)^2 = R^2/L^2 + \omega^2$,

(D-2) could be re-written as:

$$I_s(t) = \frac{V_m}{L\sqrt{R^2/L^2 + \omega^2}} \left\{ \sin\left(\arctg \frac{\omega L}{R} - \gamma\right) \cdot e^{-(R/L)t} + \sin\left[\arctg\left(\frac{-\omega \cos \gamma + R/L \sin \gamma}{\omega \sin \gamma + R/L \cos \gamma}\right) + \omega t\right] \right\} \quad (D-3)$$

Now set:

$$\frac{R/L}{\sqrt{R^2/L^2 + \omega^2}} = \cos \phi,$$

Then $\frac{\omega}{\sqrt{R^2/L^2 + \omega^2}} = \sin \phi,$

where $\phi = \arctg \frac{\omega L}{R}.$

Therefore,

$$\frac{-\omega \cos \gamma + R/L \sin \gamma}{\omega \sin \gamma + R/L \cos \gamma} = \frac{\sin(\gamma - \phi)}{\cos(\gamma - \phi)} = \tg(\gamma - \phi) \quad (D-4)$$

Substituting (D-4) into (D-1), one gets:

$$I_s(t) = \frac{V_m}{\sqrt{R^2 + \omega^2 L^2}} \left\{ \sin\left(\arctg \frac{\omega L}{R} - \gamma\right) \cdot e^{-(R/L)t} + \sin[\omega t - (\arctg \frac{\omega L}{R} - \gamma)] \right\}$$

which is equation (6-3-1-3).

Appendix E:

Calculation of the Energy for Motor Start-up

From (6-3-2-8), now set:

$$-\frac{R_s}{R_r} \int_1^{\sigma_r} \left(\frac{T_e}{T_e - T_L} \right) \sigma d\sigma = M \quad (E-1)$$

$$-\int_1^{\sigma_r} \left(\frac{T_e}{T_e - T_L} \right) d\sigma = N \quad (E-2)$$

Substituting (6-3-2-5) into (E-1), then:

$$M = \frac{R_s}{R_r} \int_1^{\sigma_r} \frac{2 \frac{T_m}{T_L} \sigma_m (1 + \sigma_m) \sigma^2}{\sigma^2 - [2 \frac{T_m}{T_L} \sigma_m (1 + \sigma_m) - 2\sigma_m^2] \sigma + \sigma_m^2} d\sigma \quad (E-3)$$

Set:

$$A = 2 \frac{T_m}{T_L} \sigma_m (1 + \sigma_m)$$

Then (E-3) can be further derived as:

$$\begin{aligned} M &= \frac{R_s}{R_r} A \int_1^{\sigma_r} \frac{[\sigma^2 - (A - 2\sigma_m^2)\sigma + \sigma_m^2] + (A - 2\sigma_m^2)\sigma - \sigma_m^2}{[\sigma^2 - (A - 2\sigma_m^2)\sigma + \sigma_m^2]} d\sigma \\ &= \frac{R_s}{R_r} A \left\{ (\sigma_r - 1) + \int_1^{\sigma_r} \frac{(A - 2\sigma_m^2)\sigma - \sigma_m^2}{(\sigma - \sigma_1)(\sigma - \sigma_2)} d\sigma \right\} \end{aligned} \quad (E-4)$$

$$\text{where, } \sigma_1 = \frac{(A - 2\sigma_m^2) + \sqrt{(A - 2\sigma_m^2)^2 - 4\sigma_m^2}}{2} \quad (E-5)$$

$$\sigma_2 = \frac{(A - 2\sigma_m^2) - \sqrt{(A - 2\sigma_m^2)^2 - 4\sigma_m^2}}{2} \quad (E-6)$$

Set:

$$M_1 = \int_1^{\sigma_r} \frac{(A - 2\sigma_m^2)\sigma - \sigma_m^2}{(\sigma - \sigma_1)(\sigma - \sigma_2)} d\sigma \quad (E-7)$$

(E-7) could be expressed as:

$$M_1 = \int_1^{\sigma_r} \left(\frac{A_1}{\sigma - \sigma_1} + \frac{A_2}{\sigma - \sigma_2} \right) d\sigma \quad (E-8)$$

where,

$$A_1 = \frac{(A - 2\sigma_m^2)\sigma - \sigma_m^2}{(\sigma - \sigma_1)(\sigma - \sigma_2)} (\sigma - \sigma_1) \Big|_{\sigma=\sigma_1} = \frac{(A - 2\sigma_m^2)\sigma_1 - \sigma_m^2}{\sigma_1 - \sigma_2} \quad (E-9)$$

$$A_2 = \frac{(A - 2\sigma_m^2)\sigma - \sigma_m^2}{(\sigma - \sigma_1)(\sigma - \sigma_2)} (\sigma - \sigma_2) \Big|_{\sigma=\sigma_2} = \frac{(A - 2\sigma_m^2)\sigma_2 - \sigma_m^2}{\sigma_2 - \sigma_1} \quad (E-10)$$

Therefore,

$$M_1 = A_1 \ln(\sigma - \sigma_1) \Big|_1^{\sigma_r} + A_2 \ln(\sigma - \sigma_2) \Big|_1^{\sigma_r} = A_1 \ln \frac{\sigma_r - \sigma_1}{1 - \sigma_1} + A_2 \ln \frac{\sigma_r - \sigma_2}{1 - \sigma_2}$$

And

$$M = \frac{R_s}{R_r} A \{ (\sigma_r - 1) + A_1 \ln \frac{\sigma_r - \sigma_1}{1 - \sigma_1} + A_2 \ln \frac{\sigma_r - \sigma_2}{1 - \sigma_2} \} \quad (E-11)$$

Again, substituting (6-2-3-5) into (E-2), then:

$$N = A \int_1^{\sigma_r} \frac{\sigma}{(\sigma - \sigma_1)(\sigma - \sigma_2)} d\sigma \quad (E-12)$$

(E-12) can be re-written as:

$$N = A \int_1^{\sigma_r} \left(\frac{A_3}{\sigma - \sigma_1} + \frac{A_4}{\sigma - \sigma_2} \right) d\sigma \quad (E-13)$$

where,

$$A_3 = \frac{\sigma}{(\sigma - \sigma_1)(\sigma - \sigma_2)} (\sigma - \sigma_1) \Big|_{\sigma=\sigma_1} = \frac{\sigma_1}{\sigma_1 - \sigma_2} \quad (E-14)$$

$$A_4 = \frac{\sigma}{(\sigma - \sigma_1)(\sigma - \sigma_2)} (\sigma - \sigma_2) \Big|_{\sigma=\sigma_2} = \frac{\sigma_2}{\sigma_2 - \sigma_1} \quad (E-15)$$

Therefore, (E-13) can be integrated to yield:

$$N = A [A_3 \ln(\sigma - \sigma_1) \Big|_1^{\sigma_r} + A_4 \ln(\sigma - \sigma_2) \Big|_1^{\sigma_r}]$$

So,

$$N = A [A_3 \ln \frac{\sigma_r - \sigma_1}{1 - \sigma_1} + A_4 \ln \frac{\sigma_r - \sigma_2}{1 - \sigma_2}] \quad (E-16)$$

Consequently, the energy required for the motor-start can be calculated as:

$$E_a = J\omega_1^2 \{M + N\} \quad (E-17)$$

The above equation could be further expressed as:

$$E_a = J\Omega_1^2 \left\{ A \left[\frac{R_s}{R_r} (\sigma_r - 1) + \left(\frac{R_s}{R_r} A_1 + A_3 \right) \ln \frac{\sigma_r - \sigma_1}{1 - \sigma_1} + \left(\frac{R_s}{R_r} A_2 + A_4 \right) \ln \frac{\sigma_r - \sigma_2}{1 - \sigma_2} \right] \right\} \quad (E-18)$$

Furthermore, from (6-3-2-4), the motor-start time can be expressed as:

$$t = -J\Omega_1 \int_1^{\sigma_r} \left(\frac{1}{T_e - T_L} \right) d\sigma \quad (E-19)$$

(E-19) can be further changed to:

$$t = J\Omega_1 \int_1^{\sigma_r} \frac{1}{T_L} \left(\frac{T_e - T_L - T_e}{T_e - T_L} \right) d\sigma = \frac{J\Omega_1}{T_L} \int_1^{\sigma_r} \left(1 + \frac{-T_e}{T_e - T_L} \right) d\sigma \quad (E-20)$$

Substituting (E-2) into (E-20), one get:

$$t = \frac{J\Omega_1}{T_L} \{ (\sigma_r - 1) + N \} \quad (E-21)$$

Additional Analysis:

The purpose of this addition analysis is to verify the existence of the solution for (E-18).

During the above derivations, one notices that to guarantee the existence of solutions of (E-18), the following equations must be satisfied:

$$\frac{\sigma_r - \sigma_1}{1 - \sigma_1} > 0, \text{ and } \frac{\sigma_r - \sigma_2}{1 - \sigma_2} > 0$$

In the previous analysis, σ_r is defined as the motor slip under a fixed load torque T_L .

σ_r can be calculated from equation (6-3-2-5):

$$T_e = \frac{2T_m(1 + \sigma_m)\sigma_r}{\sigma_r^2 + 2\sigma_m^2\sigma_r + \sigma_m^2}$$

Note that T_e is the electro-magnetic torque, which will be larger than T_L since

$$T_e = T_L + T_0$$

where T_0 is the motor no-load torque. It is typically about 3%-7% of the load torque T_L .

From (6-3-2-5), it is clear that for a given T_e , there will be two solutions (σ_{r1}, σ_{r2}).

However only in the range (0, σ_m), the induction motor can operate in the steady-state.

Therefore one will set $\sigma_r = \sigma_{r1}$ (then, $\sigma_r < \sigma_m$), as Figure E-1 shows.

Figure E-1 Induction Motor $T_e = f(\sigma)$ Curve (1)

σ_1 and σ_2 are obtained from equation (E-4), and they are the two roots of the equation:

$$\sigma^2 - (A - 2\sigma_m^2)\sigma + \sigma_m^2 = 0$$

Note that from equation (E-5) and (E-6), $\sigma_1 \geq \sigma_2$.

The above equation could also be expressed as:

$$T_L = \frac{2T_m(1 + \sigma_m)\sigma}{\sigma^2 + 2\sigma_m^2\sigma + \sigma_m^2} \quad (E-22)$$

(E-22) will have the same curve as (6-3-2-5), but since $T_e > T_L$, then $\sigma_r > \sigma_2$ will always exist. Furthermore, because $\sigma_m > \sigma_2$ and $1 > \sigma_m$, then $1 > \sigma_2$, as Figure E-2

shows. Therefore, $\frac{\sigma_r - \sigma_2}{1 - \sigma_2} > 0$

On the other hand, Figure E-2 also shows that when $\sigma = 1$, the corresponding electromagnetic torque is T_{st} , which is called the motor starting torque. Generally, from the viewpoint of motor design, the ratio of T_{st} / T_L will always lie in the range of (1, 2), that means $T_{st} > T_L$. Then from Figure E-2, it is clear that $\sigma_1 > 1$. Hence obviously, $\sigma_1 > \sigma_r$.

Consequently, $\frac{\sigma_r - \sigma_1}{1 - \sigma_1} > 0$.

Figure E-2 Induction Motor $T_e = f(\sigma)$ Curve (2)

Vita

Li Jianduo was born in P. R. China, in 1975. He received his B. Eng. degree from Tianjin University, China, in 1997 in Electrical Engineering. He then worked as an electrical engineer in power system from 1997 to 2001, in Tianjin Chengnan Power Supply Company. Since Aug 2001, he is reading for his Ph.D degree in Nanyang Technological University. His research areas include Power Quality and the use of Power Electronics devices for power system applications.

Research related to this dissertation has resulted in the following publications:

- [1] J. D. Li, S. S. Choi, D. M. Vilathgamuwa, "Impact of Load Transients on Dynamic Voltage Restorer", Proc. Int. Power System Technology Conference, PowerCon 2002, Vol. 1, Oct 2002, pp. 464 –470.
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