University of Alberta

A Self-Tuning Filter for the Mitigation of Power System Harmonics

by

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A thesis submitted to the Faculty of Graduate Studies and Research in partial fulfillment of the requirements for the degree of Master of Science

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Abstract

This thesis presents a novel self-tuning filter for the mitigation of power system harmonics. The device is based on the passive filter method with the inductive element being a variable reactor. A new type of variable reactor is introduced. The thyristor linked reactor (TLR) is a switch-able device that utilizes thyrsitors to add or remove turns of an inductor changing its inductance.

The self-tuning filter is presented within the frame work of harmonic mitigation as a method of power quality control. The device is studied in terms of steady state and transient switching response. Examples are presented to show where and how the novel self-tuning filter can be applied.

Acknowledgement

I would like to thank Dr. Wilsun Xu for his patience and persistence in reviewing this manuscript and directing my research. I would also like to thank the members of the committee Dr. Sirish L. Shah and Dr. Horacio J. Marquez for reviewing this work. I am in-debited to Emad E. Ahmed for his help with the case study found in Chapter 2.

On a personal note I would like to thank Julie for her loving support and seeing that I get this manuscript done. I would also like to thank the Bank of Mom and Dad for their generous donations and their repeated demonstrations of support.

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L – inductance

C – capacitance

X_L – inductive reactance at the fundamental

X_C – capacitive reactance at the fundemental

ΔX_L – change in inductive reactance at the fundamental

ΔX_C – change in capacitive reactance at the fundemental

TLR – abbreviation for thyristor linked reactor

R – resistance

Z – complex impedance

ω - fundamental frequency in rad/sec

X_{sc} – source (utility) reactance

h – harmonic order

V – voltage

I – current

V_h – voltage of harmonic frequency h

I_h – current of harmonic frequency h

ℓ - length of inductor

μ - magnitude unit micro- or permeabity

N – number of turns in an inductor

A – cross section area

R_S – resistive impedance of source (utility)

L_S – inductance of source (utility)

L_F – inductance of filter

R_F – resistance of filter

C_F – capacitance of filter

γ_h – pahse angle between voltage and current of some harmonic frequency

Chapter 1

Introduction

Electric power quality is an area of study that includes methods or devices designed to maintain a near sinusoidal voltage waveform, at the rated voltage magnitude and frequency for power systems. In this chapter the concepts of harmonic disturbance and the methods of harmonic disturbance mitigation are presented within the scope of power quality research. The chapter ends with a statement of the thesis objective and an outline of how the objective will be achieved.

1.1 Power Quality

Power Quality (PQ) refers to the characteristics of the power supply required for electrical equipment to operate correctly. Taking the lead of [1] a power quality problem is defined from the customer prospective as:

Any power problem manifested in voltage, current or frequency deviations that result in failure or mis-operation of customer equipment.

There are many different types and sources of power quality disturbances. It is common practice in industry for disturbances to be classified according to the electrical characteristics of the voltage experienced by the customer equipment. The characteristics include waveforms, RMS values and phase voltage balances. According to recent international efforts to standardize the definitions of power quality terms, common power quality disturbances are classified as [1]:

- Transients

  Transients are disturbances that have duration less than three cycles. These disturbances have many possible origins; examples of which are capacitor switching and lightning strikes. Transients can be further classified into impulsive and oscillatory types. Impulsive transients exist for less than 1 ms and are characterized by having sharp rise and fall edges. Oscillatory transients are generally less than a cycle in duration with a frequency content above 5kHz

- Short duration voltage variations

  These include voltage sags, swells and momentary interruptions. These disturbances cause abnormalities in the voltage periods of 0.5 cycles to 1 minute.

- Long duration voltage variations

  These are steady state over-voltages, under-voltages and sustained interruptions. These disturbances have duration in excess of 1 minute.

- Waveform distortion

  This type of disturbance is characterized by a steady state distortion of the 60 Hz sinusoidal waveform. Examples of the distortion are harmonics, voltage notching, DC offset and broadband noise.

- Voltage unbalance

  These are steady state disturbances where there is unbalance of the voltages among the three-phase power supply.

- Voltage fluctuations

  Repetitive sags and swells appear as voltage fluctuations. Voltage flickers is a well-known example of voltage fluctuations. The typical spectral content of voltage fluctuation is less than 25Hz.

- Power frequency variations
  This type of disturbance is defined as the deviation of the power system fundamental frequency from its specified nominal value. The typical duration of the frequency variation of interest is less than 10 seconds

The proliferation of power quality sensitive loads in recent years has made power quality one of the major concerns for utility companies, manufacturers and customers [2,3]. Interest in power quality was initially concentrated on the survey of power quality conditions and the improvement of monitoring techniques [4]. Such research has produced valuable information about the current power quality environment and has raised the general awareness on the subject. Consequently, the development and application of technologies for power quality disturbance mitigation have become new areas of power quality research[17]. Many innovative devices and techniques such as active harmonic filters as a means of harmonic mitigation are being developed and tested[5]. However, the progress in these areas is still far behind the expectations of industry.

1.2 Harmonic Disturbances

Harmonic distortions to the 60 Hz sinusoidal waveforms are among the most common type of waveform distortions [1]. Harmonics are present when the normally sinusoidal voltages and currents of a power system have a periodic distortion. Since these distorted waveforms are in a steady state condition they can be analyzed with Fourier analysis, to show their frequency content[17,26]. As a result of their periodicity, harmonic frequencies are integer multiples of the fundamental frequency; the first harmonic frequency is the fundamental, the second is twice the fundamental and so on. The harmonic content of a signal is defined in terms of the amplitude of each harmonic relative to the fundamental usually expressed as a percentage. Further if the waveform has symmetrical positive and negative going shape only odd harmonics will be present and if the waveform is non-symmetric both even and odd harmonics will be present.

Non-linear loads are those that cause harmonic voltage and current distortions in a power system[6]. Typical non-linear loads are the variable frequency drives and switched mode power supplies. The loads can be viewed as current sources that inject non-fundamental frequency currents back into the power system[21]. The injection of current results in the creation of voltage harmonics when the current passes through the utility impedance. Voltage and current harmonics can be detrimental to the proper function of other devices[7,8]. Examples include[9,10]:

1) In transformers harmonics currents can cause over-heating forcing the devices to be de-rated to prevent damage.
2) Many modern power electronics based devices require a clean supply current to determine proper firing angles. Harmonics may cause the device to function incorrectly or not at all.
3) The presence of any harmonic currents causes a lowering of the power factor, with equal power this means higher currents must be generated by the utility leading to an increase in line losses.
4) Capacitor bank failure
5) Inductive interference with telecommunication lines.
6) Errors in watt hour induction meters
7) Some devices are directly affected by the presence of voltage harmonics. For example capacitors are affected by the voltage harmonics because they induce harmonics currents to flow which may cause the device to exceed it's kVA rating.

Much research and effort has been directed towards the study of the origins, propagation and mitigation of harmonics within power systems. The aim of all this work is to contribute methods that will eliminate or reduce to an acceptable level the amount of harmonic content present in modern power systems.

1.3 Harmonic Mitigation

Ideally no harmonic currents should be present within a power system. The proliferation of non-linear loads has made it impossible for modern power systems to achieve this ideal. At present there are two general methods for reducing or eliminating harmonics from power systems. On one hand, harmonics can be prevented from being injected. On the other harmonics can be removed from a power system by placing harmonic filters near the current injection source. A review of present day literature and industrial practice highlights several methods.

1.3.1 Harmonics Cancellation Schemes

Utilities have little control on the type of loads connected to their systems. It is possible, however, for a utility to arrange connection schemes that can reduce the net harmonic current injection from the load into the system. An effective method is the transformer-based harmonic cancellation scheme. In this method, the transformer Y and Δ connections are employed to reduce harmonics in three-phase systems[11]. The effectiveness of the transformer cancellation schemes is dependent on the operating characteristics of the loads. If the harmonic currents produced by the loads are relatively constant and consistent over time, the scheme can be quite effective.

1.3.2 Passive Filtering

Passive harmonic filters can either shunt or block harmonic currents[1]. Shunt filters work by short-circuiting the harmonic currents injected by the loads. They keep the harmonic currents out of the supply system. The shunt filter is the most common type of filtering scheme used in industry because of economic consideration and benefits such as power factor correction[12]. Depending on the frequency response characteristics, there are different types of shunt filters. Some examples are the single-tuned filter, the 2^{nd} order filter and the high pass filter.

The series filter has the capability of blocking harmonic currents. It is a parallel-tuned circuit that offers high impedance to the harmonic current. It is not used often because it is difficult to insulate and load voltage is very often distorted. One common application is in the neutral of a grounded capacitor where it blocks the flow of zero sequence harmonics while retaining a good ground at the fundamental frequency.

The design of harmonic filters can be very difficult and are always system dependent. As a result no automated methods for harmonic filter design have been developed so far.

1.3.3 Active Filters

Active filters cancel voltage and current distortions by injecting signals that will compensate for signal distortions. The distorted signals can be analyzed in either the time or frequency domain. In the time domain the filter injects signal so as to reconstruct the desired sinusoidal signal. For the frequency domain the filter analyzes the frequency content then injects signals of equal magnitude but opposite phase so as to cancel the harmonics. Active filters inject signal through an inverter that allows either a dc current or voltage source to be disconnected, connected positively or connected negatively from the system[5].

Active filters in general have complex control systems for deciding how the inverter should be controlled. Because of the complexity they are generally expensive, difficult to maintain devcies. For these reasons active filters are at present an unattractive solution for industry harmonics mitigation.

1.3.4 De-tuning Distribution Circuits

One of the most significant problems caused by harmonics is resonance. Resonance can magnify harmonic distortions to a level that can damage the utility apparatus or cause malfunction to customer equipment. The possibility of resonance at harmonic frequencies can be prevented by de-tuning distribution circuits. De-tuning is achieved by changing or relocating power factor correction and filtering capacitors.

1.4 Thesis Objective

Harmonic filtering today is performed almost exclusively using static passive filters. Present active filter technology falls short of being implemented on a large scale due to the complexity of filter design[5]. In view of the increasing in harmonic content, the need for a more flexible filter that is easy to implement is obvious. To this end the development of variable reactors and self-tuning filter techniques is warranted.

The goal of this research is to investigate the feasibility and application of a novel self-tuning filter. The filter is based on a variable reactor called thyristor linked reactor. The scope of this thesis consists of the following:

1) Determination of the feasibility of the proposed self-tuning filter: Several different topologies for the self tuning-filter are presented. For each topology the required thyristor ratings will be calculated from steady state operation requirements. The self-tuning filter will be compared to present filters designs in terms functionality.
2) Investigation of the electromagnetic transients associated with the filter: Since the thyrsitor linked reactor has switching elements the transient response of the filter will be investigated. The level of transients created by the switching will be studied to show that they are not prohibitive. Elements

of the filter design that are critical to the limiting of transients created by switching are highlighted.

3) Development of control schemes for the filter. Three methods for tuning the novel filter are presented in order of increasing complexity. The thyristor linked reactor is controlled in one of two ways

a) The frequency response of the self-tuning filter meets design requirements.

b) The current through the filter is optimized to a programmed value.

The self-tuning filter researched in this thesis is a novel device. There are few experiences on its feasibility, application and control. The current work is to perform a preliminary investigation on the subject and to identify key areas that need further research.

1.5 Outline of Thesis

Present harmonic mitigation techniques fall short of industry expectations. The weaknesses of these methods must be understood so that a better method, the self-tuning harmonic filter, can be developed. Further options must be presented to respond to the demands of industry. On one hand the present methods for implementing harmonic mitigation must be improved so that uncontrollable changes in system variation can be managed. While on the other hand the complexity of the self-tuning filter must be kept at a minimum so as to make the usefulness of the design obvious and the implementation financially feasible. To this end this thesis is conducted as follows.

- Chapter 2 presents harmonic mitigation methods focusing on the passive and active harmonic filter technologies. Passive topologies are presented and a simple case study is preformed to show their advantages and short comings.
- The concept of a thyristor linked reactor (TLR) and its application to self-tuning harmonic filters is presented in chapter 3.

- Chapter 4 details the transient performance of the self-tuning harmonic filter. This chapter includes an analytical and a numerical study of the self-tuning filter switching.
- Several methods of control are outlined in chapter 5. These methods are presented in order of increasing complexity.
- Finally in chapter 6 conclusions about the self-tuning filter are drawn and possible steps for future research are suggested.

Chapter 2

Power System Harmonics and Mitigation

This chapter presents a review of the process of harmonic generation, harmonic load flow techniques and passive and active filtering methods. It is also demonstrated through a harmonic load flow case that passive filtering can be used to reduce harmonic propagation.

2.1 Harmonics in Power Systems

Many modern devices use power electronic switching equipment that draws current in a manner non-linearly related to the voltage across it. With these devices the waveforms of current are distorted but periodic with the voltage. The fact that the current waveform is periodic allows them to be analyzed with Fourier techniques. Using standard Fourier methods the periodic current waveform can be described as the sum of a series of harmonically related sine functions:

(2.1)

(2.2)

where

$|c_k|$- the magnitude of each harmonic component

$F_o=1/T_P$ – is the fundamental frequency

k – the harmonic order

θ_k= is the phase of the harmonic

If the current wave form has symmetric positive and negative going half periods , which is true for the devices to be studied, only odd harmonic will be present[14]. Further for practical considerations only harmonics up to the 50^{th} order are considered; at frequencies above this harmonic the contributions are generally negligible. A load drawing distorted current can also be viewed as a current source injecting harmonic currents[1]. In this way non-linear loads generate harmonics currents .

Two common harmonic sources are the switch mode power supply (SMPS) and the pulse width modulated adjustable speed drives (PWM ASD). Both draw on an ac source and use a diode bridge to create a signal that is used to feed a capacitor. This capacitor is regularly charged in short bursts to maintain a dc voltage bus. The SMPS is shown in Figure 2.1. It is a common power supply for PCs, battery chargers and electronic ballast. The SMPS draws current in short bursts every half cycle to charge to dc side capacitor. A typical current waveform is presented in Figure 2.2 and its harmonic content as a percentage of the fundamental in Figure 2.3. This type of power supply shows a distinctive high 3^{rd} harmonic content.

The PWM ASD shown in Figure 2.4 is a popular ac drive configuration. The PWM ASD rectifier circuit operates much like a three phase SMPS, it charges a capacitor at regular intervals to create the rectified signals. From the dc side bus a three phase motor is controlled using PWM techniques. The three phase power converter results in a harmonic current spectrum missing all even harmonics and those that are multiplies of three. An exemplary waveform and its spectral content are given in Figures 2.5 and 2.6.

General quantitative indices have been developed to judge the degree to which a current waveform is distorted. A single valued index for comparing the harmonic distortion of a particular loads is the total harmonic distortion (THD)

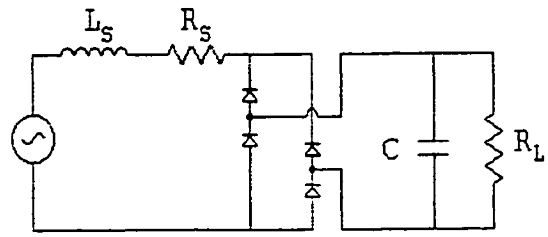

Figure 2.1 Switch Mode Power Supply

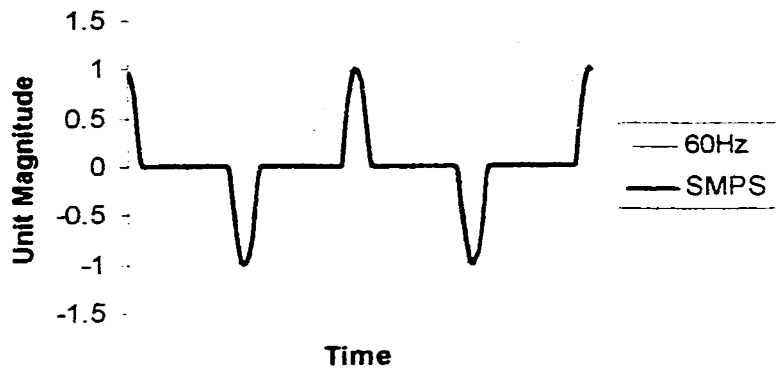

Figure 2.2 SMPS and 60Hz Current Waveform

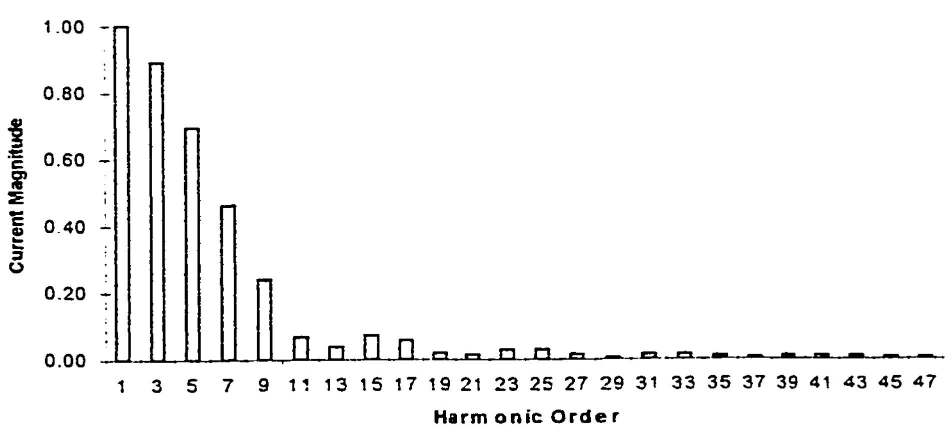

Figure 2.3 SMPS Harmonic Content

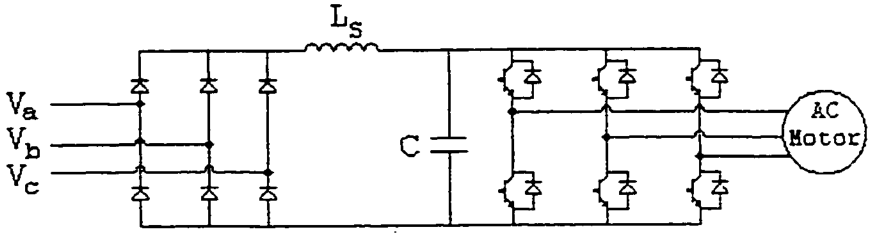

Figure 2.4 Pulse Width Modulated Adjustable Speed Drive

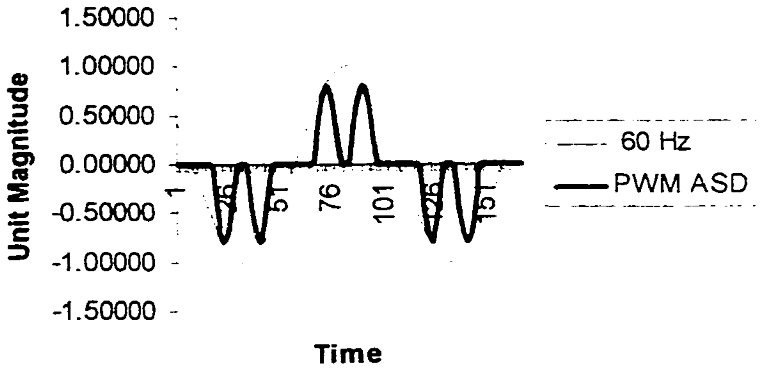

Figure 2.5 60Hz and PWM ASD Current Waveform

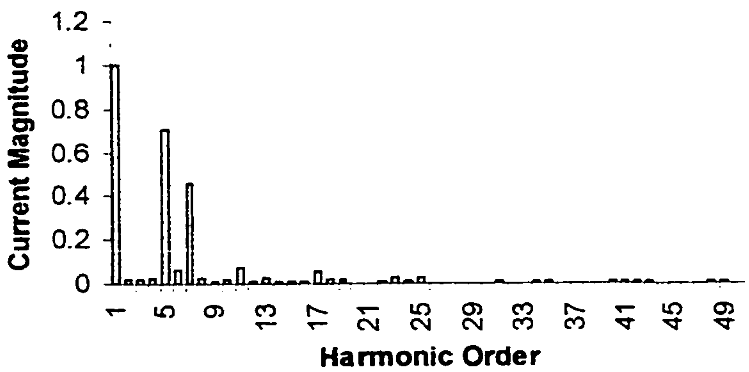

Figure 2.6 PWM ASD Harmonic Content

$$THD = \frac{\sqrt{\sum_{k=2}^{k_{MAX}} M_k^2}}{M_1} \tag{2.3}$$

where

M_1 is the RMS value of the fundamental

M_k is the rms value of the k^{th} harmonic

THD is a measure of the effective value of the harmonic content relative to the fundamental component. The harmonic current drawn by a non-linear load can be viewed as a harmonic current injection into the power system. These harmonic currents will propagate along the path of least resistance until they are returned to ground or dispersed resistively.

To determine the propagation of harmonics in a power network, traditional nodal Y matrix solution methods can be used[26]. The harmonic producing loads are represented using current sources with magnitudes and phases determined from their harmonic spectrum. The network admittance is described by nodal admittance matrix at the harmonic frequency of interest.

The standard node equations relating network voltage, current and network admittance for an n node network are given by

$$\begin{bmatrix} I_1 \\ \vdots \\ I_n \end{bmatrix}_h = \begin{bmatrix} Y_{11} & \cdots & Y_{1n} \\ \vdots & \ddots & \vdots \\ Y_{n1} & \cdots & Y_{nn} \end{bmatrix}_h \begin{bmatrix} V_1 \\ \vdots \\ V_n \end{bmatrix}_h \tag{2.4}$$

where

$V_1 \ldots V_n$ are the voltages at the network nodes

$I_1 \ldots I_n$ are the currents being injected into the network nodes

Y_{jk} are node admittance

h is the harmonic number

The admittance along the diagonal Y_{jj} are self-admittance and are equal to the sum of all admittance terminating on the node j identified in the repeating subscript. The other admittance are mutual admittance Y_{jk} which are equal to the negative of the sum of all admittance connected directly between the nodes j and k in the subscripts.

Solving the above equation will give the nodal voltages. The solution process is repeated for all harmonic frequencies of interest.

2.2 Passive Filters

One method of power system harmonic mitigation is passive filtering. A passive filter is composed of capacitive, inductive and resistive elements selected so that they either:

1) Present a high series impedance to the desired range of frequencies
2) Present a low shunt impedance to the desired range of frequencies

The series filter usually takes the form of suitably sized elements parallel-tuned to block the harmonics in question. The tuned circuit must be selected so that it can carry full load line current at the line voltage. The series filter must therefore be heavily insulted making the device expensive for all but low power application[27,21].

Passive shunt filters are the most utilized method of harmonic mitigation[15]. They are designed to present a low impedance path to ground for the harmonics to be filtered[28]. Three types of passive shunt filters are presented left-to right in Figure 2.7 in order of increasing complexity. The first-order high pass filter is used to significantly attenuate all non-fundamental currents. The first order filter has a frequency response given by:

$$|Z| = \left| R - \frac{j}{\omega C} \right| \qquad (2.5)$$

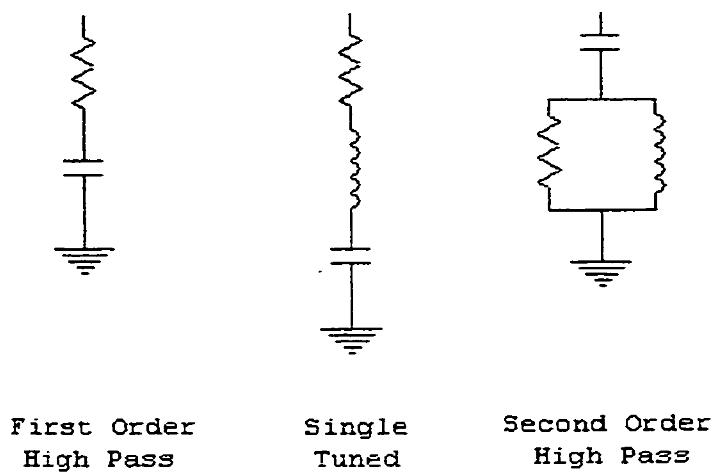

Figure 2.7 Passive Shunt Filters

The capacitance C must be selected so that acceptable attenuation is seen at the fundamental (Figure 2.8). Further the capacitor must be rated to carry all the harmonic current content. The first order filters present a simple solution to attenuate all harmonic frequencies but due to the large losses and high capacitor ratings its application is not always feasible.

Single-tuned or notch filters are designed to short circuit particular harmonic currents. A notch filters frequency response is given by

$$|Z| = \left| R - \frac{j}{\omega C} + j\omega L \right| \tag{2.6}$$

Figure 2.9 plots the frequency response of a notch filter tuned near the 5^{th} harmonic. At frequencies close to the harmonic frequency to be filtered, the term approaches resonance leaving only the resistive impedance. In other words the terms ωL and $1/\omega C$ become equal and cancel out. To control the amount of current flowing through the filter it is often tuned to a frequency just below the desired harmonic[1]. Single tuned filters are very effective at removing specific

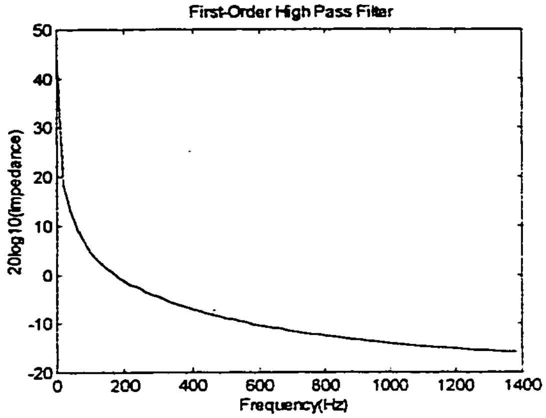

Figure 2.8 First Order High Pass Filter

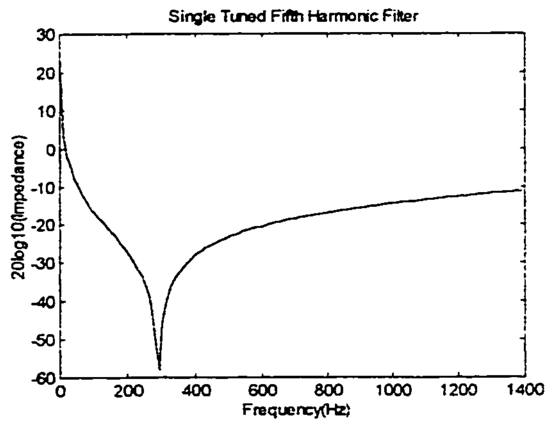

Figure 2.9 Single Tuned Filter

frequency currents with low losses. Their drawback is that due to their sharp tuning any deviations of inductance or capacitance *ie.* due to temperature or failure, will cause considerable changes in the amount of harmonic current filtered to ground. This problem is heightened when the filter interacts with the system impedance. Dramatic changes in impedance can occur if the frequency of parallel resonance between the utility and the filter is near the harmonic of interest[1,21,28]. Figure 2.10 shows the frequency response of the system of a single-tuned filter forming a parallel resonance with the system impedance. The equivalent circuit is shown in Figure 2.11.

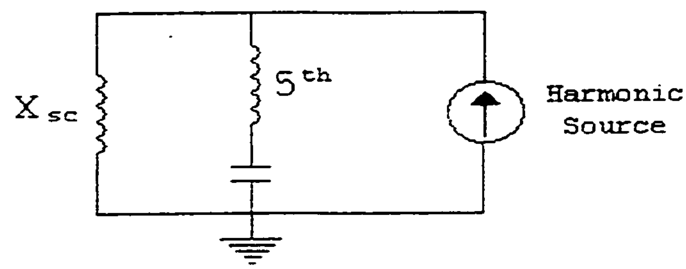

Figure 2.10 Single Tuned Filter Interaction with Source

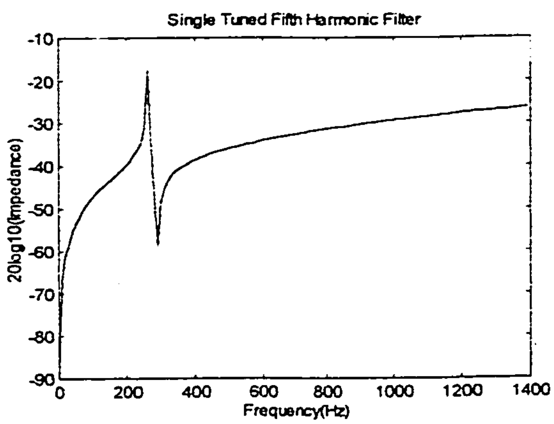

Figure 2.11 Frequency Response of Single Tuned Filter and System Interaction

Second order high pass filters can exhibit elements of both the first order and the single tuned filters. The frequency response of the filter is given by:

$$|Z| = \left| -\frac{j}{\omega C} + \frac{\omega LR}{R + j\omega L} \right| \quad (2.7)$$

Figure 2.12 shows four curves of second order high-pass frequency response for Q factors of 300, 30, 3 and 0.3. Q factor is defined as

$$Q = \frac{R}{k\omega L} \quad (2.8)$$

where k is the harmonic frequency to be filtered

R is the resistance of the filter

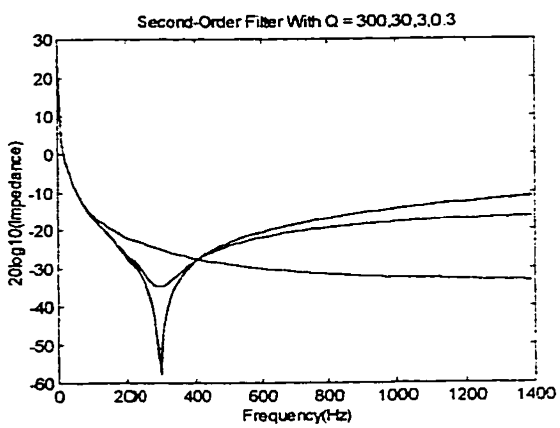

Figure 2.12 Second Order High Pass Filter with Several Q Values

A second order high pass filter with a large Q resembles a single tuned filter frequency response and a first order filter with a small Q. Depending on the value of Q, a second order high pass filter has the advantages and disadvantages of the first order and notch filters. Second order high pass filters with Q between 0.5 and 2 are used in filter banks to attenuate all frequencies above a certain order.

Harmonic filter banks, composed of several single tuned filters and a second order filter, are connected at points of harmonic injection to prevent excessive current from entering the system. Shown in Figure 2.13 is a filter bank with single tuned filters at the 5^{th}, 7^{th}, 11^{th} and 13^{th} with a second order filter designed to attenuate all frequencies above the 17^{th}. Using filters in banks allows designers to take advantage of the low loss/sharp tuning qualities of the notch filters to remove the harmonic that have large currents and reserving the broad-spectrum attenuation capabilities of the second order filters for the harmonics with relatively low content. A typical frequency response for a filter bank like the one shown in Figure 2.13 is given in Figure 2.14. From the frequency response plot it can be seen that there is a frequency range near the tuning point that has high impedance. This resonant peak occurs as a result of parallel resonance between the filter and the system. If this peak falls near to a problem harmonic frequency, large harmonic currents may be created. When filters are designed the filter system interactions must be taken into account to avoid parallel resonance.

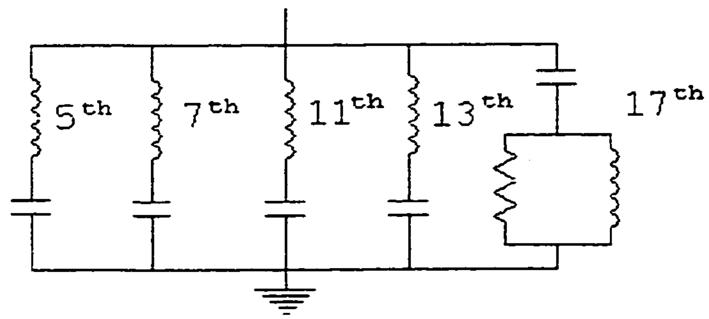

Figure 2.13 Filter Bank of Single Tuned Filter with a Second Order Filter for Higher Harmonics

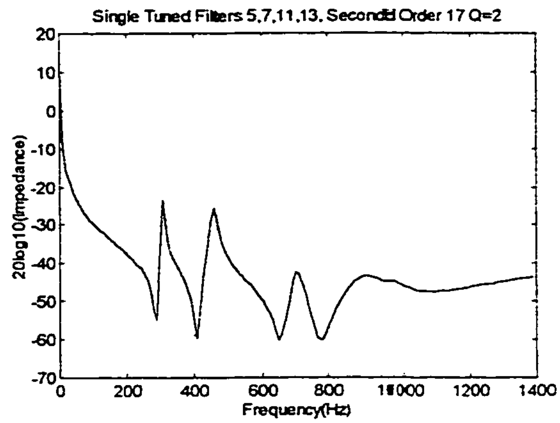

Figure 2.14 Filter Bank with 5th, 7th, 11th, 13th, and 17th+ Single Tuned Filters

The prime disadvantage of passive filters stems from their static frequency response. If the filter impedance changes, harmonic content will vary possibly rendering the filter ineffective. Both system and filter element parameters may drift due to temperature or damage. Also since utilities are continually reconfiguring sections of the their networks to accommodate customers, these changes may inadvertently introduce unwanted system and filter interaction.

2.3 Active Filters

An alternative to using static elements to passively filter harmonic currents is the active filter. These devices use a power converter to supply the power source line with the harmonics required by the non-linear load. The power supplied, from the utilities point of view, is in the form of current or voltage of equal magnitude but opposite phase to that injected by the harmonic causing load.

There are two basic active filter types used to correct distortion[5]:

1) Voltage inverters – mainly used to correct voltage distortions
2) Current inverters – mainly used to correct current distortions

Voltage or current injection refers to the controlled connection of a dc source of voltage or current. The dc source is connected through an inverter, Figure 2.15 shows inverters used for both voltage and current injection. Both devices can positively connect, negatively connect or disconnect a dc source to resist changes in voltage or current. For voltage injection a capacitor is connected because it resists voltage changes while for current an inductor is used. The dc sources can be powered either through the switching action of the inverter or through a secondary charging source.

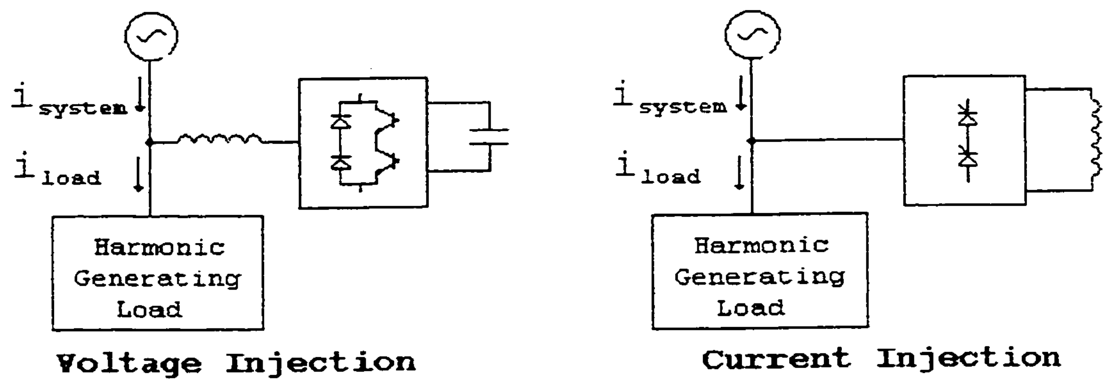

Figure 2-15 Voltage and Current Inverters For Active Filters

Current injection systems require simple control mechanisms but they also tend to incur large losses and are therefore only suitable for low power applications. Voltage injection systems on the other hand can be added in parallel to increase the overall rating of the filter. While the primary drawback for voltage injection is related to the complexity of their control systems.

The actions taken by the inverter to correct the distorted current or voltage varies depending on the method of analysis. Frequency domain analysis involves a time averaged estimate of the periodic waveforms spectral content. The inverter, either through injection of voltage or current, supplies the utility with harmonic content of opposite phase to cancel out the distortions. The frequency analysis may be performed once when the filter is first designed so that the same signal is always injected. The filter may also be used for "real-time" compensation of harmonic changes that occur. The system is only delayed by the length of the cycle sampled and the time to compute the content of the highest harmonic.

Time based analysis methods differ from frequency methods in that they correct the distorted signal "on-line". Voltage or current injections are preformed so as to make the distorted signal sinusoidal. An error signal is calculated by subtracting the present value of the voltage or current from the desired ideal one. Two or three-state switching functions take this error signal and generate a course of action for the inverter. Examples of a two-state and three-state switching function are shown in Figure 2.16. A two state switching function will reverse the connection of the inverter every time the error signal intersects a high frequency comparison function. In figure 2.16 a triangular comparison function is used. A three-state switching function will disconnected the inverter as long as the error stays within certain limits. When the calculated error leaves this region the inverter is switched so as to correct the distortion.

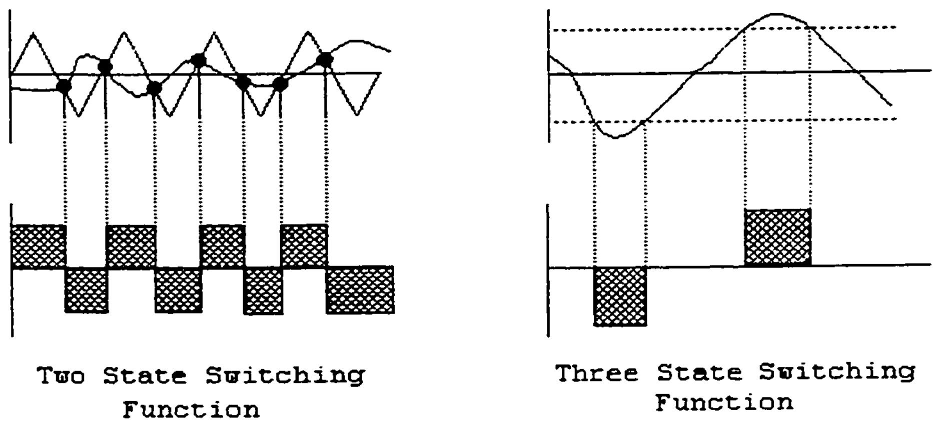

Figure 2-16 Time Domain Correction

Both methods have the net effect of reducing the calculated error and correcting harmonic distortion. Frequency based methods suffer from computation delay incurred during the Fourier transform but have the advantage of providing a time averaged moderated response. Time based methods on the other hand are real-time but may as a result respond dramatically to transients that could be ignored.

2.4 Filter Placement

The obvious location for active and passive filters is near the source of the harmonic currents. In this way harmonics can be removed to prevent any damage to the utility. Harmonic currents and therefore harmonic power flow in the direction of least impedance[28] which is generally toward the utility[1]. Figure 2.17 shows the general propagation of harmonics within a power system. This direction of propagation can be seen in the case studies presented in this chapter. A result of harmonic propagation towards the utility bus is that harmonics generated by one the utilities customers system may propagate into another customers through the utility system. Depending on the level of the harmonic distortion, the utility may require the offending customer install new harmonic filters or re-tune the presently ineffective ones.

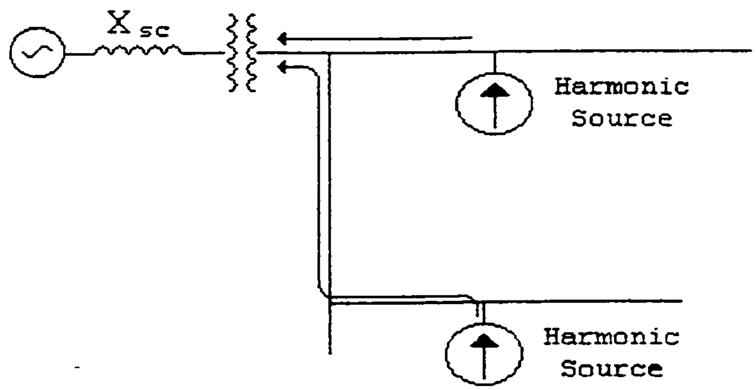

Figure 2-17 General Propagation of Harmonics

2.5 Example Case

The test system to be studied is representative of a medium-sized industrial plant. The system is extracted from a common system that is being used in many of the calculations and examples in the IEEE Color Book Series. The system depicted in Figure 2.18 has 13 buses fed from a utility supply of 69kV, a local plant system of 13.8kV and several loads. The system is balanced and free of line capacitance. Line impedance and transformer data are presented in Tables 2.1 and 2.2 respectively, Table 2.3 contains all the load bus data while Table 2.4 contains the harmonic profile of a adjustable speed drive that will be used as a harmonic current source.

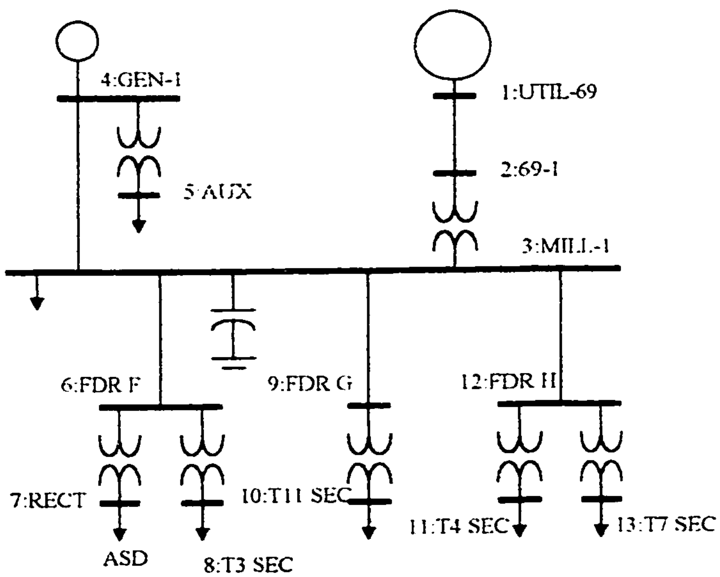

Figure 2.18 Test System

Table 2.1 Per-Unit Line Impedance Data

From	To	R (pu)	X (pu)
UTIL-69	69-1	0.0139	0.0296
MILL-1	GEN-1	0.0243	0.0243
MILL-1	FDR F	0.0075	0.0063
MILL-1	FDR G	0.0157	0.0131
MILL-1	FDR H	0.0109	0.0091

Table 2.2 Transformer Data

From	To	Voltage	kVA	R (pu)	X (pu)
69-1	MILL-1	69:13.8	15000	0.004698	0.079862
GEN1	AUX	13.8:0.48	1500	0.009593	0.056694
FDR F	RECT	13.8:0.48	1250	0.007398	0.044388
FDR F	T3 SEC	13.8:4.16	1725	0.007442	0.05937
FDR G	T11 SEC	13.8:0.48	1500	0.008743	0.056831
FDR H	T4 SEC	13.8:0.48	1500	0.008363	0.054360
FDR H	T7 SEC	13.8:2.4	3750	0.004568	0.054810

Table 2.3 Load Bus Data

Bus	P_{load} (kW)	Q_{load} (kVAr)
MILL-1	2240	2000
AUX	600	530
RECT	3000	800
T3 SEC	1310	1130
T4 SEC	370	330
T7 SEC	2800	2500
T11 SEC	810	800

Table 2.4 Harmonic Source

Harmonic	Magnitude (% of Fund.)	Relative Phase Angle
1	100	0
5	18.24	-55.69
7	11.90	-84.11
11	5.73	-143.56
13	4.01	-175.58
17	1.93	111.58
19	1.39	68.30
23	0.94	-24.61
25	0.86	-67.64
29	0.71	-145.46
31	0.62	176.83
35	0.44	97.40
37	0.38	54.36

Additional information

1) The utility supply has an impedance of 0.05+j1 pu at a voltage of 1.00Vpu

2) The in plant generator has an impedance of j0.25 pu at a voltage of 0.995 pu and is limited to 2000kW
3) The power factor correction capacitor is rated at 5,000 kVAr.
4) The system has base values of 13.8 kV and 100 MVA.
5) As a benchmark for evaluating the cases the utility will tolerate voltage THD up to 1.00% at its connection point.

Using PSA-H developed by Power Tech Software several load flows cases were calculated. Case 0 shows the system behavior when harmonic sources are not present. In Case 1 an adjustable speed drive (ASD) is connected to bus 7 to inject harmonic currents. A comparison of this case with the Case 0 will show how the presence of a non-linear load can distort voltage and current waveforms throughout the system. For the last two cases a filter is connected at bus 7. By varying the capacitance of the filter between Case 2 and Case 3 the sensitivity of the system to filter parameters can be demonstrated.

In each case when the harmonic load flow is performed on the system, the resulting bus voltages, line currents and their THD are used to make quantitative comparisons. Further for each case a frequency scan is performed. A frequency scan reveals the system impedance at the point of harmonic injection for all harmonics of interest. The frequency scan aids in understanding how a harmonic filter effects the system behavior when harmonics are present. A summary of the different cases is presented in Table 2.5.

Table 2.5 System Setup Summary

Case Number	Harmonic Source	Harmonic Filter	Capacitive Variation
0	No	No	N/A
1	Yes	No	N/A
2	Yes	Yes	0%
3	Yes	Yes	+5%

For Cases 2 and 3 a harmonic filter is placed at bus 7. The capacitor used in the filter is rated for 800 kVAr. To calculate the reactive impedance of the capacitor

$$X_C = \frac{1}{\omega C} = \frac{kV^2}{MVAr} = \frac{(0.480)^2}{0.800} = 0.288\Omega \qquad (2.9)$$

The value of the reactor required can be obtained from

$$X_L = \omega L = \frac{X_C}{h^2} = \frac{0.288}{(4.8)^2} = 0.0125\Omega \qquad (2.10)$$

where h is the harmonic to which the filter is to be tuned

ω is the fundamental frequency

The filter used in Case 2 is tuned to the 4.8th harmonic or 288Hz. Table 2.6 contains a summary of the filter parameters for the last two cases. Figure 2.19 shows the impedance frequency response of both cases. For Case 3 the value of the capacitor has deviated from that of Case 2 by +5% to demonstrate the system sensitivity to filter parameter fluctuation.

Table 2.6 Filter Parameters

Case Number	Tuned Harmonic h	Reactor Value L (mH)	Capacitor Value C (uF)
2	4.8	0.0331572	9210.36
3	4.6843	0.0331572	9670.87

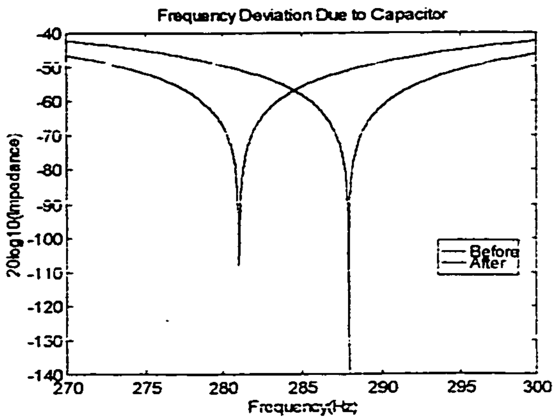

Figure 2.19 Frequency Response Change Due to Filter Capacitor Fluctuation

Case 0: No Harmonic Source or Filter

In the first case no harmonic sources or filters are present. Table 2.7a presents the per-unit voltage at each bus and Table 2.7b the per-unit current through each branch. Since there are no harmonic currents all voltage and current THDs are zero. The fundamental current and the total RMS current are identical because no harmonic currents exist. A frequency scan at bus 7 (Figure 2.20) shows the system impedance dominated by reactance.

Table 2.7a Case 0: Bus Voltage and THD

Bus	Voltage Magnitude (pu)	Voltage Angle (deg)	THD (%)
UTIL-69	1.0000	0.00	0.0000
69-1	0.9985	-0.16	0.0000
MILL-1	0.9939	-3.15	0.0000
GEN-1	0.9950	-3.16	0.0000
AUX	0.9702	-4.30	0.0000
FDR F	0.9934	-3.16	0.0000
RECT	0.9302	-10.45	0.0000
T3 SEC	0.9453	-5.62	0.0000
FDR G	0.9937	-3.15	0.0000
T11 SEC	0.9567	-4.72	0.0000
T4 SEC	0.9789	-3.83	0.0000
FDR H	0.9933	-3.15	0.0000
T7 SEC	0.9505	-5.45	0.0000

Table 2.7b Case 0: Branch Currents and THD

From	To	Current Fundamental(pu)	THD (%)	Total Current RMS (pu)
UTIL-69	69-1	0.098	0.000	0.098
MILL-1	GEN1	0.039	0.000	0.039
MILL-1	FDR F	0.056	0.000	0.056
MILL-1	FDR G	0.012	0.000	0.012
MILL-1	FDR H	0.045	0.000	0.045
69-1	MILL-1	0.098	0.000	0.098
GEN1	AUX	0.008	0.000	0.008
FDR F	RECT	0.038	0.000	0.038
FDR F	T3 SEC	0.018	0.000	0.018
FDR G	T11 SEC	0.012	0.000	0.012
FDR H	T4 SEC	0.005	0.000	0.005
FDR H	T7 SEC	0.039	0.000	0.039
PF Cap	Ground	0.050	0.000	0.050

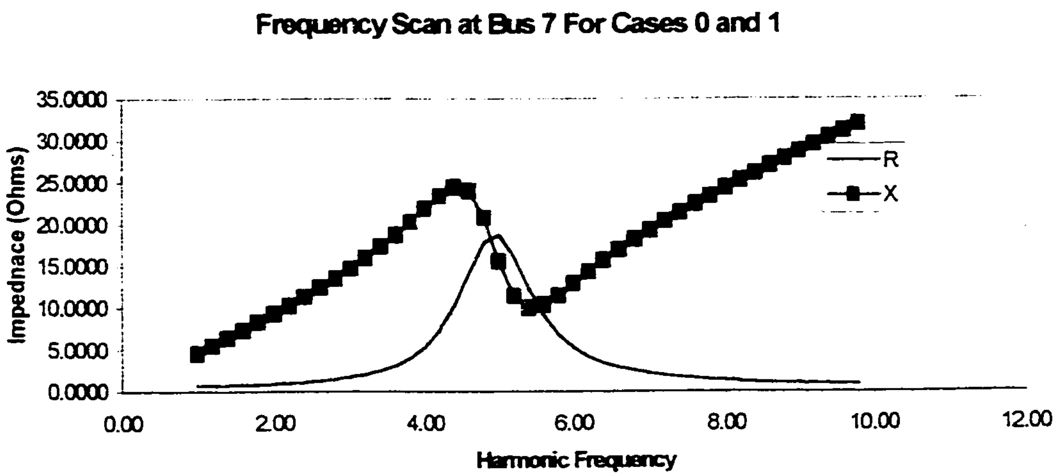

Figure 2-20 Frequency Scan at Bus 7 For Case 0 and Case 1

Case 1: ASD Harmonic Source and No Harmonic Filter

For case 1 the harmonic source is added at bus 7. As shown in Tables 2.8a and 2.8b the harmonic source causes voltage and current distortion on all buses and branches. The voltage distortion is worst at the injection bus and exceeds the 1.00% limit. The current distortion is largest along the path from the injection bus to ground through the power factor correction capacitor. This path has the lowest impedance for the harmonics generated by the ASD and thus show a large increase in the total RMS current. There is no change in the frequency scan since the system impedance was not changed.

Table 2.8a Case 1: Bus Voltage and THD

Bus	Voltage Magnitude (pu)	Voltage Angle (deg)	THD (%)	Total RMS Voltage (pu)
UTIL-69	1.0000	0.00	7.218	1.0026
69-1	0.9985	-0.15	7.442	1.0013
MILL-1	0.9940	-3.00	11.342	1.0004
GEN-1	0.9950	-3.00	11.298	1.0013
AUX	0.9702	-4.14	9.966	0.9750
FDR F	0.9935	-3.00	11.340	0.9999
RECT	0.9384	-9.27	21.586	0.9600
T3 SEC	0.9454	-5.46	9.012	0.9492
FDR G	0.9938	-2.99	11.336	1.0001
T11 SEC	0.9568	-4.56	9.477	0.9611
T4 SEC	0.9790	-3.67	10.489	0.9844
FDR H	0.9934	-2.99	11.328	0.9997
T7 SEC	0.9506	-5.29	9.121	0.9545

Table 2.8b Case 1: Branch Currents and THD

From	To	Current Fundamental (pu)	THD (%)	Total Current RMS (pu)
UTIL-69	69-1	0.09303	15.343	0.09412
MILL-1	GEN1	0.03645	7.238	0.03654
MILL-1	FDR F	0.05051	14.912	0.05107
MILL-1	FDR G	0.01190	10.666	0.01196
MILL-1	FDR H	0.04455	10.211	0.04478
69-1	MILL-1	0.09303	15.343	0.09412
GEN1	AUX	0.00825	11.321	0.00830
FDR F	RECT	0.03308	23.065	0.03395
FDR F	T3 SEC	0.01830	9.685	0.01838
FDR G	T11 SEC	0.01190	10.665	0.01197
FDR H	T4 SEC	0.00506	12.150	0.00510
FDR H	T7 SEC	0.03949	9.966	0.03968
PF Cap	Ground	0.04970	58.118	0.05748

Case 2: ASD Harmonic Source and Filter Tuned to the 4.8 Harmonic

In case 2 a harmonic filter has been added at bus 7 to shunt some harmonic current to ground. The frequency scan in Figure 2.21 shows the low impedance path near the 5th harmonic due to the harmonic filter. The voltage THD given in Table 2.9a shows a marked reduction in voltage distortion. The utilities 1.00% limit at its connection point is once again met. The harmonic filter has removed a large portion of the 5th harmonic current injected by the ASD. Table 2.9b reveals that while the path from the injection bus to ground through the power factor correction capacitor has relatively large current THD, the largest value is found through the harmonic filter. This shows that the THD reduction is due to the presence of the filter.

Table 2.9a Case 2: Bus Voltage and THD

Bus	Voltage Magnitude (pu)	Voltage Angle (deg)	THD (%)	Total RMS Voltage (pu)
UTIL-69	1.0000	0.00	0.940	1.0000
69-1	0.9986	-0.15	0.969	0.9986
MILL-1	0.9942	-2.99	1.477	0.9943
GEN-1	0.9950	-2.99	1.471	0.9951
AUX	0.9702	-4.14	1.276	0.9703
FDR F	0.9938	-3.00	1.468	0.9939
RECT	0.9688	-9.35	7.916	0.9718
T3 SEC	0.9457	-5.46	1.132	0.9457
FDR G	0.9940	-2.99	1.476	0.9941
T11 SEC	0.9570	-4.56	1.207	0.9571
T4 SEC	0.9792	-3.67	1.351	0.9793
FDR H	0.9936	-2.99	1.475	0.9937
T7 SEC	0.9508	-5.29	1.154	0.9508

Table 2.9b Case 2: Branch Currents and THD

From	To	Current Fundament al(pu)	THD (%)	Total Current RMS (pu)
UTIL-69	69-1	0.09292	1.661	0.09294
MILL-1	GEN1	0.02917	1.004	0.02917
MILL-1	FDR F	0.04693	5.416	0.04700

MILL-1	FDR G	0.01189	1.236	0.01190
MILL-1	FDR H	0.04454	1.186	0.04454
69-1	MILL-1	0.09292	1.661	0.09294
GEN1	AUX	0.00825	1.326	0.00825
FDR F	RECT	0.03096	7.887	0.03106
FDR F	T3 SEC	0.01829	1.119	0.01829
FDR G	T11 SEC	0.01190	1.236	0.01190
FDR H	T4 SEC	0.00506	1.430	0.00506
FDR H	T7 SEC	0.03948	1.156	0.03948
PF Cap	Ground	0.04971	9.801	0.04995
Filter	Ground	0.00810	73.134	0.01004

In order to demonstrate the sensitivity of the system to harmonic filter variation and therefore the usefulness of the self-tuning filter, one more case is presented. Case 3 shows what happens to the system bus voltage, line current and THD when the capacitor drifts.

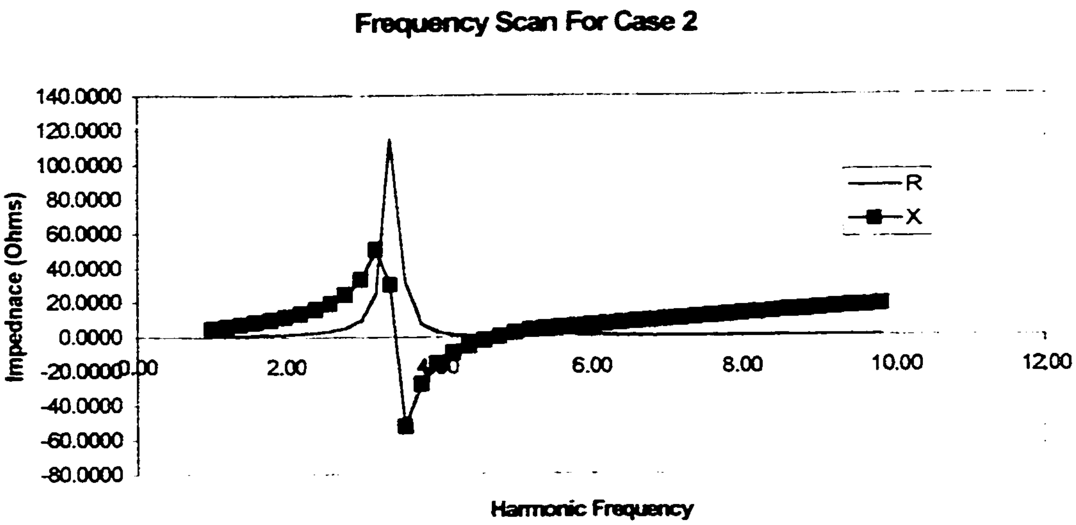

Figure 2.21 Frequency Scan at Bus 7 For Case 2

Case 3:ASD Harmonic Source and +5% Change in Filter Capacitance

In case 3 the value of the capacitance increases by 5% placing the tuning frequency at the 4.6843th harmonic instead of the 4.8th harmonic. The frequency scan in Figure 2.22 at the injection bus shows that at the 5th harmonic the reactive impedance is greater than that for case 2. This results in less 5th harmonic current being removed from the system increasing both the voltage and current THD. Notably the voltage at the utility is now greater than the 1.00% limit. Table 2.10a and 2.10b show the results of the harmonic load flow calculation.

Table 2.10a Case 3: Bus Voltage and THD

Bus	Voltage Magnitude (pu)	Voltage Angle (deg)	THD (%)	Total RMS Voltage (pu)
UTIL-69	1.0000	0.00	1.144	1.0001
69-1	0.9986	-0.15	1.180	0.9986
MILL-1	0.9942	-2.99	1.798	0.9944
GEN-1	0.9950	-2.99	1.791	0.9952
AUX	0.9702	-4.14	1.562	0.9703
FDR F	0.9938	-3.00	1.791	0.9939
RECT	0.9705	-9.35	8.062	0.9736
T3 SEC	0.9457	-5.46	1.395	0.9458
FDR G	0.9940	-2.99	1.797	0.9941
T11 SEC	0.9570	-4.56	1.481	0.9571
T4 SEC	0.9792	-3.67	1.651	0.9793
FDR H	0.9936	-2.99	1.796	0.9937
T7 SEC	0.9508	-5.29	1.419	0.9509

Table 2.10b Case 3: Branch Currents and THD

From	To	Current Fundamental (pu)	THD (%)	Total Current RMS (pu)
UTIL-69	69-1	0.09292	2.165	0.09294
MILL-1	GEN1	0.02879	1.312	0.02879
MILL-1	FDR F	0.04678	5.614	0.04685
MILL-1	FDR G	0.01189	1.568	0.01190
MILL-1	FDR H	0.04454	1.504	0.04454
69-1	MILL-1	0.09292	2.165	0.09294
GEN1	AUX	0.00825	1.676	0.00825
FDR F	RECT	0.03091	8.191	0.03102
FDR F	T3 SEC	0.01829	1.421	0.01829
FDR G	T11 SEC	0.01190	1.568	0.01190
FDR H	T4 SEC	0.00506	1.804	0.00506
FDR H	T7 SEC	0.03948	1.466	0.03948
PF Cap	Ground	0.04971	11.118	0.05002
Filter	Ground	0.00854	67.219	0.01029

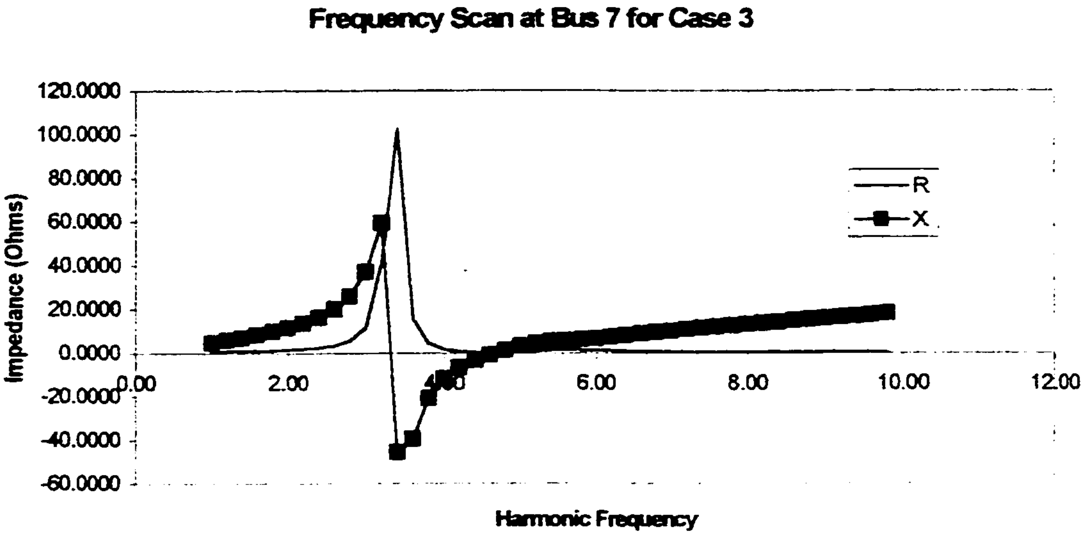

Figure 2.22 Frequency Scan at Bus 7 For Case 3

2.6 Conclusions

In this chapter the methods of harmonic generation and quantification were presented. The benefits and drawbacks of passive and active filtering were outlined. Passive filtering for harmonic mitigation is weak in terms of flexibility, but is preferable over active filtering due to less complexity. In the last section of the chapter four cases of an example system were presented. The inclusion of a harmonic source increases the THD on all voltage buses and branch current. A harmonic filter placed near the source of the harmonics can be used to reduce the voltage and current THD to acceptable limits. Variation of the filter tuning due to changes in capacitance will affect the THD of all bus voltages and line currents. As a result of the changes in voltage THD some values exceed specified limits and the system may need to be altered by adding more filters or correcting the old one.

Chapter 3

Self-Tuning Harmonic Filter

In this chapter a device is proposed that will resolve some of the problems associated with passive harmonic filtering. The solution is a self-tuning filter using a thyristor linked reactor as the variable element. References to other self-tuning filters as well as outlines of their designs are given.

3.1 Motivation for a Variable Inductor

Cases 2 and 3 presented in the last chapter show the effects of a harmonic filter capacitor fluctuation. When there is a change in the capacitance of the harmonic filter the impedance frequency response of the filter changes. In the case presented in chapter 2 the capacitance increased by 5%. Table 2.6 and Figure 2.19 show the filter parameters and frequency response for cases 2 and 3. In Figure 2.19 it can be seen that the tuning frequency of the filter has shifted to a lower value. The result of this shift was a increase in the impedance at the 5^{th} harmonic. With less harmonic current going to ground the amount injected into the system increased. This resulted in a violation of the 1% voltage THD limit at the utility bus.

To solve the problem of increased voltage THD at the utility bus using passive filter technology one of the following must be done:

1) The damaged component must be replaced
2) The temperature sensitive device must be housed in a controlled environment
3) More filters must be added to the system.

All of these solutions result in additional material and engineering costs. Further they provide little guaranty from future parameter fluctuations causing similar problems. Active filtering methods may be applicable but have up to present day not provided feasible solutions. What is required is a technology that:

1) Is similar in application to passive filter technology
2) Configurable in the sense that it can be changed to meet foreseeable fluctuations.
3) Ideally self-reliant in that it does not require continual re-engineering

Obvious solutions to the first two technology requirements are the introduction of a variable capacitor or inductor into the filter to compensate for any fluctuations that occur in the filter. In this way if the filter becomes de-tuned the reactive impedance of the variable device can be changed to re-tune the filter. The third requirement will be discussed later in this chapter and at length in chapter 5.

Conceptual diagrams of the variable capacitor and inductor are presented in Figure 3.1a and b respectively. The variable capacitor is a set of parallel switch-able capacitors. The capacitance value of the variable device is equal to the sum of each capacitor with a closed switch. These switch-able devices are presently used alone in power systems for power factor correction and voltage support[18,19,20]. The primary drawback of using a variable capacitor for a tune-able filter is the financial cost due to the fact that multiple capacitors each with full voltage rating are required. Another disadvantage of this scheme is the need to control capacitor switching using sophisticated digital controls.

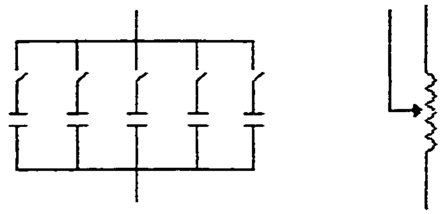

Figure 3.1 Variable Reactive Elements

There are two standard sets of methods for implementing a variable inductor. The first method uses mechanical devices to control either the number of flux linkages of the inductor or the reluctance of the magnetic circuit passing through the coils. The number of flux linkages is altered mechanically by using a tapped inductor coil [21] and a tap-changing mechanism which is a relay controlled motor[13]. Altering, again mechanically, the reluctance of the coil is achieved by controlling the movement of a low reluctance material within the inductor. As an example, in the device under Patent Number 4357489[22] (seen presented in Figure 3.2) by using a hand crank any length of a threaded iron core can a moved into a current carrying coil. The second set of methods for implementing a variable inductor relies upon electric methods of altering the permeability of the inductor core. An example of this is shown in Figure 3.3 and is found in Patent Number 4393157[23]. In that device two orthogonally placed cores magnetically intersect in two areas. One core carries the alternating magnetic current and the other a user controlled DC magnetic current. The strength of the DC current alters the permeability of the intersecting regions and in this way the over all permeability of the core and therefore the inductance of

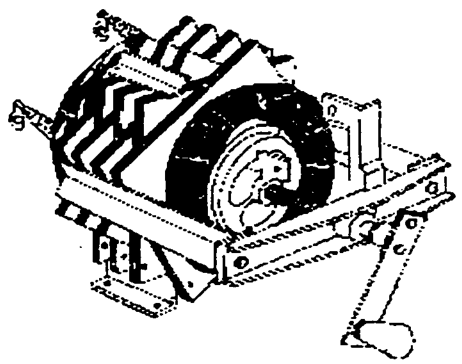

Figure 3.2 Patent Number 4357489

Variable Reactor

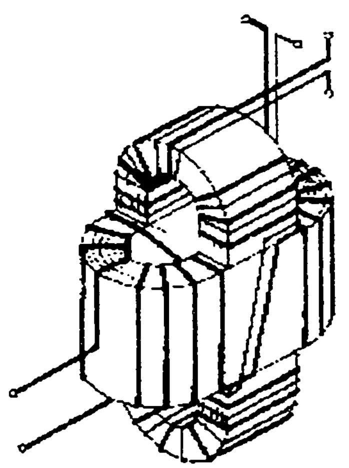

Figure 3.2 Patent Number 4393157

Variable Reactor

the alternating current coil. These two sets of methods suffer from the same drawbacks:

1) They need secondary power supplies and thus incur more losses than that of a simple inductor
2) They require extra equipment in the form motors and relays or secondary windings and power rectifiers

In this thesis a novel topology for a variable inductor is presented. The device proposed has an advantage over the variable capacitor in that only a single, tapped inductor of rated value is required. Further the device is electronic in nature having no mechanical apparatus and does not require a secondary power source. These factors make the variable inductor device favorable over the present implementation.

3.2 Thyristor Linked Reactor

The key elements in the novel device to be presented are thyristors. In the thyristor linked reactor thyristors are connected back to back so the opposite polarity ends connect. Pulsing the thyristor gates simultaneously allows current to travel in either direction. When the gate stimulus is removed the thyristors will maintain continuity until the current flow is zero.

In the rest of this work the thyristors will be treated as ideal. The back to back thyristors will act as a switch which turns on when gate signal is present and off when both gate signal and current are zero. Three possible topologies for the variable inductor are given in Figure 3.4. With each topology the number of loops of the inductor coil can be varied electronically. Using the thyristor switches the passive elements can be added or removed from the circuit changing the inductance. Several of these devices connected in series allow multiple inductance steps to be achieved. These novel devices are labeled as Thyristor Linked Reactors (TLR).

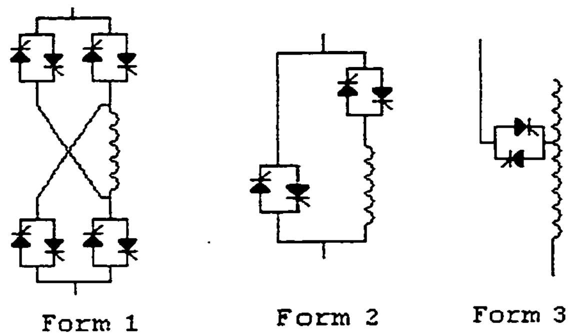

Figure 3.4 Three TLR Implementations

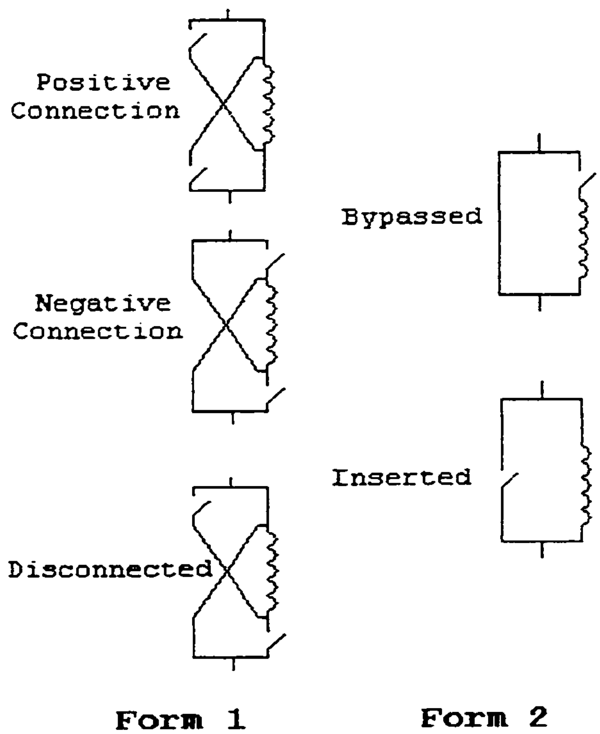

Figure 3-5 Useful Connections of Form 1 and 2

Forms 1 and 2 presented in Figure 3.4 use similar methods to either include or bypass an individual inductor segment. Figure 3.5 shows how the thyristors of Form 1 and 2 can be gated to include, reverse or remove any combination of individual inductor segments. Form 3 differs in that only the leading inductor segments can be added or removed. An example of each form is presented in Figure 3.6.

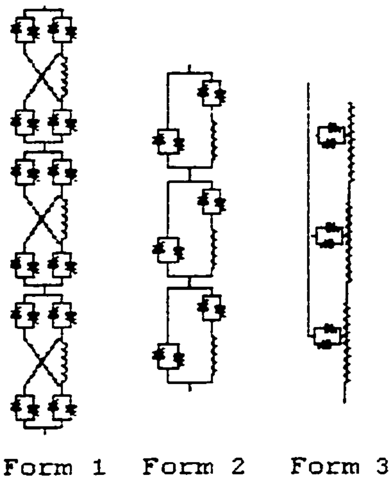

Figure 3.6 Examples of Each TLR Form

Forms 1 and 2 have 8 possible inductance combinations while form 3 has only 3. It can be seen that by adding one more segment (those found in Figure 3.4) the number of possible inductance combinations doubles for forms 1 and 2 but only increments by one for form 3. Form 2 is the topology used in the rest of this work. The reason for this decision is based on a tradeoff between three factors

1) The number of thyristors required per step
2) The number of possible steps per inductor element
3) The maximum reverse voltage experienced by thyristors when they are turned off.

Ideally it is desirable to have a TLR that uses a minimum number of thyristors to achieve the maximum number of steps while maintaining reasonable voltages across thyristors that are turned off. The last factor arises because thyristors have a breakdown voltage at which the element will begin to conduct even if there is no gate signal present. To maintain normal operation the thyrsitor must not have a voltage exceeding the breakdown voltage across it at any time.

To evaluate the three forms in terms of the above trade-offs three circuits are studied. In Figure 3.6 harmonic filters using TLRs with forms 1, 2 and 3 are given. A TLR functions as a variable reactor with finite resolution and dynamic range. The magnitude of the inductance is dictated by which segments of the TLR have a gate pulse. The TLR is proposed as a tapped reactor mounted on a single core.

In terms of the number of thyristors required per step Table 3.1 rates the forms with form 3 being the best and form 1 the worst. Also in Table 3.1 the number of steps that result from an n segment inductor is presented. Comparing these factors form 2 is the most acceptable model for the TLR implementation.

Table 3.1 Comparison of Forms

Form	Thyristors Per Step	Steps Per n Inductor Element
1	8	2^n
2	4	2^n
3	2	n

The third factor to consider is in the steady state operation of the circuit. When the thyristors are gated there is conduction in the direction of current. When the gates are not pulsed there is no conduction and thus a potential difference may exist across the thyrsitors. Thyristors have a physical parameter called the reverse breakdown voltage. At this voltage the thyristor will begin to conduct and current will flow backwards. Since each pair of thyristor is connected back to back, the polarity of the voltage difference for the TLR is irrelevant. To maintain proper operation the reverse breakdown voltage must not be exceeded.

For each form the maximum voltages seen by thyristors that are turned off can be calculated in terms of inductance (either bypassed or included), magnitude and frequency of current. For forms 1 and 2 large voltages will be experienced by the bypass thyristors for any included inductor element, in phasor form the voltages can be written:

$$V = j\omega L_{included} I \qquad (3.1)$$

For form 3 the thyristos that are turned off will experience a potential drop equal to that generated by all included elements.

$$V = j\omega I \sum L_{included} \qquad (3.2)$$

Each TLR device is mounted on a signal core. This results in voltage potential being generated on the bypassed inductor segments. For all the forms the voltage generated across a bypassed element can be calculated as

$$V_{bypassed} = \frac{N_{bypassed}}{N_{included}} V_{included} \qquad (3.3)$$

where $V_{bypassed}$ is the voltage induced across the bypassed turns
$V_{included}$ is the voltage across the turns that are carrying current
$N_{bypassed}$ is the number of turns in the bypassed segment
$N_{included}$ is the number of turns carrying current

To the relate the number of turns in equation 3.3 to the inductance of the respective element the core is considered to be a solenoid thus

$$L = \frac{\mu N^2 A}{\ell} \qquad (3.4)$$

where L is the inductance of the element in Henrys

μ is the permeability of the material of the core

N is the number of turns carrying current around the coil

A is the cross sectional area of the coil

ℓ is the length of the coil

If ℓ is proportional to the number of turns then

$$\ell = RN \qquad (3.5)$$

Then L the inductance becomes proportional to N and equation 3.4 can be written

$$L = \frac{\mu NA}{R} \qquad (3.6)$$

Writing equation 3.3 using 3.6 gives

$$V_{bypassed} = \frac{L_{bypassed}}{L_{included}} V_{included} = j\omega I L_{bypassed} \qquad (3.7)$$

As for the bypass thyristors form 3 has the ability to generate the largest reverse voltages since multiple segments may be bypassed. Therefore in-terms of the third factor listed above form 1 and 2 are superior to 3 be equal to each other. Comparing the three forms using all criteria form 2 is the most desirable. With relatively low off state voltages form 2 has the largest number of combinations of inductive elements using the least number of thyristors.

The proposed scheme used in the rest of this work uses form 2 with binary related inductor segments. The inductor segments are binary in that their inductance values successively double. The smallest inductance represents the resolution with the range of the TLR being 0 H to $(2^n-1)L_{smallest}$ H where $L_{smallest}$ is the smallest (resolution) inductance step and n is the number of inductor segments in the TLR.

An example of a 4 segment TLR suitable for use in a harmonic filter is proposed here. This example is an extension of the work done in chapter 2. In chapter 2 a harmonic filter tuned to the 4.8^{th} harmonic was designed with an inductor with a reactive impedance of $X_L = 0.0125\Omega$. In order to make the reactor variable a TLR is introduced. The TLR is specified to vary the inductive reactance +10%/-10% around a central value of X_L. Working in-terms of the inductance the resolution and range of the TLR are specified in Table 3.2.

Table 3.2 TLR Range and Resolution

Parameter	Value
Minimum Step Size	0.4421uH
Inductance range	36.4729µH to 29.8414mH

The inductor segment nearest to ground is not switch-able because it represents the minimum inductance value of the TLR. To calculate the resolution, the inductance value of the smallest inductor:

$$L_1 = \frac{L_{Largest} - L_{smallest}}{2^n - 1} = \frac{L_{range}}{2^n - 1} = \frac{(36.471\mu H) - (29.841\mu H)}{15} = 0.4421\mu H \quad (3.8)$$

The rest of the inductor values for the TLR are successive multiplies of 2

$$L_2 = 2^1 \times L_1 = 0.8842\mu H$$

$$L_3 = 2^2 \times L_1 = 1.7684\mu H$$

$$L_4 = 2^3 \times L_1 = 3.5368\mu H \tag{3.9}$$

The value of the un-switched inductor segment is

$$L_{smallest} = 29.841\mu H$$

The complete TLR is depicted in Figure 3.7

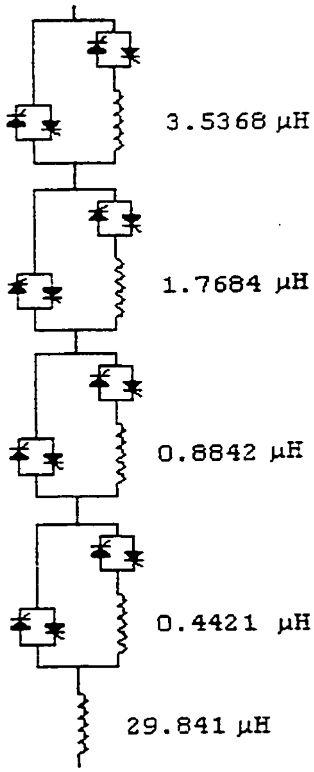

Figure 3.7 Binary Form 2 TLR

3.3 Self-Tuning Filter

At the beginning of this chapter three requirements for an improved filter technology were presented. The first of these was that the new type of filter would be applied and designed in a manner similar to the present day passive filter technology. Replacing the standard inductor with a TLR (Figure 3.6) of similar ratings will produce a harmonic filter that can play the same role as the standard filter. The second requirement was that the new filter technology should be re-configurable to correct for fluctuations including parameter changes within the filter itself. The filter presented in Figure 3.6 again meets this criteria to the limit of the range and the resolution of the TLR. Equation 3.11 shows the relationship of the resistance R, inductance L, capacitance C and frequency ω to the impedance of the filter.

$$Z = R + j\omega L - \frac{j}{\omega C} \tag{3.11}$$

From equation 3.1 it can be seen that varying the inductance can provide some compensation for changes in the capacitance

The third requirement was that the filter should be self-reliant in that it does not require "re-engineering" when fluctuations occur. Particular values of L and C for a filter lead to minima in filter impedance. Several examples of C and L combinations are presented Table 3.3. In these examples the base value of the capacitor is 9210.36μF while the base value of the inductor is 33.1572μH and the tuning frequency was selected as 2π*60*4.8 rads. Table 3.3 shows selected examples of how a variable inductor can compensate for changes in capacitance.

Table 3.3 Selected Examples of Capacitor Fluctuation and Compensating Inductance

Change In Capacitance (%)	Capacitor Value μF	Compensating Change in Inductance (%)	Compensating Inductance Value μH
-15	7828.8	17.64	39.008
-5	8749.8	5.263	34.902
5	9670.9	95.24	31.578
15	10591.9	86.95	28.832

The results in Table 3.3 show how using a variable inductor fluctuations in the capacitance can be corrected.

An ideal filter technology would use information about L and C values to correct for fluctuations in filter tuning. In this way the filter would be self-tuning. When a frequency shift occurs the filter would vary the TLR until the impedance minima is close to the original position. This would help limit the need for capacitor replacement and engineering.

3.4 Economical Feasibility

To determine the economical feasibility of the proposed self-tuning filter scheme the basic equipment costs of the new filter is compared to the price of the regular passive filter alone. No estimate is made of the cost of the TLR construction or control mechanism. A unit price of $200/kVAr [32] is assumed for a single phase tuned harmonic filter. A TLR based harmonic filter therefore will have the same base price of the normal filter plus the cost of the thyristors required to do the switching. Cost equations for each type of filter are given in equations 3.12 and 3.13.

Price of a Regular Filter = (Rating kVAr)x($200/kVAr) (3.12)

Price of a Self-tuning Filter = (Rating kVAr)x($200/kVAr) + (Cost of Thyristors) (3.13)

Equation 3.13 can be further broken down to account for the number of TLR steps, the thyristors per step and the unit price of the thyristors.

$$\text{Cost of Thyristors} = (\text{Number of Thyristors Per Step}) \times (\text{Number of Steps}) \times (\text{Thyristor Unit Price}) \tag{3.14}$$

To find the unit cost of the thyristor it is necessary to calculate the required on state current and the worst case off state voltage. Once these values are determined a suitable device and its bulk unit price can be found[29]. To calculate the on state current it is assumed that the bulk of the current through the filter is reactive fundamental current being injected into the system. To factor in the effects of harmonic current is it safe to multiply the fundamental value by 1.2 to get a reasonable estimate of the filter current[30]. The fundamental impedance seen by the power system at the filter is

$$Z = j(X_l - X_C) \tag{3.15}$$

X_C can be calculated using the line voltage and the desired rating of the capacitor

$$X_C = \frac{(kV)^2}{MVAr} \tag{3.16}$$

The value of the reactive impedance is given by

$$X_L = \frac{X_C}{h^2} \tag{3.16}$$

where h is the tuned harmonic

The fundamental reactive current is then

$$I_1 = \frac{kV_{LN}}{X_C - X_L} \qquad (3.17)$$

The fundamental current can then be multiplied by the estimation factor of 1.2 to obtain an estimate of the total current flowing through the filter. A thyristor must be rated to carry this amount of current, if a suitable device cannot be found several thyristors whose sum of rated currents exceed the fundamental current may be used in parallel.

To calculate the maximum reverse voltage for a binary form 2 TLR the voltage drop across the largest inductor segment must be calculated. The largest inductor segment will have an impedance value of

$$X_L = \omega\left(2^{S-1} L_1\right) \qquad (3.18)$$

where S is the number of inductor segments

L_1 is the inductance value of the smallest inductor segment

If a suitable device cannot be found then several thyristors whose reverse breakdown voltage rating sum to the required value may be used in series. Table 3.4 contains the above calculations for a 10kVAr and 1000 kVAr rated filter with a 3 segment TLR and suitable thyristors for line voltages of 480V and 13.8kV. The final comparison of the passive filter and the self-tuning one is in the form of a percent cost increase

$$\%\ Cost = \frac{Self-tuning}{Passive} \times 100\% \qquad (3.19)$$

Table 3.4 Cost Ratios for a 3 Step Self-Tuning Filter

Rated kVAr	Line Voltage (kV)	Estimated Current (A)	Largest Reverse Voltage (V)	Thyristor Unit Price ($)	Passive Filter Cost ($)	Cost Ratio (%)
10	0.480	15.03	0.023	31.87	2000	19.1
1000	0.480	1503.52	2.3	212.96	200000	1.28
1000	13.8	52.30	0.079	57.41	200000	0.344

Since the capacitor carries the bulk of the impedance of the filter little, voltage drop is seen across the thyristors. Therefore the device chosen is based on it current rating. The material cost of the thyristors is negligible for large filters. Even with a large number of steps the overall price of the filter outweighs the cost of the thyristors. For small filters the extra equipment needed for construction of a self-tuning filter can be considerable.

3.5 Academic and Industrial Precedent

Self-tuning harmonic filters have been designed in [24,25] and implemented using variable inductors. In [24] the variable inductor is a tapped inductor controlled by a tap changing mechanism. The system is designed to compensate for capacitance changes due to ambient temperature variation and temperature variation due to self-heating. The filter and its control system is depicted in Figure 3.7. The control systems objective is to maintain the harmonic VAr (reactive power) injected by the filter between preset limits. Using potential, current transform signals proportional to voltage and current are measured. To calculate the reactive power the current is phase shifted by 90°and multiplied by voltage. The multiplier then produce and direct current output that is proportional to the value of the VAr injected by the filter. If this signal is greater than expected, ie. the filter has injected to many VArs, the inductance is increased by the tap changer relay conversely if the signal is too small the amount of inductance is decreased by the tap changer.

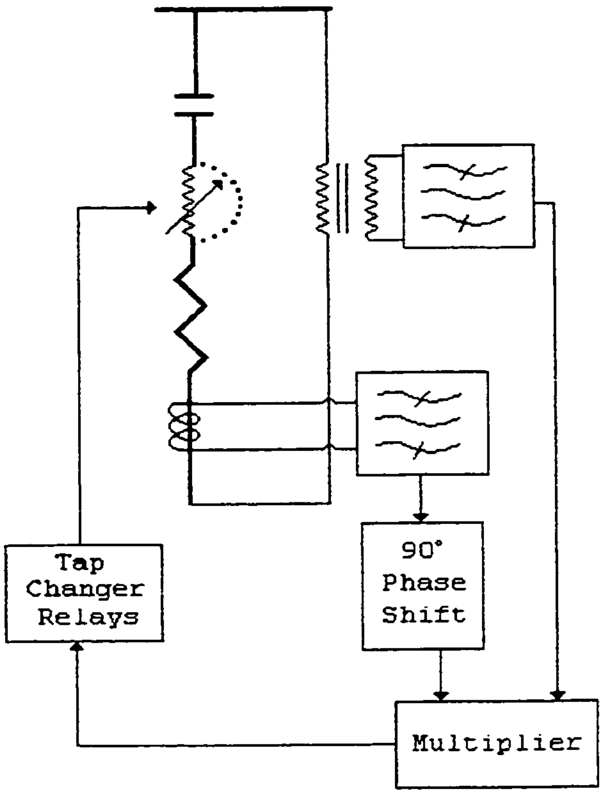

Figure 3.7 Self-Tuning Filter Using Tapchanger

An industrial application of the self-tuning filter technology is presented in [25]. ABB implements a variable inductor using a DC current to alter the permeability of the inductor core. The system is designed to optimize filter performance by maintaining specified filter tuning. The filter and its control system are depicted in Figure 3.8. Using closed loop control the system maintains the amount of harmonic current flowing through the filter. If the amount of harmonic current begins to exceed its limit the inductance value is changed so as to move the filter tuning frequency away from the harmonic. If there is room for more current to be shunted to ground the tuning point is moved closer to the harmonic.

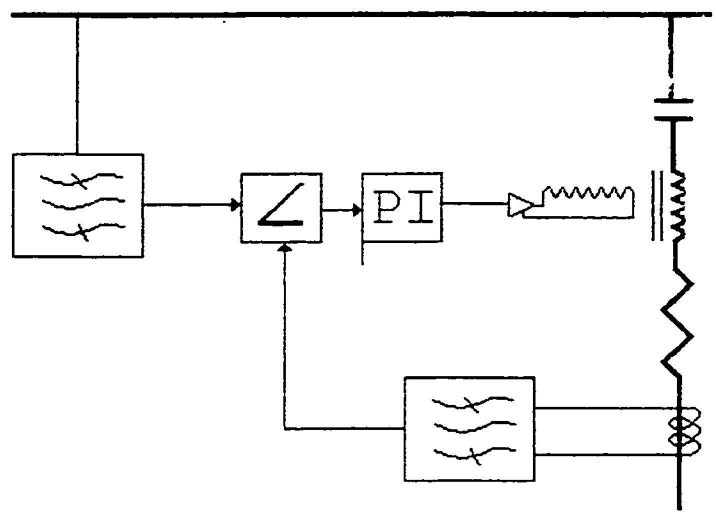

Figure 3.8 ABB Con-tune Filters using DC control Winding

3.6 Conclusion

In this chapter a novel self-tuning filter was presented. This method differs from the others referenced in that the variable element within the filter is a TLR or thyristor linked reactor. The TLR is a novel device that can be controlled electronically to step between inductance values of limit dynamic range and resolution. Since the TLR is electronically controlled and does not require a large secondary source it is a favorable device when compared to other method of producing a variable reactive impedance. Chapter 5 presents several methods for controlling the TLR within a self-tuning filter. The next chapter highlights a particular complication that arises from using the TLR. This complication is the creation of transients that occurs during TLR switching.

Chapter 4

Transient Characteristics of the Self-Tuning Filter

In chapter 3 the self-tuning filter using a thyristor linked reactor (TLR) was introduced. The goal of the present chapter is to demonstrate the feasibility of the TLR with respect to thyristor switching. To this end two cases will be studied using both analytic and numeric methods.

4.1 Introduction

To provide a complete description of any circuit containing reactive elements, the initial condition of each element must be known[16]. When in operation the self-tuning filter alters its reactive impedance to move its tuning frequency. If the system otherwise remains unchanged there will be a change in the steady state current through the filter. During transitions between configurations transients will occur. The voltage and current transients may propagate into the system and cause power quality problems. Therefore there is a necessity to asses the extent of the switching transients so that the usefulness of the TLR can be evaluated.

For the present study the system is assumed to be an ideal source with fixed impedance. The transients created by switch action will be estimated by Laplace transform techniques, time domain analysis as well as PSPICE and MATLAB numerical simulations. The circuits under study are presented in Figure 4.1. Both circuits have an ideal source with series system impedance. The circuit on the left (Figure 4.1a) has a single thyristor linked reactor connected to the source while the one on the right (Figure 4.1b) has a small but complete self-tuning filter.

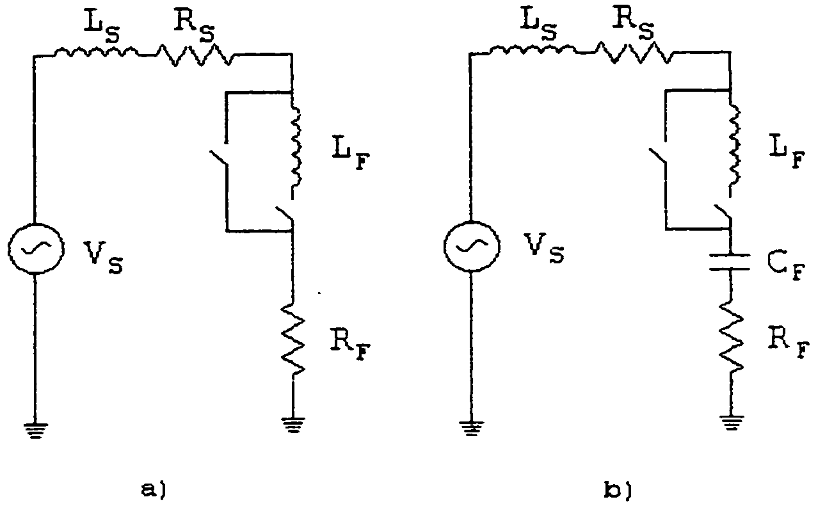

Figure 4.1 Test Circuits for Transient Analysis

The study of the first circuit will demonstrate how an individual reactor will behave under ideal switching. An understanding of this simple circuit will provide information necessary to understanding the behavior of the TLR. The second circuit a complete harmonic filter, will be used to show how the addition of capacitive energy storage element effects the transient behavior of the TLR.

4.2 Simple Case

In studies that follow the thyristors used in the switching circuits are assumed to be ideal. Figure 4.2 shows an ideal thyristor switch which has two states 'on' and 'off'. The thyristor switch that is on will have negligible forward impedance conversely when it is off it will have nearly infinite impedance. For analytic studies the thyristor will function as a switch.

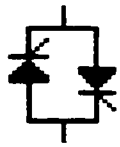

Figure 4.2 Ideal Thyristor Switch

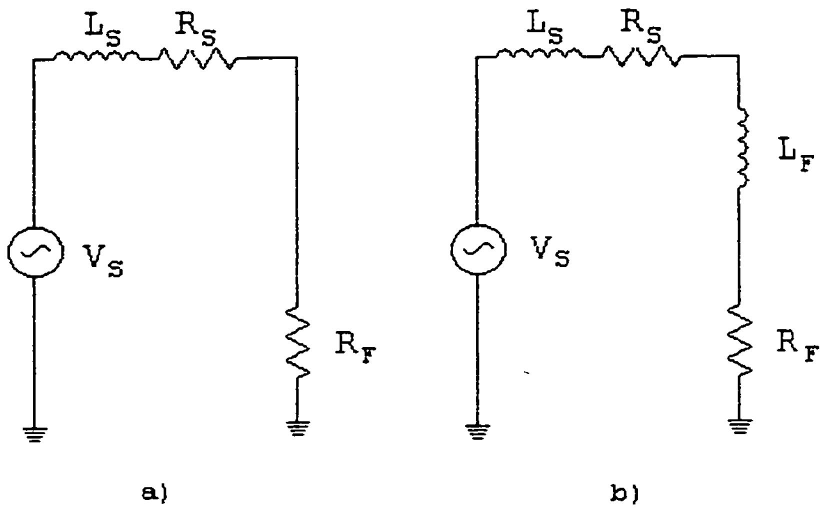

Figure 4.3 States of the Simple Case

The single TLR presented in Figure 4.1a has two switches a and b. The four different states of switches a and b lead to two different configurations. When switch a is closed as shown in Figure 4.3a the reactor is a short circuited conversely when a is open and b is closed the reactor contributes to the circuit impedance (Figure 4.3b). The matching time-dependant circuit equations are:

$$V_M \sin(\omega t + \phi) = (R_S + R_F) i(t) + (L_S) \frac{di(t)}{dt} \qquad (4.1)$$

and

$$V_M \sin(\omega t + \phi) = (R_S + R_F)i(t) + (L_S + L_F)\frac{di(t)}{dt} \qquad (4.2)$$

The general solution for the current can be written as:

$$i(t) = ce^{-\frac{R}{L}t} + \frac{V_M}{\sqrt{R^2 + \omega^2 L^2}}\cos\left(\omega t + \phi - \tan^{-1}\left(\frac{\omega L}{R}\right)\right) \qquad (4.3)$$

$$c = -\frac{V_M}{\sqrt{R^2 + \omega^2 L^2}}\cos\left(\phi - \tan^{-1}\left(\frac{\omega L}{R}\right)\right) \qquad (4.4)$$

where R is R_S+R_F - the total resistance

L is the total inductance of the circuit

ϕ is the phase angle of the source

ω is frequency

L_S is source reactive impedance

L_F is the filter reactive impedance

When the switching of a and b result in the reactor L_F being added to or removed from the circuit the time domain response must be recalculated using the new value of inductance. At the time of switching the current through the circuit becomes the initial condition of the new circuit. As presented in chapter 3 the ideal thyristor switches can be gated so that they change state at zero current. With the new circuit time is set to t=0 and the constant c is calculated using the circuit parameters before the time of switching.

The constant ϕ represents the phase shift between voltage and current when the switching occurs. For the current through the circuit to be zero before switching:

$$\omega t + \phi - \tan^{-1}\left(\frac{\omega L_0}{R}\right) = \frac{\pi}{2}, \frac{5\pi}{2}, \cdots \qquad (4.5)$$

where L_0 is the inductance before switching

The time of switching is arbitrary so say t=0:

$$\phi = \frac{\pi}{2} + \tan^{-1}\left(\frac{\omega L_0}{R}\right) \qquad (4.6)$$

The general solution for the magnitude of the transients for the case where L_F is switched into the circuit is then:

$$i(t) = ce^{-\frac{R}{(L_S+L_F)}t} + \frac{V_M}{\sqrt{R^2 + \omega^2 (L_S + L_F)^2}} \cos\left(\omega t + \phi - \tan^{-1}\left(\frac{\omega (L_S + L_F)}{R}\right)\right)$$

(4.7)

where

$$c = -\frac{V_M}{\sqrt{R^2 + \omega^2 (L_S + L_F)^2}} \cos\left(\frac{\pi}{2} + \tan^{-1}\left(\frac{\omega L_S}{R}\right) - \tan^{-1}\left(\frac{\omega (L_S + L_F)}{R}\right)\right)$$

(4.8)

$$\phi = \frac{\pi}{2} + \tan^{-1}\left(\frac{\omega L_S}{R}\right) \qquad (4.9)$$

The magnitude of the transients is determined by the circuit parameters both before and after the switching. When the amount of inductance L_F that is switched in is small compared to the amount already present L_S the transients will be small. Similar results are obtained from switching out inductance.

4.3 Full Case

The circuit presented in Figure 4.1b is a complete filter tunable to one of two frequencies. The circuit in both its configurations can be written in the time domain as:

$$V_M \sin(\omega t) = (R_S + R_F)i(t) + (L_{total})\frac{di(t)}{dt} + \frac{1}{C}\int_0^t i(\tau)d\tau + v_C(0) \qquad (4.10)$$

and in the frequency domain as:

$$\frac{V_M \omega}{s^2 + \omega^2} = (R_S + R_F)I(s) + L(sI(s) - i_L(0)) + \frac{I(s)}{sC} + \frac{v_C(0)}{s} \qquad (4.11)$$

where L is the total inductance of the circuit

$v_c(0)$ is the initial (t=0) voltage value across the capacitor

$i_L(0)$ is the initial current through the inductors

With different TLR configurations only the value of L will change. The introduction of the capacitive element has introduced another initial condition $v_c(0)$. This condition specifies the voltage across the capacitor at switching time. Unlike the inductors initial condition $v_c(0)$ is not zero because the voltage across the capacitor is not in phase with the current through it.

As a first step in understanding the behavior of this circuit a simplifying assumption is made. The voltage source is assumed to be constant V_s. Re-writing Equation 4.11 the frequency domain description of the circuit is:

$$\frac{V_S}{s} = (R_S + R_F)I(s) + L(sI(s) - i_L(0)) + \frac{I(s)}{sC} + \frac{v_C(0)}{s} \qquad (4.12)$$

solving for I(s)

$$I(s) = \frac{V_S - v_c(0) + sLi_L(0)}{s^2L + (R_S + R_F)s + \frac{1}{C}} \tag{4.13}$$

and performing the inverse Laplace transform an approximation of the time domain behavior of the current is:

$$i(t) = \frac{(V_S - v_C(0))}{bL}e^{-at}\sin(bt) + i_L(0)e^{-at}\cos(bt) + \frac{a}{b}i_L(0)e^{-at}\sin(bt) \tag{4.14}$$

where

$$a = \frac{(R_S + R_L)}{2L} \tag{4.15}$$

$$b = \sqrt{\frac{1}{LC} - \frac{(R_S + R_F)^2}{4L^2}} \tag{4.16}$$

From this it can be seen that the requirements for non-transient behavior are:

1) V_S-$v_C(0)$-aL$i_L(0)$=0.
2) $i_L(0)$=0

The voltage across the filter can be calculated in the time domain as:

$$V_f(t) = V_S\sin(wt) - R_Si(t) - L_S\frac{di(t)}{dt} \tag{4.17}$$

where i(t) is the current through the filter with $i_L(0)$=0.

Again assuming that the source is constant and substituting the expression for i(t):

$$v(t) = V_S - \frac{(V_S - v_C(0))}{bL}\left((R_S - aL_S)e^{-at}\sin(bt) + bL_S e^{-at}\cos(bt)\right) \qquad (4.18)$$

For the voltage transients to be zero the same initial conditions are required as for the current transients.

The second requirement is already met because the thyristors switch off only when i_L=0. The first requirement is met if the voltage across the capacitor is equal to the voltage across the source at the time of switching. The voltage V_C across the capacitor is given by the phasor:

$$V_C = \frac{V_S}{(R_S + R_F) + j\left(\omega L - \frac{1}{\omega C_F}\right)} \frac{1}{j\omega C_F} \qquad (4.19)$$

in actual application

$$(R_S + R_F) \approx 0 \qquad\qquad j\omega L << \frac{1}{j\omega C}$$

and therefore

$$v_C(0) \approx V_0$$

That is, the condition V_S-$v_C(0)$-$aLi_L(0)$=0 can be met approximately. As a result the current and voltage transients will not be significant if the reactor is switched at $i_L(0)$=0. This is always the case due to the SCR operating characteristics.

To gain more understanding otf the magnitude of the transients further analysis of the equation 4.10 is performed in the time domain. Differentiating Equation 4.10 gives

$$V_M \omega s \cos(\omega t + \phi) = L\frac{d^2 i(t)}{dt^2} + (R_S + R_F)\frac{d}{dt}i(t) + \frac{1}{C}i(t) \qquad (4.20)$$

The general solution for this second order differential equation is given in the time domain as:

$$i(t) = c_1 e^{rt} + c_2 e^{st} + K\cos(\omega t + \phi - \delta) \qquad (4.21)$$

where

$$K = \frac{V_M \omega}{\Delta} \qquad (4.22)$$

$$\sin(\delta) = \frac{(R_S + R_F)\omega}{\Delta}$$

$$r = -\alpha + \sqrt{\alpha^2 - \omega_0^2}$$

$$\cos(\delta) = \frac{L(\omega_0^2 + \omega^2)}{\Delta}$$

$$\Delta = \left(L^2(\omega_0^2 - \omega^2)^2 + (R_S + R_F)^2\omega^2\right)^{1/2}$$

$$s = -\alpha - \sqrt{\alpha^2 - \omega_0^2}$$

$$\alpha = \frac{R_S + R_F}{L}$$

$$\omega_0 = \frac{1}{\sqrt{LC}}$$

In this analysis the complete circuit in Figure 4.1b is switched from one value of inductance to another at time $t=t_{sw}$. Before $t=t_{sw}$ the circuit has reached steady state so the current through the circuit can be described as:

$$i(t) = K\cos(\omega t_{sw} - \delta) \tag{4.23}$$

Due to the thyristor switching this current will be zero at the time $t=t_{sw}$. This can only be true when one of the following conditions is met:

$$\omega t_{sw} - \delta = \frac{\pi}{2}, \frac{3\pi}{2}, \cdots \tag{4.24}$$

At the time $t=t_{sw}$ the current is zero so the voltage across the capacitor is the difference of the voltage generated by the source and the voltage across the inductor:

$$v_C(\omega t_{sw}) = V_M \sin(\omega t_{sw}) + LK\omega\sin(\omega t_{sw} - \delta) \tag{4.25}$$

where L is the inductance of the circuit before switching

After a switch is closed, changing the value of the total inductance to L_N, the current through the circuit is described as:

$$i(t) = c_{1N}e^{r_N t} + c_{2N}e^{s_N t} + K_N \cos(\omega t - \delta_N + \omega t_{sw}) \tag{4.26}$$

where r_N, s_N, K_N, δ_N are the "N"ew values of the indicated symbol with the changed value of L_N.

The second circuit evolves from t=0 with the initial conditions:

1) $i_L(0) = 0$
2) $v_C(0) = V_M\sin(\omega t_{sw}) + LK\omega\sin(\omega t_{sw} - \delta)$

and a phase shift of ωt_{sw} in the steady state term. The phase shift occurs because the forcing sinusoidal voltage is out of phase with the steady state current through the circuit. When the thyristor switching occurs there will be voltage across the forcing source.

To determine the coefficients c_{1N} and c_{2N} the initial conditions are used. For current:

$$i(0) = 0 = c_{1N} + c_{2N} + K_N \cos(-\delta_N + \omega t_{sw}) \tag{4.27}$$

For the voltage across the capacitor the general expression relating the new transient coefficients to the initial conditions is:

$$\begin{aligned} &V_M \sin(0 + \omega t_{sw}) - (R_S + R_F) i(0) - L_N \frac{di(0)}{dt} \\ &\quad = v_C(0) = V_M \sin(\omega t_{sw}) + LK\omega \sin(\omega t_{sw} - \delta) \\ &\quad = V_M \sin(\omega t_{sw}) - L_N\left(c_{1N} r_N + c_{2N} s_N + K_N \sin(\omega t_{sw} - \delta_N)\right) \end{aligned} \tag{4.28}$$

The term $V_M \sin(\omega t_{sw})$ cancels out because at the time of switching this term is common to the initial condition and the voltage loop about the capacitor. This leaves the relation as the second equation required to calculate the two transient coefficients as:

$$\begin{aligned} &L_N\left(c_{1N} r_N + c_{2N} s_N + K_N \sin(\omega t_{sw} - \delta_N)\right) \\ &\qquad = L\, K\omega \sin(\omega t_{sw} - \delta) \end{aligned} \tag{4.29}$$

Making substitutions Equations 4.28 and 4.29 can be written as

$$0 = c_{1N} + c_{2N} + A \qquad\qquad B = c_{1N}D + c_{2N}E + F$$

where

$$A = K_N \cos(\omega t_{sw} - \delta_N) \tag{4.30}$$

$$B = KL\omega \sin(\omega t_{sw} - \delta) \tag{4.31}$$

$$D = -L_N r_N \tag{4.32}$$

$$E = -L_N s_N \tag{4.33}$$

$$F = K_N \omega \sin(\omega t_{sw} - \delta_N) \tag{4.34}$$

solving for c_{1N} and c_{2N}

$$c_{1N} = \frac{B - F + AE}{D - E} \tag{4.35}$$

$$c_{2N} = -A - c_{1N} \tag{4.36}$$

The voltage and current will be out of phase by less than 180° so

$$\omega t_{sw} - \delta = n\frac{\pi}{2} \qquad\qquad n = 1,5,9\cdots \tag{4.37}$$

for n=1

$$\omega t_{sw} = \frac{\pi}{2} + \delta \tag{4.38}$$

A (Equation 4.30) can be re-written as:

$$A = K_N \cos(\frac{\pi}{2} + \delta - \delta_N) = -K_N \sin(\delta - \delta_N) \qquad (4.39)$$

Finally the constants c_{1N} and c_{2N} can be written as:

$$c_{1N} = \frac{KL_S\omega \sin(\omega t_{sw} - \delta) - L_T K_N \omega \sin(\omega t_{sw} - \delta_N) + L_T s_N K_N \sin(\delta - \delta_N)}{-L(r_N - s_N)} \qquad (4.40)$$

$$c_{2N} = K_N \sin(\delta - \delta_N) - c_{1N} \qquad (4.41)$$

The expression for current transients are now complete using Equations 4.40 and 4.41 above in the general expression for current 4.26. This expression for current can now be used to find the voltage transients across the filter. The voltage across the filter can be written as.

$$V_f(t) = V_S \sin(wt + \omega t_{sw}) - R_S i(t) - L_S \frac{di(t)}{dt} \qquad (4.42)$$

Substituting Equation 4.26 for i(t) the voltage across the filter can be written as:

$$V_f(t) = [V_S \sin(wt + \omega t_{sw}) - R_S K_N \cos(wt + \omega t_{sw} - \delta_N) - K_N \omega L_S \sin(wt + \omega t_{sw} - \delta_N)]$$

$$+ \left[- R_S \left(c_{1N} e^{r_N t} + c_{2N} e^{s_N t}\right) - L_S \left(c_{1N} r_N e^{r_N t} + c_{2N} s_N e^{s_N t}\right)\right] \qquad (4.43)$$

The first bracket terms are the steady state response while the second bracketed terms are the transient response.

As an analysis the worst case magnitude of the transients are calculated as a percentage of the steady state value. These calculations are used in the next section to numerically demonstrate the magnitude of transients. This calculation is performed for system inductance that changes from 40% through 220% of their original magnitude. For the current transients through the filter the worst case expressed as a percentage of steady state is:

$$I_{trans}(L_T, R_T, \omega, C_F, V_M) = \frac{|c_{1N}| + |c_{2N}|}{K_N} = \frac{\Delta(|c_{1N}| + |c_{2N}|)}{V_M \omega} \tag{4.44}$$

For the voltage across the filter the steady state magnitude is calculated as:

$$\begin{aligned} |V_{FSS}| &= \left| I\left(j\omega L_F - \frac{j}{\omega C_F} \right) \right| = \left| \frac{V_M}{R_T + j\omega L_T - \dfrac{j}{\omega C_F}} \left(j\omega L_F - \frac{j}{\omega C_F} \right) \right| \\ &= \frac{V_M \omega}{\Delta} \left(\omega L_F - \frac{1}{\omega C_F} \right) \end{aligned} \tag{4.45}$$

A key feature to note is that when $L_T \rightarrow L_S$, $K_N \rightarrow K$ and $\delta - \delta_N \rightarrow 0$ as a result small changes in the system inductance will give rise to relatively small transients coefficients.

With the analytic representations of the transients created by TLR switching complete a numerical assessment of the their magnitude with system values obtained from Chapter 2 test system is now conducted.

4.4 Numerical Assessment

In this section using MATLAB the magnitude of the transient coefficients are calculated with respect to changes in system inductance using the simple and full case equations presented in the previous section. These calculations are augmented by simulations using PSPICE which show the response of the circuits to switching. Both sets of numerical calculations use the same basic system and filter values.

$$R_S = 1.728 m\Omega$$

$$L_S = 28.543 \mu H$$

$$L_F = 33.1572 \mu F$$

$$R_F = 0$$

$$C_F = 9.21036 mF$$

Figure 4.4 shows the effect on transient magnitude of switching inductance in and out of the circuit in Figure 4.1. The graph shows the transients magnitudes as a percentage of the steady state value of the current after switching. The graphs are plotted versus the post to pre inductance ratio of the circuit. The figure can be viewed as having two sections separated by the point of no inductance change. On the left side inductance is removed from the circuit during switching as a result the Post to Pre inductance ratio is less than 1. To the right the act of switching adds inductance so the Post to Pre ratio is greater than 1. Figure 4.4 shows that transient are zero when the inductance switched in or out of a circuit is small percentage of the pre-switching value.

Graphs of the worst case current and voltage transients for the full filter case are presented in Figures 4.5 and 4.6. These graphs were generated using MATLAB and Equations 4.44 and 4.45. The graphs are plotted as a function of percent change in system inductance caused by switching and show the worst case transients that can arise. The left hand side of Figure 4.5 shows an increase in current transient magnitude when filter inductance is removed from the system. When

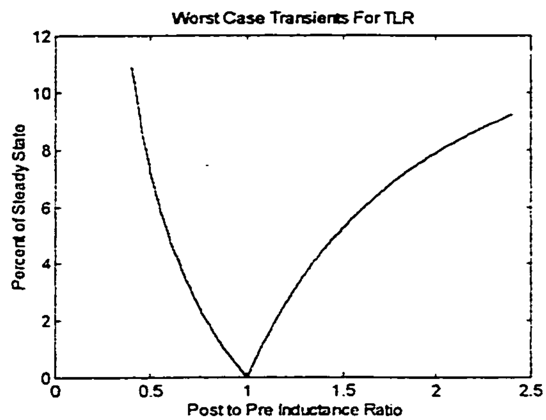

Figure 4.4 Current Transients Caused by TLR Switching

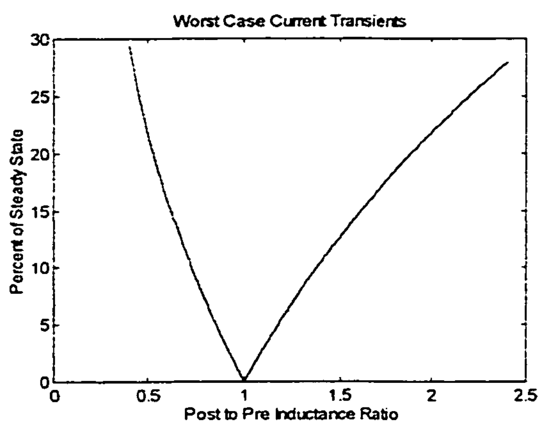

Figure 4.5 Current Transients In Self-Tuning Filter

the change in inductance is small compared to the system inductance small transients are seen. If large step changes are taken the transients created can have larger magnitudes. The right hand side of Figure 4.5 shows the current transient magnitudes when filter inductance is removed from the system. When a small decrease is made in the filter inductance, small transients occur; when large values of inductance are switched the transients increase. Similar results are shown in Figure 4.6 for voltage across the filter. The range of the X-axis is selected to include both the switching in and switching out of the entire filter inductance. The Y-axis represents the magnitude of the worst case transients created by the switching.

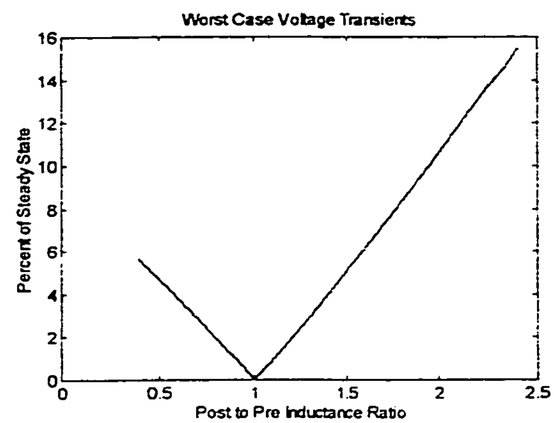

Figure 4.6 Voltage Transients In Self-Tuning Filter

As a final demonstration of the magnitude of the transients created during switching two, simulations were run using PSPICE. For both simulations the current through the filter and the voltage across it were calculated as a function of time. Again switching was done using near-ideal voltage controlled switches timed manually so as to switch when current was zero. For the first set of simulations the total inductance was changed from L_S+L_F to L_S very close to the 1 second mark. This results in a post to pre inductance ratio of ~0.46. For the second set the total inductance was changed from L_S to L_S+L_F. The resulting ratio is ~2.16. In each set of simulations both the simple and the full filter cases were tested. The results of the simulations are present in Waveforms 4.7 through 4.12, the percent overshoots and their analytically calculated maximums are presented in Table 4.1

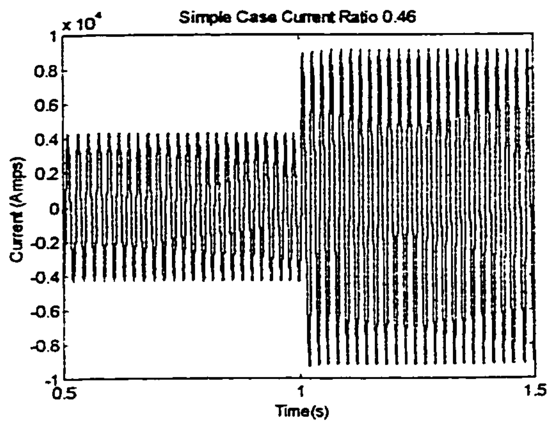

Figure 4.7 Current Waveform For the Simple Case with 0.46 Reactance Change Ratio

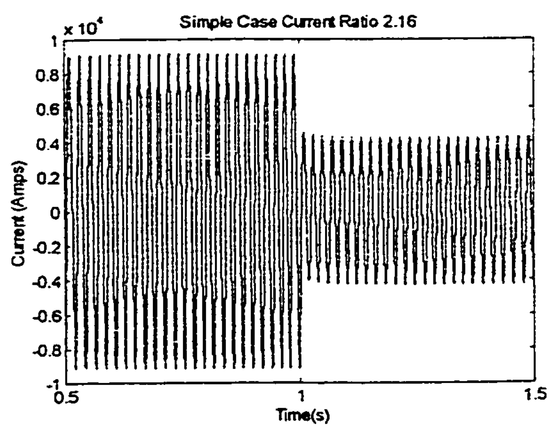

Figure 4.8 Current Waveform For the Simple Case with 2.16 Reactance Change Ratio

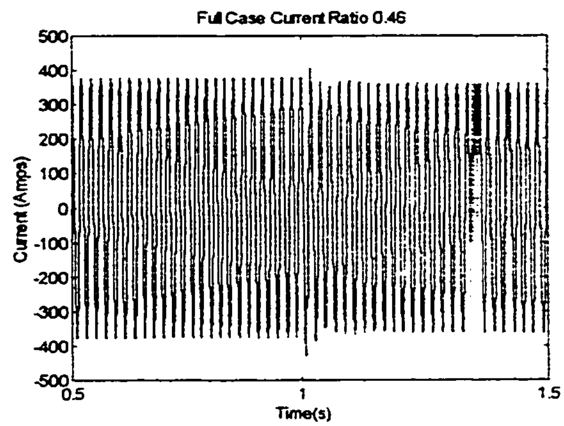

Figure 4.9 Current Waveform For the Full Case with 0.46 Reactance Change Ratio

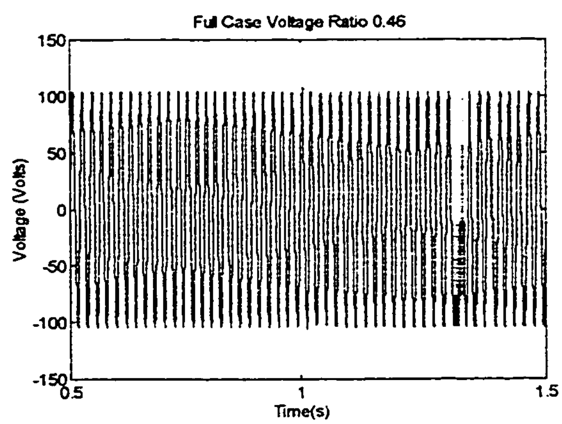

Figure 4.10 Voltage Waveform For the Full Case with 0.46 Reactance Change Ratio

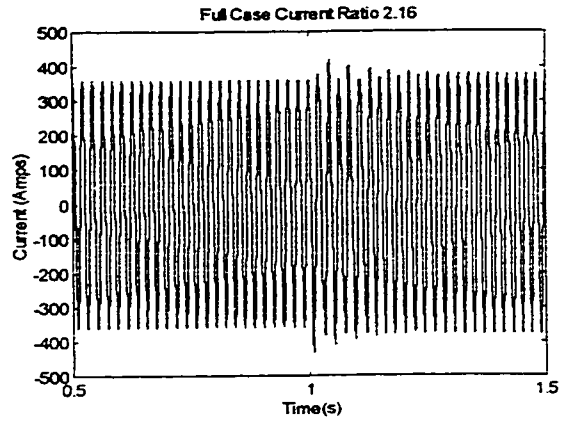

Figure 4.11 Current Waveform For the Full Case with 2.16 Reactance Change Ratio

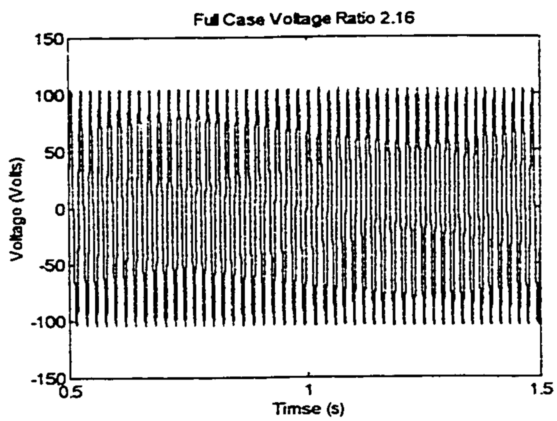

Figure 4.12 Voltage Waveform For the Full Case with 2.16 Reactance Change Ratio

Table 4.1 Worst case Calculated and Simulated Transient Peaks

Figures	Post to Pre Inductance Ratio	Maximum Transient Peak (% of SS)	Worst Case Calculated (%)
4.4 & 4.7	0.46	2.56	8.59
4.4 & 4.8	2.16	7.52	8.5
4.5 & 4.9	0.46	19.0	24.6
4.6 & 4.10	0.46	3.85	5.08
4.5 & 4.11	2.16	14.5	24.4
4.6 & 4.12	2.16	1.94	12.5

4.5 Conclusions

In this chapter it has been demonstrated that a self-tuning filter using a TLR as its variable reactive device will produce transients during switching. It was also shown that as the switching steps get smaller so do the transients created. As a result if the self-tuning filter is required to compensate for small variations in filter reactance the TLR based self-tuning filter would be a reasonable choice.

Chapter 5

Operation and Control of the Self-Tuning Filter

In chapter 4 the feasibility of the TLR was demonstrated by showing that the conditions for minimizing transients during switching could be met. Switching of the self-tuning filter is required to adjust the reactive impedance of the filter to meet filter specifications for tuning frequency or current. In this chapter three methods for determining the required reactive impedance adjustment are presented.

5.1 Introduction

A complete description of TLR control under dynamic harmonic content is out of the scope of this work. Instead two simplifying assumptions will be made:

1) System harmonic content varies slowly with time
2) All switching decision are made during steady state

With these assumptions the reactive impedance adjustment can be calculated from instantaneous values.

Three methods are presented in this chapter:

1) Direct Method
2) Current Limiting Method
3) Parameter Estimation Method

All three methods use harmonic voltage or current data to estimate a reactive impedance adjustment required for meeting filter specifications. The direct method requires accurate RMS measurements of voltage, current and phase angle at the harmonic frequency being filtered as well as knowledge of the desired tuning frequency and an estimate of the present values of the filter capacitance and inductance. The current limiting method requires accurate values of the RMS current and qualitative information about the voltage leading or lagging. The current limiting method then uses this information to increase or decrease the TLR inductance. The third method presented is the parameter estimation method. This method uses the voltage and current content at several significant harmonic frequencies to estimate present filter parameters. From the calculation of filter impedance the TLR adjustments necessary to achieve the required tuning frequency are made. In the following sections each method is presented with analytic methods, qualitative function and quantitative examples.

5.2 Direct Method

This method uses the TLR to adjust the filters reactive impedance to maintain a specific tuning frequency. Capacitor parameter variation due to temperature or damage can cause changes in the phase angle between voltage and current. If the filter resistance remains unchanged the relation ship between γ_h and the reactive impedance can be calculated as:

$$\frac{V_h\angle 0}{I_h\angle\gamma} = Z_h\angle -\gamma_h = R + j(hX_L - \frac{X_C}{h}) \qquad (5.1)$$

where V_h is the voltage harmonic across the filter

I_h is the harmonic current through the filter

γ_h is the phase angle between harmonic voltage and current

R is the resistance of the filter

X_L is the fundamental inductive reactance of the filter

X_C is the fundamental capacitive reactance of the filter

h is the harmonic being shunted

Using Euler's formula the angular component can be equated to the reactive impedance and can be written as follows

$$\frac{hX_L - \frac{X_C}{h}}{|Z_h|} = \frac{hX_L - \frac{X_C}{h}}{\left|R + j(hX_L - \frac{X_C}{h})\right|} = \sin(\gamma_h) \qquad (5.2)$$

Due to a fluctuation in some physical parameter either through temperature change or capacitor damage the reactance of the filter changes. Equation 5.2 must be re-written to include a change in capacitance ΔX_C.

$$hX_L - \frac{1}{h}(X_C + \Delta X_C) = \left|R + j\left(hX_L - \frac{1}{h}(X_C + \Delta X_C)\right)\right| \sin(\gamma_h + \Delta\gamma_h) \qquad (5.3)$$

where ΔX_C is the change in fundamental capacitve reactance

The previous equation has been arranged with measured values of $|Z|$ and γ_h on the right hand side and the resultant reactive impedance on the left hand side. The value of ΔX_L can be calculated from ΔX_C:

$$\Delta X_L = \frac{\Delta X_C}{h^2} \qquad (5.4)$$

where ω is the fundamental frequency

ΔX_L is the required change in inductive reactance

To calculate the value of ΔX_L the values of X_L, X_C, R, ω, γ_h and $|Z_h|$ are required. γ_h and $|Z_h|$ are assumed to be measurable and R and ω are assumed fixed. X_L and X_C are values that are known initially and updated at every switching event. For switching to be possible the magnitude of ΔX_L must be larger than half the smallest TLR step. Further $X_L + \Delta X_L$ must be within the TLR physical capacity. The number of minimum steps that have to be made to change

X_L by ΔX_L can be calculated by dividing ΔX_L by the minimum step size of the TLR and following normal integer rounding rules. If these conditions are not met switching does not occur. The procedure for making TLR adjustments using the direct method is outlined in Figure 5.1.

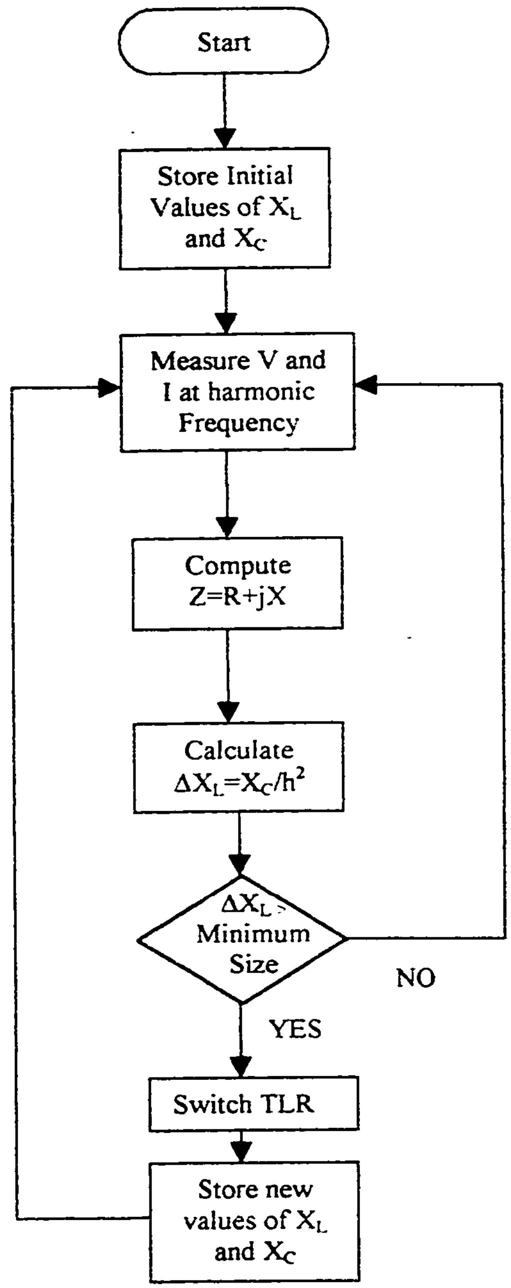

Figure 5.2 Flow Chart For the Direct Method

The direct method makes assumptions about measurable values. It is assumed that both voltage and current phasors can be measured at the tuning frequency. The obstacle for this assumption is that the filter will have low impedance for the harmonic being measured. Therefore it is probable that accurate voltage information will not be available at the harmonic being filtered. Two solutions to this problem will now be discussed. The first relies only on the value of the current for tuning decisions and the second employs a best-fit method for estimating the value of the capacitor.

5.3 Current Limiting Method

The current limiting method bases its reactive impedance changes on whether the filter current is above or below the rated filter current. If the filter is carrying current exceeding the filter specifications the filter impedance must be increased. If the current is below the specification there is room to increase the harmonics being shunted by decreasing the filter impedance. In essence, the filter attempts to control the current in a narrow band as shown in Figure 5.2.

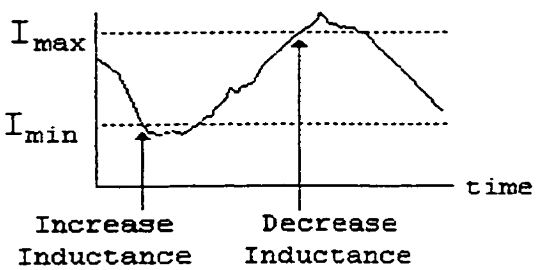

Figure 5.2 Current Limiting TLR Control Scheme

Since changing the filter reactance will change its impedance at all frequencies the filter current must be specified as the total RMS current as specified by:

$$I_{Total} = \sqrt{\sum_{h=1}^{k} I_h^2} \qquad (5.5)$$

where h is the harmonic order up to the k^{th}

At each switching event ΔX_L will always be the minimum step size. This choice is justified by the fact that slow variations in harmonic content may be compensated with a series of small steps. The sign of ΔX_L (+/-) can be determined from whether the harmonic current leads or lags the harmonic voltage. The filter impedance at any harmonic h is given by:

$$Z_{filter} = \sqrt{R^2 + \left(hX_L - \frac{X_C}{h}\right)^2} \qquad (5.6)$$

As a result:

- If I_h lags V_h, the filter is reactive which implies:

$$hX_L > \frac{X_C}{h} \qquad (5.7)$$

Thus decreasing X_L will decrease Z_{filter}.

- If I_h leads V_h, the filter is capacitive which implies:

$$hX_L < \frac{X_C}{h} \qquad (5.8)$$

Thus increasing X_L will decrease Z_{filter}.

The information about leading or lagging current may be gleaned from the crossing points of the harmonic voltage and current to be filtered. A flow chart of the TLR switching control for the current limiting method is present in Figure 5.3.

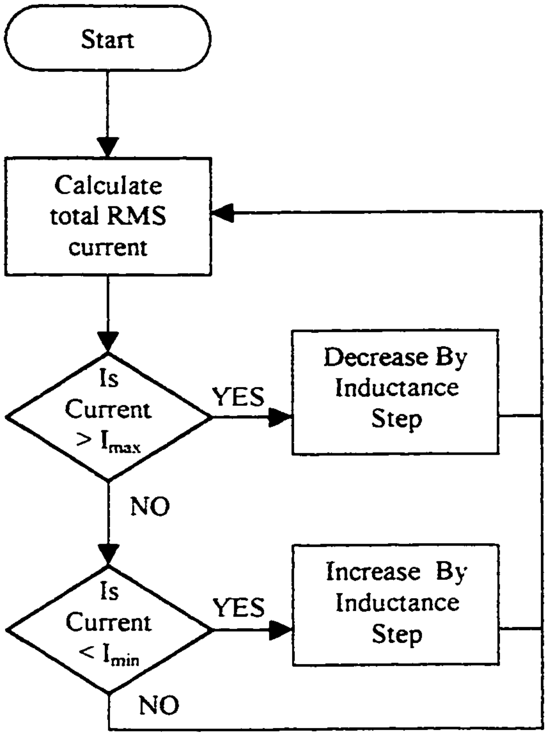

Figure 5.3 Flow Chart for Current Limiting Method

The current limiting method relies on accurate information about significant harmonic currents while only requiring leading or lagging information about harmonic voltage. Since the method limits filter current to a specified values, it can minimize damage to the filter due to long term harmonic current variation. The method is limited by the lack of information in calculating ΔX_L magnitude. The last method to be presented removes this limitation by using information from several significant harmonic voltages and currents to model system parameter and to calculate ΔX_L.

5.4 Parameter Estimation Method

The parameter estimation method uses harmonic voltage and current data to calculate the filter impedance frequency response. This data is used to find a best fit to the model:

$$Z = R + j\omega L - \frac{j}{\omega C} \qquad (5.9)$$

If the value of the capacitance C has varied significantly the TLR is switched so as to compensate for the change. The information required by the parameter estimation method is the harmonic voltages and currents at frequencies other than the one being filtered and the specified tuning frequency. The method used to fit the data to the model is simultaneous solution of linear equations.

The first step in calculating the values of L, C and R is to gather the required data. The impedance frequency response can be calculated at each harmonic (including the fundamental) where V and I can be accurately measured. The filtered frequency is ignored because of the possible inaccuracy of voltage measurement at that frequency. For each pair of voltage and current measurement the impedance can be written as:

$$\frac{V_h}{I_h} = Z_h = R + jhX_L - \frac{j}{hX_C} \qquad (5.10)$$

where h is the harmonic frequency

These equations can be written in matrix form:

$$\begin{bmatrix} Z_1 \\ \vdots \\ Z_h \end{bmatrix} = \begin{bmatrix} 1 & j1 & -j1 \\ \vdots & \vdots & \vdots \\ 1 & jh & -\frac{j}{h} \end{bmatrix} \begin{bmatrix} R \\ X_L \\ X_C \end{bmatrix} \qquad (5.11)$$

or simply

$$Y = AX \qquad (5.12)$$

Calculating the pseudo-inverse, the solution is :

$$X = \left(A^T A\right)^{-1} A^T Y \qquad (5.13)$$

The value of ΔX_L that the TLR must be adjusted by can be gathered from:

$$\Delta X_L = \frac{X_C}{h^2} - X_L \qquad (5.14)$$

The TLR can then be switched to the specified frequency. A flow chart of the parameter estimation method can be found in Figure 5.4. The parameter estimation technique requires only three sets of voltage and current data and the specified tuning frequency.

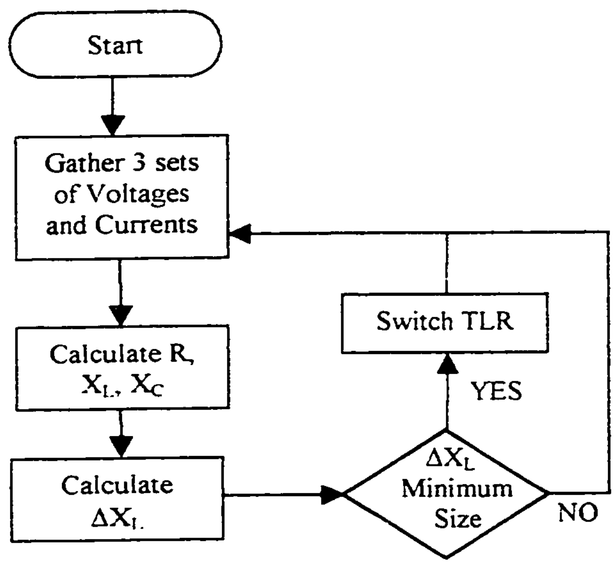

Figure 5.4 Flow Chart for Parameter Estimation Method

5.5 Numerical Examples

In this section examples of the direct, current limiting and parameter estimation methods are given. The three examples are all based on the case study presented at the end of chapter 2. As in that study an adjustable speed drive (ASD) is present within the system. It acts as a non-linear load and thus as a source of harmonics. The values for the harmonic currents generated by the ASD in relation to the fundamental and the relative phase angles were presented in Table 2.4. It was also shown how a harmonic filter tuned near to the fifth harmonic could reduce the total harmonic distortion (THD) at buses and branches within the system. The main focus being on the level of voltage THD at the utility bus.

In chapter 2 it was demonstrated that when the harmonic filter experienced a 5% increase in its capacitance the maximal 1% THD limit at the utility bus was exceeded.

The filter parameters and the resulting voltage THD at the utility bus before and after the fluctuation are given in Table 5.1.

Table 5.1 Filter Parameters Before and After with Voltage THD

Parameter	Before Fluctuation	After Fluctuation
L	0.03315mH	0.03315mH
C	9210.36uF	9670.9uF
Harmonic	4.80	4.68
Voltage THD at UTIL-69	0.940%	1.144%

The goal of the examples in this section is to show how the change in tuning of the filter can be corrected if the inductor L is a thyristor linked reactor (TLR). In each example the TLR is switched so as to re-tune the filter. The TLR used has 4 switch-able segments giving a linear distribution of 16 inductor values in a range +/-10% of the central value. The minimum step size and range values are given in Table 5.2.

Table 5.2 TLR Parameters

Parameter	Value
Minimum Step Size	0.41441uH
Inductance range	36.05807μH to 29.8414mH

Examples of the direct, current limiting and parameter estimation methods were developed using MATLAB and PSA-H. The fundamental phasor voltage at the ASD bus and the current through the filter for various inductance and capacitance values was determined using PSA-H. Using an equivalent circuit involving only the source impedance, the filter and the ASD (Figure 5.5) the MATLAB Script 3 presented in Appendix A was used to calculate the harmonic phasor voltages and currents of the filter. The system impedance is given in Table 5.3.

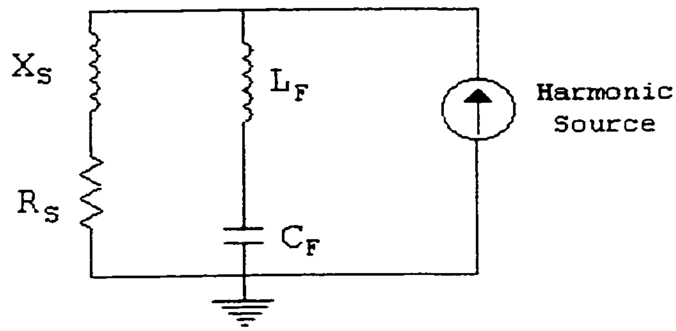

Figure 5.5 System/Filter Model for Parameter Estimation Technique

Table 5.3 System Parameters

Parameter	Value
R_S	0.001728Ω
X_S	28.543uH

The equivalent circuit functions as a current divider. Changes in the filter impedance produce changes in the voltage across and the current through the filter at each harmonic. The most dramatic changes occur at the fifth harmonic because of the sharp tuning of the notch filter. Selected harmonic voltage and currents across and through the filter before and after the fluctuations are presented in Table 5.4 and 5.5.

Table 5.4 Currents Before and after capacitor fluctuations

Harmonic Number	Current Before (pu)	Current Angle Before (Deg)	Current After (pu)	Current Angle After (Deg)
1	0.0081021	80.6500	0.0085415	80.6500
5	0.0055253	-35.0414	0.0053431	-36.9075
7	0.0018749	-63.0967	0.0018321	-63.1631
11	0.0007936	-119.9695	0.0007872	-119.9729
13	0.0005417	-151.8158	0.0005384	-151.8175

Table 5.5 Voltage Before and After Capacitor Fluctuations

Harmonic Number	Voltage Before (pu)	Voltage Angle Before (Deg)	Voltage After (pu)	Voltage Angle After (Deg)
1	0.9688	-9.35	0.9705	-9.35
5	0.0117505	54.9586	0.0177238	53.0925
7	0.0377224	26.9033	0.0384210	26.8369
11	0.0383440	-29.9695	0.0384609	-29.9729
13	0.0329994	-61.8158	0.0330455	-61.8175

5.5.1 Direct Method Example

After the fluctuation the total impedance is calculated by dividing magnitudes of the voltage and current at the fifth harmonic from Tables 5.4 and 5.5:

$$hX_L - \frac{1}{h}(X_C - \Delta X_C) = |Z|\sin(\gamma) = \frac{0.0177238pu}{0.0053431pu} = 3.3171\Omega_{pu} \qquad (5.15)$$

X_L and X_C are the reactive impedance of the filter before fluctuation. These values are known and updated each time a change takes place. Z and γ are the impedance of the filter and the relative phase angle of voltage and current of the filter. The impedance is converted from per-unit to an impedance value so that the results of the calculation will be in Farads and Henrys:

$$3.3171\Omega_{pu}\left(\frac{.480^2 kV}{100MVA}\right) = 7.6427x10^{-3}\Omega \qquad (5.15)$$

The term in brackets is the reciprocal of base impedance of the system at the bus in question. Since there is no resistance, the impedance of the filter is purely reactive and furthermore is dominated by the capacitor therefore $\sin(\gamma) = 1$. The equation 5.15 is reordered solving for ΔX_C, the deviation of the capacitor in Farads:

$$\frac{-\Delta X_C}{h} = 0.0076427 - hX_L + \frac{X_C}{h} = -0.0027428 \qquad (5.17)$$

the change in inductance required is calculated as

$$\Delta L = \frac{\Delta X_C}{h^2 \omega} = -1.4551 \mu H \qquad (5.18)$$

Diving this value by the minimum step size and following normal rounding rules give the total number of minimum TLR steps necessary to correct for the capacitor fluctuation.

$$\frac{-1.4551 \mu H}{0.41441 \mu H} = -3.5188 \rightarrow -4 \quad steps \qquad (5.19)$$

The sign determines whether the inductance should be increased or decreased in this case the inductance value is decreased by four steps or 1.65764μH the parameters for the filter before and after the switching are presented in Table 5.6.

Table 5.6 Filter Parameter Before and After TLR Switching

Parameter	Before Fluctuation	After Fluctuation
L	0.03315mH	0.03150mH
C	9670.9uF	9670.9uF
Harmonic	4.68	4.81

As a result of the finite switching steps the filter is not tuned to exactly the same the harmonic. Tables 5.7 and 5.8 show the harmonic voltages and currents before the fluctuation in the capacitor and after the TLR is switched.

Table 5.7 Currents Before and after TLR switching

Harmonic Number	Current Before (pu)	Current Angle Before (Deg)	Current After (pu)	Current Angle After (Deg)
1	0.0081021	80.6500	0.0085203	80.6500
5	0.0055253	-35.0414	0.0055405	-34.7728
7	0.0018749	-63.0967	0.0019227	-63.0108
11	0.0007936	-119.9695	0.0008157	-119.9540
13	0.0005417	-151.8158	0.0005571	-151.8045

Table 5.8 Voltage Before and After TLR switching

Harmonic Number	Voltage Before (pu)	Voltage Angle Before (Deg)	Voltage After (pu)	Voltage Angle After (Deg)
1	0.9688	-9.35	0.9704	-9.35
5	0.0117505	54.9586	0.0108673	55.2272
7	0.0377224	26.9033	0.0366714	26.9892
11	0.0383440	-29.9695	0.0374201	-29.9540
13	0.0329994	-61.8158	0.0322247	-61.8045

Further the THD at the utility bus has dropped to 0.898%, below the required 1% value. To calculate the reactance change the direct method uses the both the phasor voltage and the current at the harmonic being filtered. Since the harmonic filter is tuned such that is has a low impedance to the harmonic currents to be filtered accurate voltage information may not be known to sufficient accuracy. The current limiting method presented next assumes that only limited phase information of the harmonic voltage is known.

5.5.2 Current Limiting

The fluctuation in the capacitor changes the amount of 5^{th} harmonic current flowing through the filter. Using an established 5^{th} harmonic limit the TLR takes minimal switching steps until the limit is met. For this example the limit will be selected as any current value within 1% of the value before fluctuation. The decision to increase or decrease the reactance is based on the amount of 5^{th} harmonic current flowing through the filter and whether the filter is tuned below or above the harmonic in question. If after a fluctuation the amount of current and the voltage leads the current then the reactance must be decreased. Conversely if the amount of current has increased but the voltage lags the current then the reactance must be increased.

In this example after the fluctuation four steps are required to bring the amount of harmonic current filtered within the required range. The step by step values for L and I_5 fifth harmonic current are presented in Table 5.9.

Table 5.9 Step by Step change of Current Limiting Self-Tuning Filter

Step	Phase Relationship	L (μH)	I_5(pu)	Percent of Original Current
Before Fluctuation	V Leads I	33.157	0.00055253	100%
After Fluctuation. No Steps	V Leads I	33.157	0.0053431	96.70%
1	V Leads I	32.742	0.0053922	97.59%
2	V Leads I	32.328	0.0054413	98.48%
3	V Leads I	31.914	0.0054907	99.37%

By the third step the current is within 1% of the original value. Tables 5.10 and 5.11 show the harmonic voltages and currents before the fluctuation in the capacitor and after the TLR is switched.

Table 5.10 Currents Before and after TLR switching

Harmonic Number	Current Before (pu)	Current Angle Before (Deg)	Current After (pu)	Current Angle After (Deg)
1	0.0081021	80.6500	0.0085355	80.6500
5	0.0055253	-35.0414	0.0053922	-36.3880
7	0.0018749	-63.0967	0.0018541	-63.1263
11	0.0007936	-119.9695	0.007942	-119.9683
13	0.0005417	-151.8158	0.005430	-151.8143

Table 5.11 Voltage Before and After TLR switching

Harmonic Number	Voltage Before (pu)	Voltage Angle Before (Deg)	Voltage After (pu)	Voltage Angle After (Deg)
1	0.9688	-9.35	0.9705	-9.35
5	0.0117505	54.9586	0.016584	53.6120
7	0.0377224	26.9033	0.038020	26.8737
11	0.0383440	-29.9695	0.0382104	-29.9683
13	0.0329994	-61.8158	0.0328479	-61.8143

As a final comparison the THD at the utility bus has dropped to 0.932%, below the required 1% value. The current limiting technique relies upon knowledge of whether the voltage leads or lags the current. If at any point in the switching the phase relation ship changes by 180° the tuning point has shifted from one side of the tuning point to another. If the current limit has not been met then either there is not enough harmonic current or the filters switching steps are not small enough.

The current limiting method differs from the direct or parameter estimation method in that it does not strive to remain tuned to a particular frequency. Instead the current limiting method tries to maximize the harmonic current removed from the system without exceeding its physical ratings. The next method utilizes linear methods to fit a model of the filter based on harmonic voltages and currents at frequencies other than the one of filtering interest.

5.5.3 Parameter Estimation

From the values presented in Table 5.4 and 5.5 the parameter estimation method requires three voltage and current values. These values are used to calculate the impedance of the filter seen by each harmonic. The impedance values calculated for the 1^{st} ,7^{th} ,11^{th} harmonic are presented in Table 5.12.

Table 5.12 The impedance at 1st ,7th ,11th harmonic

Harmonic	Voltage (pu) x10^{-3}	Current (pu) x10^{-3}	Z (pu)
1	970.5∠-9.35	8.5415∠80.65	113.6∠-90
7	38.4210∠26.8369	1.8321∠-63.1631	20.97∠90
11	38.4609∠-29.9729	0.7872∠-119.9729	48.86∠90

The linear equations used to model the filter can be written in matrix form:

$$\begin{bmatrix} 113.6\angle -90 \\ 20.97\angle 90 \\ 48.86\angle 90 \end{bmatrix} Z_{base} = \begin{bmatrix} 1 & j & -j \\ 1 & 7j & -j/7 \\ 1 & 11j & -j/11 \end{bmatrix} \begin{bmatrix} R \\ 120\pi L \\ \dfrac{1}{120\pi C} \end{bmatrix} \tag{5.20}$$

Inverting and solving for R, L, C gives

$$R=0.000\text{pu} \qquad L=33.1572\mu\text{H} \qquad C=9.6739\text{mF} \tag{5.21}$$

The impedance expressed on the right hand side of equation 5.20 are in per-unit; multiplying by the base value converts them into impedance values. Using the specified frequency the new value of L required is

$$120\pi L = \frac{1}{120\pi C}\frac{1}{(h)^2} = \frac{1}{120\pi(9.6379mF)(4.8)^2} = 0.01190 \tag{5.22}$$

resulting in a total inductance of:

$$L=31.568\mu\text{H}$$

the change in inductance required is calculated as

$$\Delta L = 31.568\mu H - 33.1572\mu H = -1.589\mu H \tag{5.23}$$

Diving this value by the minimum step size and following normal rounding rules give the total number of minimum TLR steps necessary to correct for the capacitor fluctuation.

$$\frac{-1.589\mu H}{0.41441\mu H} = -3.835 \rightarrow -4 \quad steps \tag{5.24}$$

The sign determines whether the inductance should be increases or decreased in this case the inductance value is decreased by four steps or 1.65764μH the parameters for the filter before and after the switching are presented in Table 5.13.

Table 5.13 Filter Parameter Before and After TLR Switching

Parameter	Before Fluctuation	After Fluctuation
L	0.03315mH	0.03150mH
C	9670.9uF	9670.9uF
Harmonic	4.68	4.81

Since the parameter estimation method resulted in the same number of step the harmonic voltages and currents before the fluctuation and after the TLR switching are the same as the direct method example. Similarly the THD at the utility bus has dropped to 0.898%, below the required 1% value.

5.6 Conclusions

Chapter 3 introduced the idea that a TLR could be used to construct a tunable harmonic filter. This chapter presented three methods of self-tuning for such a filter. These methods use information about the voltage across and the current through the filter and correct for harmonic filter parameter changes. The examples studied the case of a filter that experiences a capacitance fluctuation due to either temperature or damage. The fluctuation was such that the tuning point of the filter was moved to a lower frequency. The result of this was that more harmonic current was allowed to flow into the system raising the voltage THD at the utility bus.

Using the direct, current limiting and parameter estimation methods numerical examples were given to show how the TLR within the harmonic filter could be switched to re-tune the filter. For the direct and parameter estimation methods this involved calculating the change in inductance required to bring the filter, within the limits of finite TLR step changes, back to the original tuning frequency. The current limiting method based its TLR switching decisions on the amount of harmonic current flowing through the filter. Using this method the filter seeks for the optimum tuning frequency based on its physical limitations. A summary of the results obtained from Chapter 2 and this chapter are presented in Table 5.14.

Table 5.14 Summary of Utility Bus Voltage THD

Type of Adjustment	UTIL 69Voltage THD%
Before capacitor fluctuation (no adjustment)	0.940
After fluctuation but before adjustments	1.144
After direct method adjustment	0.898
After current limiting method adjustment	0.932
After parameter estimation method adjustment	0.898

The methods were presented in the order of decreasing requirements of filter harmonic voltage or current data. The direct method needs accurate measurements of voltage and current magnitudes and phase angles. The current limiting method requires accurate information on current magnitude and whether voltage is leading or lagging the current. While the parameter estimation method required no knowledge of the filter harmonic voltage and current but relies on at least three other harmonics.

Chapter 6

Conclusions

The goal of this work was to investigate the feasibility and application of a novel self-tuning filter. This goal was achieved through a series of steps. First to orient the reader to the necessity and purpose of the self-tuning filter, the concepts of power quality, the role of harmonics and methods of their mitigation were presented. The final section of this orientation was a test system example where harmonics generated by an ASD where shown to violate utility harmonic distortion limits.

The second step was to present the self-tuning filter and its central element the thyristor linked reactor. The TLR, a variable reactor, when used in conjunction with a fixed capacitor was shown to function as a tunable harmonic filter. The tunable filter was so named because by changing the configuration of the TLR the frequency response and therefore the tuning point was changed.

The third step was to examine the economic and theoretical feasibility of the self-tuning filter. The economic aspects were analyzed by comparing the equipment cost of a regular passive filter and those additional costs of the self-tuning filter due to the thyristors. It was shown that as filter ratings increased the cost differences between the two filters became negligible.

The theoretical feasibility was demonstrated analytically and through simulation. It was shown that the worst case transients created during TLR switching increased as the value of the inductance switched in and out of the filter was increased. Most importantly the transients approached zero as the inductance steps approached zero. The theoretical results were supported by simulations.

The fourth and final step was to present the method of operation of the self-tuning filter. Three methods were presented. Each used information either gathered from the filter during operation or preprogrammed to control the TLR switches. Two of the methods control the filters function so as to maintain filter tuning, that is it adjusts the TLR net inductance to compensate for reactive changes in the filter. The third method controls the amount of harmonic current flowing through the filter so that it never exceeds filter ratings. Each of these methods were tested on an example system similar to that found in the first step. All methods were shown to operate properly under the specified conditions.

With this last step the original goal was achieved. The novel self-tuning filter was investigated in terms of feasibility and application. The results show that the device is feasible economically and that it is possible to control transient magnitudes. Further in application the device is reasonable in that simple control mechanisms suffice for operation under static conditions. This final statement points in the direction of future self-tuning filter research.

The present work investigates TLR operation assuming that all thyristors act in an ideal manner. In reality the switching action will have a short non-linear conduction period. The effects of this behavior have not been considered. A concept related to this is the effect of switching transients on the simplified control methods. These transients may require a post switch dead time. This solution while acceptable for known transients, does not account for external sources of transients (from the utility). Further on the subject of transients the effect of TLR switching on the system must be assessed. In a manner similar to a switched capacitor a TLR may serve as a source of unwanted transients.

The factors involved in TLR control were only briefly presented in this work. All three methods rely on simple calculations for control. There is certainly the need for a more realistic look at the complications associated with accuracy of measurement and the time delays associated with the control methods.

Another question is the construction costs of a self-tuning filter. While a tapped inductor is a standard device the control system, thyristor connections and supporting superstructure are unique and their cost is unknown. These questions and others make the topic of a self-tuning filter using thyristor linked reactors a fertile field of academic interest.

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Appendix A

MatLab Script 1

This MatLab script is used to generate the coefficient data and graphs found in Chapter 4.

```
% Plot the magnitudes of transient coefficients
% as a function of percentage deviance from some
% set value.
% In this way the magnitudes can be used to estimate
% the effect on capactitors

w = 2*pi*60;

% Source values
VS = w*25000;
ZB = ((0.48)^2)/100;

% The source impedance
LS = 0.028543/1000;
RS = 0.001728;

% Filter values
RF = 0.0000;
LF = 0.0331572/1000;
CF = 9.21036/1000;

% Summed values that don't change
RT = RS+RF;
CT = CF;

% This is only used for the changed value
% There are two experiemnts
% One where the value of the system impedance
% can be increase for 28uH to 63uH
% and the other where it is deceased form 63 to 28
%
%LD = LS+LF;
%LN = LD*[0.4:0.001:0.999];
%LNtemp(1:600)=LN/LD;


%Ztemp(1:600) = 100*abs(-cos(-atan(w*LN./RT)+pi/2+atan(w*LD./RT)));

%LD = LS;
```

```
%LN = LD*[1:0.001:2.4];
%LNtemp(601:2001)=LN/LD;
%Ztemp(601:2001) = 100*abs(-cos(-
atan(w*LN./RT)+pi/2+atan(w*LD./RT)));

% This is for the case where there is no capacitor
% only a switched inductor

%plot(LNtemp,Ztemp,'w')
%title('Worst Case Transients For TLR')
%xlabel('Post to Pre Inductance Ratio')
%ylabel('Percent of Steady State')



% For the new circuit (hence the N postfix)
% These are the general souluoin parameters

LS = LS+LF;
LN = LS*[0.4:0.001:0.999];
LNtemp(1:600)=LN/LS;

woN = sqrt(1./(LN*CT));
DN = sqrt((LN.*LN).*((woN.*woN-w^2).*(woN.*woN-w^2))+RT^2*w^2);
KN = VS./DN;
TN = asin(RT*w./DN);
alpha = RT./LN;
r1N = -alpha+sqrt(alpha.*alpha-woN.*woN);
r2N = -alpha-sqrt(alpha.*alpha-woN.*woN);

% For the old circuit
% These are the general solution parameters
wo = sqrt(1./(LS*CT));
D = sqrt((LS^2)*((wo^2-w^2)^2)+RT^2*w^2);
K = VS/D;
T = asin(RT*w/D);

%For the switch time
n = 5;
tsw = ((n*pi/2)+T)/w;

A = -KN.*sin(T-TN);
B = LS*K*w;
D = -LN.*r1N;
E = -LN.*r2N;
F = LN.*KN.*w.*sin(w.*tsw-TN);

c1 = (B-F+A.*E)./(D-E);
c2 = -A-c1;

Zcurrent(1:600) = 100*abs(c1)./KN+100*abs(c2)./KN;

VSMM(1:600) = (VS./(DN)).*(w*LN-(1./(w*CF)));
VSMM(1:600) = abs(VSMM(1:600));
```

```
Zvoltage(1:600) =
(100./VSMM(1:600)).*(RT.*(abs(c1)+abs(c2))+LN.*(abs(c1.*r1N)+abs(c1.*
r1N)));



LS = LS;
LN = LS*[1:0.001:2.4];
LNtemp(601:2001)=LN/LS;

woN = sqrt(1./(LN*CT));
DN = sqrt((LN.*LN).*((woN.*woN-w^2).*(woN.*woN-w^2))+RT^2*w^2);
KN = VS./DN;
TN = asin(RT*w./DN);
alpha = RT./LN;
r1N = -alpha+sqrt(alpha.*alpha-woN.*woN);
r2N = -alpha-sqrt(alpha.*alpha-woN.*woN);

% For the old circuit
% These are the general solution parameters
wo = sqrt(1./(LS*CT));
D = sqrt((LS^2)*((wo^2-w^2)^2)+RT^2*w^2);
K = VS/D;
T = asin(RT*w/D);

%For the switch time
n = 5;
tsw = ((n*pi/2)+T)/w;

A = -KN.*sin(T-TN);
B = LS*K*w;
D = -LN.*r1N;
E = -LN.*r2N;
F = LN.*KN.*w.*sin(w.*tsw-TN);

c1 = (B-F+A.*E)./(D-E);
c2 = -A-c1;

Zcurrent(601:2001) = 100*abs(c1)./KN+100*abs(c2)./KN;


% These are the current transuients
plot(LNtemp,Zcurrent,'w');
title('Worst Case Current Transients');
xlabel('Post to Pre Inductance Ratio')
ylabel('Percent of Steady State')



VSMM(601:2001) = (VS./(DN)).*(w*LN-(1./(w*CF)));
VSMM(601:2001) = abs(VSMM(601:2001));
```

```
Zvoltage(601:2001) =
(100./VSMM(601:2001)).*(RT.*(abs(c1)+abs(c2))+LN.*(abs(c1.*r1N)+abs(c
1.*r1N)));


%plot(LNtemp,Zvoltage,'w');
%title('Worst Case Voltage Transients');
%xlabel('Post to Pre Inductance Ratio')
%ylabel('Percent of Steady State')
```

MatLab Script 2

This script is used in chapter 5 to generate the parameter estimation example data and results.

```
% To calculate the current through the filter and the %voltage
% across the filter at all harmonics
% The fundamental is given from PSA-H
% The only source of harmonics is the ASD which is acting %as ca
current
% source
% So all all non-fundamental frequencies the System %impedance
% from the frequency scan wih no filter is in parallel with %the
% filter impedance.
% Use a current splitter or voltage divider to calculate
% the voltage acroos the filter and the current through it


% Base impedance at the bus 7
Zb = ((0.480)^2)/100;



% To calculate the impedance of the filter and the system %seperately
% Use the system impedance with no filter
% and change the frequency of the filter
%Zs = (0.6949+i*4.4994)*Zb;
%Zf = i*2*pi*60*33.143/1000000-i*(1/(2*pi*60*9.210/1000));
%((Zs*Zf)/(Zs+Zf))/Zb;

% All this impedance data comes from bus 7
% Impedance data with properly tuned filter
Zwf0= [%      1.00  0.7501      4.6704      4.730
  5.00  0.0208      0.8115      0.812
  7.00  0.5127      9.4180      9.432
  11.00 0.2387      20.3276     20.329
  13.00 0.2208      25.0578     25.059
  17.00 0.2101      34.0937     34.094
  19.00 0.2086      38.5033     38.504
  23.00 0.2077      47.2109     47.211
```

```
 25.00 0.2076        51.5284      51.529
 29.00 0.2078        60.1180      60.118
];

% Impedance data with capacitor changed by -5%
Zwf5 = [%1.00 0.7532        4.6796        4.740
  5.00  0.1159        1.9114       1.915
  7.00  0.5370        9.6374       9.652
  11.00 0.2412        20.4351      20.436
  13.00 0.2223        25.1442      25.145
  17.00 0.2109        34.1564      34.157
  19.00 0.2092        38.5586      38.559
  23.00 0.2081        47.2557      47.256
  25.00 0.2080        51.5694      51.570
  29.00 0.2081        60.1530      60.153

];


%Impedance data with no filter
Znf=[%1.00    0.6949       4.4994      4.553
  5.00  18.7205       15.5507      24.337
  7.00  2.1682        19.2742      19.396
  11.00 0.7808        36.7604      36.769
  13.00 0.6927        44.3822      44.388
  17.00 0.6325        59.1560      59.159
  19.00 0.6210        66.4353      66.438
  23.00 0.6097        80.8923      80.895
  25.00 0.6068        88.0892      88.091
  29.00 0.6035        102.4450 102.447
  31.00 0.6025        109.6094 109.611
  35.00 0.6013        123.9196 123.921
  37.00 0.6009        131.0676 131.069
];

% These are the source impedances at the particular %frequencies
Zs = (Znf(:,2) + i*Znf(:,3))*Zb;
%Zs = abs(Zs)

% These are the frequencies in Hertz
w = 2*pi*60*Znf(:,1);

% The physical parameters of the filter
LF = 33.1572/1000000;
LF = 31.7422/1000000;


CF = 1.05*9.21036/1000;
% The impedance of the filter
Zf = i*w.*LF-i.*(1./(w*CF));

% The total impedance of system and filter
Zt = ((Zs'.*Zf')./(Zs'+Zf'))/Zb;
```

```
hs = [%   1            100.00          0
   5            18.24          -55.68
        7             11.90          -84.11
       11              5.73         -143.56
       13              4.01         -175.58
       17              1.93          111.39
       19              1.39           68.30
       23              0.94          -24.61
       25              0.86          -67.64
       29              0.71         -145.46
       31              0.62          176.83
       35              0.44           97.40
       37              0.38           54.36

];

% For the direct current method need to know
% The voltage across the filter and the current through
% it.  Magnitudes and relative phase angles are
% required

% To calculate the fundamental current drawn by the load
% use the rated power and voltage at the bus
% These are obtained from the sim
MagPuFund = 0.9704;
AngDegFund = pi*(-9.35)/180;

Load = (3000+i*800)/100000;  % PU
LoadCurr = Load/(MagPuFund.*(cos(AngDegFund)+i*sin(AngDegFund)));
LoadCurrAng = (180/pi)*atan(imag(LoadCurr)/real(LoadCurr));

MagPer5th = abs(LoadCurr)*hs(:,2)/100;
Angl5th = pi*(hs(:,3))/180+pi*LoadCurrAng/180;

% The current through the filter at the fundamental is out of phase
by
FundVolt = MagPuFund.*(cos(AngDegFund)+i*sin(AngDegFund));
Zf1 = (i*2*pi*60*LF-i*(1./(2*pi*60*CF)))/Zb;
FundCurr = FundVolt/Zf1;

abs(FundCurr)*1000
FundAng = 180*atan(imag(FundCurr)./real(FundCurr))/pi

% This is now the 5th harmonic current generated by the
% ASD in rectangular cooriduantes
I5th = MagPer5th.*(cos(Angl5th)+i*sin(Angl5th));

Zs = (Znf(:,2)+i*Znf(:,3));
Zf = (i*2*hs(:,1)*pi*60*LF-i*(1./(2*hs(:,1)*pi*60*CF)))/Zb;

Zd = ((Zs.*Zf)./(Zs+Zf));
Ifil=I5th.*((Zs)./(Zs+Zf));
Vfil=Ifil.*Zf;
```

```
abs(Ifil)*1000
180*atan(imag(Ifil)./real(Ifil))/pi;
180*angle(Ifil)/pi

abs(Vfil)*1000
180*atan(imag(Vfil)./real(Vfil))/pi;
180*angle(Vfil)/pi

sqrt(sum(abs(Ifil).^2))/abs(FundCurr)
```

Appendix B

PSPICE Scripts

This PSPICE scripts models a simple switched inductor using voltage controlled switches. At first the inductor is present at the one second mark it is removed. The results of these are found in chapter 4 in the transient analysis.

```
Example 8.6
.OPTIONS NUMDGT=6 NOECHO NOBIAS NOMOD NOPAGE
*Ex 8.6
.MODEL   SWITCH      VSWITCH(RON=1e-7  ROFF=1E+7   VON=1
   VOFF=0);

* Models a TLR with a single switched element

* The signal source is set at 100V to show
*changes as a percentage

V  1     0     sin(0 100   60)

* The switches are used as the thyristors
* for simpliticy these are manually timed
VSbypass 10    0      PULSE(0V 5V 1.0037s 0.0001s 0.0001s
6s 10s);
VSinline 11    0      PULSE(5V 0V 1.0037s 0.0001s 0.0001s
6s 10s);

* The values were obtained from the Freq Scan of
* case 2. Take the frequency scan value and multiply
* it by the base at THAT BUS!
* So for bus 7 in the project case (.480^2)/100
RS 1     2     0.001728
LS 2     3     28.543uH


Sinline 4      0      11     0      SWITCH
Sbypass  3     0      10     0      SWITCH

* These are the inductor values values
LF 3     4     0.0000331572H
*RF      5     0      0.001

* Start time is delayed to allow circuit to
* reach steady state
.TRAN    0.0001      1.4000      0.6000 0.0001
```

```
* The current through the circuit
* and the voltage across the filter
* inductor are used as measurments
.PRINT TRAN I(RS) V(LF)

.END
```

Opposite case of the last script. In this one the inductor is first missing then later it is added

```
Example 8.6
.OPTIONS NUMDGT=6 NOECHO NOBIAS NOMOD NOPAGE
*Ex 8.6
.MODEL   SWITCH      VSWITCH(RON=1e-7  ROFF=1E+7   VON=1
   VOFF=0);

* The signal source is set at 100V to show
* changes as a percentage
V  1      0      sin(0 100   60)

* The switches are used as the thyristors
* for simplicity these are manually timed
VSbypass 10     0      PULSE(5V 0V 1.0037s 0.0001s 0.0001s
6s 10s);
VSinline 11     0      PULSE(0V 5V 1.0037s 0.0001s 0.0001s
6s 10s);

* The values were obtained from the Freq Scan of
* case 2.
RS 1      2      0.001728
LS 2      3      28.543uH


Sinline 4       0      11     0      SWITCH
Sbypass  3      0      10     0      SWITCH

* These are the filter values
LF 3      4      0.0000331572H
*RF       5      0       0.001

* Start time is delayed to allow circuit to
* reach steady state
.TRAN     0.0001       1.4000       0.6000 0.0001

* The current through the circuit
* and the voltage across the filter
* inductor are used as measurments
.PRINT TRAN I(RS) V(LF)

.END
```

This example uses a complete LC filter with a single element switched TLR. First the segment is present then it is removed.

```
Example 8.6
.OPTIONS NUMDGT=6 NOECHO NOBIAS NOMOD NOPAGE
*Ex 8.6
.MODEL   SWITCH      VSWITCH(RON=1e-7  ROFF=1E+7   VON=1
   VOFF=0);

* A simple filter with switched inductor
* The signal source is set at 100V to show
* changes as a percentage
V  1     0     sin(0 100   60)

* The switches are used as the thyristors
* for simplicity these are manually timed
VSbypass 10    0     PULSE(0V 5V 1.0037s 0.0001s 0.0001s
6s 10s);
VSinline 11    0     PULSE(5V 0V 1.0037s 0.0001s 0.0001s
6s 10s);

* The values were obtained from the Freq Scan of
* case 2.
RS 1     2     0.001728
LS 2     3     28.543uH


Sinline 4      5     11     0     SWITCH
Sbypass  3     5     10     0     SWITCH

* These are the filter values
LF 3     4     0.0000331572H
*RF      5     0     0.001
CF 5     0     9.21036mF

* Start time is delayed to allow circuit to
* reach steady state
.TRAN    0.0001      1.4000      0.6000 0.0001

* The current through the circuit
* and the voltage across the filter
* inductor are used as measurments
.PRINT TRAN I(RS) V(3,0)  V(CF)

.END
```

Opposite case of the one above

```
Example 8.6
.OPTIONS NUMDGT=6 NOECHO NOBIAS NOMOD NOPAGE
*Ex 8.6
.MODEL   SWITCH      VSWITCH(RON=1e-7  ROFF=1E+7   VON=1
   VOFF=0);

* Models a LC filter when the reactance is initially
bypassed
* then later added
* The signal source is set at 100V to show
* changes as a percentage
V  1     0     sin(0 100   60)

* The switches are used as the thyristors
* for simpliticy these are manually timed
VSbypass 10    0      PULSE(5V 0V 1.0037s 0.0001s 0.0001s
6s 10s);
VSinline 11    0      PULSE(0V 5V 1.0037s 0.0001s 0.0001s
6s 10s);

* The values were obtained from the Freq Scan of
* case 2.
RS 1     2     0.001728
LS 2     3     28.543uH


Sinline 4      5      11     0     SWITCH
Sbypass  3     5      10     0     SWITCH

* These are the filter values
LF 3     4     0.0000331572H
*RF      5     0      0.001
CF 5     0     9.21036mF

* Start time is delayed to allow circuit to
* reach steady state
.TRAN    0.0001      1.4000      0.6000 0.0001

* The current through the circuit
* and the voltage across the filter
* inductor are used as measurments
.PRINT TRAN I(RS) V(3,0) V(CF)

.END
```