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Adaptive Harmonic Blocking Compensators.

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ADAPTIVE HARMONIC BLOCKING COMPENSATORS

A Dissertation

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in

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by

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ABSTRACT

This dissertation explores a new method of adaptive compensation and harmonic suppression in distribution systems. The compensator under development and investigation integrates a harmonic blocking compensator (HBC) with an adaptive balancing compensator (ABC) resulting in an adaptive harmonic blocking compensator (AHBC). Such a compensator can be used for compensating reactive power, for symmetrizing supply currents, for protecting the distribution system against load generated current harmonics and, consequently, the voltage harmonic distortion caused by non-linear loads.

Harmonic blocking compensators are designed for compensating fixed loads, and they are not adaptive devices, while the adaptive capability of compensators is becoming more and more important. This dissertation investigates the possibility of the conversion of a HBC into an adaptive device. A HBC and an ABC are combined to form a new device which is referred to as an "adaptive harmonic blocking compensator." A HBC provides the harmonic suppression. An ABC which contains three thyristor controlled susceptance circuits provides the variable susceptances needed for reactive and unbalanced power compensation. A digital signal processing system (DSPS) for three-phase quantities has been developed for the AHBC control. This DSPS consists of signal processing hardware and a control algorithm. The hardware includes transducers, a microcontroller and some associated circuits for signal processing. The control algorithm provides the thyristor firing angles calculated based on the current orthogonal decomposition method.

The dissertation presents the fundamentals of AHBC operation and design, results of computer modeling using PSpice and experimental results obtained from a laboratory prototype.
CHAPTER 1
INTRODUCTION

The increasing cost of energy has created an increased interest in the efficient usage of electric energy [1-5]. It is well known that to improve this efficiency, to improve the supply capability, to reduce the power loss and to improve the voltage profile, the load reactive power should be compensated [6-11], i.e., the power factor should be improved. Capacitor banks are installed for that purpose. They improve the power factor of the supply to an economically justified value.

With the proliferation of non-linear loads, mainly of power electronic equipment but also other loads such as fluorescent lamps, arc furnaces, etc., there is an increasing concern regarding voltage distortion in distribution systems. This distortion appears mainly because the current harmonics generated in non-linear loads cause a voltage drop on the system impedance. Moreover, harmonic distortion could increase substantially in the presence of a capacitor bank. Series resonance of this bank with the distribution system inductance at a harmonic frequency or in its vicinity is responsible for that increase.

Harmonics cause a number of harmful effects to customer devices and to distribution system equipment. All of these harmful effects are referred to as the main cause of the supply quality degradation in distribution systems. Thus additional equipment is needed to reduce harmonic distortion if its level reaches an unacceptable value. It would be advantageous, of course, if the same equipment could provide not only the reactive power compensation but also the harmonic suppression to achieve power factor and supply quality improvement.

Resonant harmonic filters (RHF) are traditionally used for reducing harmonic distortion [12-20]. They also reduce the reactive power of the supply, thus they fulfill the tasks of traditional compensators. RHF are usually tuned to a few dominating harmonics of the load current and provide a low impedance path for those harmonics,
thus, reducing their contents in the supply current. Unfortunately, the remaining harmonics can be amplified since the filter creates a parallel resonance with the distribution system inductance. Moreover, the supply voltage distortion degrades the performance of RHF's since RHF's branches have low impedance for supply voltage harmonics, which increases the supply current distortion.

It is also known [21-27] that not only reactive current and current harmonics but also unbalanced current, caused by the load unbalance, deteriorate the power factor. Adaptive balancing of a load requires that the branches of the compensator have an impedance controlled in an inductive and/or capacitive range. This control can be provided by a thyristor controlled susceptance (TCS) circuit [28-32] that includes a thyristor-switched inductor (TSI). A suitable TCS for this purpose should have the following properties:

(1) it has to provide susceptance, inductive and/or capacitive, needed for unbalanced and reactive current compensation at the fundamental frequency.
(2) it should inject the smallest possible amount of harmonic currents to the rest of the system.
(3) it should not cause a resonance between the TCS and the distribution system inductance.

Adaptive balancing compensators (ABCs) built of TCSs have been studied and developed as adaptive devices for the compensation of the reactive and unbalanced power in References [29,31]. However, ABCs are not designed for harmonic suppression. It may be desirable that adaptive compensators not only have the capability to reduce reactive current and unbalanced current but also the capability to reduce the harmonic distortion.

In recent years, active filters (AFs) have been studied for power factor and supply quality improvement by injecting a compensating current [33-40]. However, due to their complexity, difficulties with building a large-rated AF and the high cost of
building them, AFs are not very widespread devices. They are still mainly under
development.

The objective of this dissertation is to develop and evaluate a new compensator
for power factor and supply quality improvement, referred to as an "adaptive harmonic
blocking compensator" (AHBC). Before discussing the development of this new
compensator, all existing devices, as mentioned above, are reviewed with respect to
their properties. Conclusions are drawn regarding desirable properties of a new
compensator by analysis of the advantages and disadvantages of these devices. Current
orthogonal decomposition is the main theoretical tool for this study. It decomposes the
load current into a useful component, referred to as the active current, and a non-useful
component composed of the scattered, reactive, unbalanced and generated harmonic
currents. All of these four current components are mutually orthogonal. The
compensation is a matter of minimizing those current components which are not useful,
without causing a noticeable effect on the useful current component. It should be noted
that although in some situations total compensation is theoretically possible, it may not
be economical to improve power factor to unity in practice. With the current harmonic
distortion and voltage harmonic distortion as the performance indicators, computer
modeling using PSpice is used to compare the performances of these devices.

A harmonic blocking compensator (HBC) [41-43] is a newly developed device
that can provide higher effectiveness in harmonic suppression than traditional filters.
The study of the new compensator will begin with a compensator for fixed loads. This
compensator is converted next into an adaptive compensator.

A HBC is built of a series filter tuned to the fundamental frequency and a shunt
capacitor. Due to the voltage resonance, the series filter does not increase the line
impedance for the fundamental, but increases its impedance for higher order harmonics.
The shunt capacitor provides the needed susceptance for the reactive power
compensation and a low impedance shunt path for harmonic currents generated by the
load. Computer modeling using PSpice is used to verify the properties of HBCs expected from theoretical study. The capacitor bank, resonant harmonic filter and harmonic blocking compensator are compared with respect to the effectiveness of harmonic suppression.

HBCs are effective for compensation of fixed loads, while over the last few decades more and more fast varying loads have been installed in distribution systems. This is an effect of development of modern industrial processes with automatic robots and adjusting energy flow to consumer needs. The dissemination of power controlling electronic equipment causes higher variability of the reactive power and waveform distortions than before. This variability may require that compensators and harmonic suppression equipment have an adaptive property. Traditionally used resonant harmonic filters do not have adaptive properties, of course, and they usually have too many components to convert RHFs into adaptive devices. Having less components, a HBC is more suitable for conversion into an adaptive device than a RHF.

An adaptive harmonic blocking compensator (AHBC) can be obtained if a HBC is combined with a thyristor controlled susceptance circuit. A TCS, which contains a thyristor switched inductor, provides the variable susceptance, capacitive and/or inductive, needed for an adaptive compensation of reactive and unbalanced power. Therefore, the combination of a HBC with a TCS results in an adaptive device not only for reactive power compensation but also for unbalanced power compensation. An AHBC should minimize the reactive and unbalanced current dynamically and eliminate, or at least reduce, most of the current harmonics. However, TCSs generate harmonics and may cause resonances with the rest of the system. Therefore, various structures of TCSs and of AHBCs are considered and carefully chosen to avoid these effects.

This dissertation presents the fundamentals of AHBC operation and design. Computer modeling of the AHBC using PSpice for single-phase and three-phase systems are used for the verification of the AHBC effectiveness. Computer simulation
may not provide, however, full information about a new device under development. This research is supported with physical experiments for that purpose. These experiments are an essential component of this dissertation. Thus the properties of the AHBC are concluded based on the theoretical analysis, computer modeling and physical experiments.

To provide the control signals for the load compensation and the load current balancing with the AHBC, the supply voltages and currents are measured and adequately processed. A digital signal processing system (DSPS) for three-phase quantities has been developed for the AHBC control. The DSPS hardware consists of two current and two voltage transducers, a microcontroller and some associated circuits. To achieve high-speed control, a 16-bit microcontroller (8XC196MC by Intel) designed especially for high-speed event control is used. It uses assembly language, a low-level language, to accomplish its high-speed feature. To compensate the reactive and unbalanced current based on the information of the system provided by DSPS, a control algorithm has been developed. This control algorithm provides the thyristor firing angles calculated based on the current orthogonal decomposition method. Then, these calculated firing angles are converted into time delay for the actual firing triggers.

This dissertation is arranged as follows. Chapter 2 briefly reviews the fundamentals of the current decomposition and the reduction of the non-useful component for linear loads. Properties of an adaptive balancing compensator and its application are studied in that chapter. Chapter 3 extends the current decomposition from linear loads to non-linear loads. Properties of capacitor banks, resonant harmonic filters and active filters are reviewed and compared in Chapter 3. In Chapter 4, the fundamentals and design of harmonic blocking compensators are developed. The comparison between a CB, RHF and HBC is based on computer modeling using PSpice. The conversion of a HBC into an adaptive harmonic blocking compensator is discussed in Chapter 5. Experimental results of HBCs and AHBCs are presented in Chapter 6.
Conclusions of this study are included in Chapter 7. Some additional information, circuit configurations used for experiments such as driver circuits for TCSs, microcontroller pin connection etc., are included in Appendixes.
CHAPTER 2
REACTIVE AND UNBALANCED CURRENT REDUCTION

2.1 Introduction

Electric energy is generated, transmitted and distributed mainly in three-phase systems. To be specific, electric energy is generated mainly by three-phase symmetrical generators, then transformed, transmitted and distributed in three-phase systems, except for the lowest voltage levels of distribution systems where single-phase systems are used. Symmetry is a very important feature of such three-phase systems. This is because three-phase symmetrical systems are more effective than single-phase systems with respect to the capability of energy transmission. At the lowest voltage level, 120 volts in the United States, energy is consumed mainly by single-phase loads, since low power single-phase devices are cheaper than three-phase devices and some devices, such as bulbs, are not built as three-phase devices. The single-phase loads are combined together to form a three-phase load. In such a case, the three-phase system is most likely unbalanced with respect to its load impedances. Consequently, the supply currents are asymmetrical with respect to current rms values and phase angles between them. Such combined single-phase loads stand for three-phase unbalanced loads and cause supply current asymmetry and supply voltages asymmetry that are undesirable to both power consumers and utilities. This asymmetry may have substantially detrimental effects on power factor and on the performance of some three-phase equipment where symmetrical supply voltage is crucial. This is mainly related to three-phase rotating machines and to power electronic devices. In the former case, the supply voltage asymmetry, even of a relatively low degree, may result in a substantial increase in power loss, in machine overheating and in motor efficiency deterioration. For example, a 3.5% voltage unbalance can result in a 25% increase in the motor temperature of an induction machine [25]. In the latter case, the asymmetrical voltage applied to three-phase power electronic devices, such as AC/DC converters and AC voltage controllers, causes non-
characteristic harmonics to occur in the current and voltage [25-27] and results in load current asymmetry. Therefore, it is advantageous for a consumer if the utility supplies him with a symmetrical voltage. The current asymmetry causes degradation of power factor due to an increase in the supply current rms value, i.e., an increase in apparent power, while the active power remains the same. As a result, asymmetry causes an extra energy loss and increased heating, referred to as the operating cost, or it may require that sizes of transmission lines, transformers, breakers, etc., referred to as the investment cost, be increased in order to reduce the power loss. In each of these cases, either the operating cost or the investment cost will increase. It is desirable for utility companies if the supply voltages are symmetrical and the supply currents are symmetrical and minimized.

2.2 Current decomposition and reduction

To study power factor improvement, it is advantageous to decompose current into the useful and non-useful components. A current or a voltage can be expressed by Fourier series, for instance, for a current:

\[ i = I_0 + \sqrt{2} \Re \sum_{n \in N} I_n e^{j\omega_n t}, \]  

(1)

where \( \omega_n = 2\pi/T \), \( N \) is the set of the harmonic orders of supply voltage and the boldface capital letter \( I_n \) is the complex rms value (CRMS) of the \( n \)-th order harmonic and can expressed as follows:

\[ I_n = I_n e^{j\alpha_n}. \]  

(2)

The rms value of current \( i \) can be calculated as

\[ ||i|| = \sqrt{\sum_{n \in N} I_n^2}. \]  

(3)

In three-phase systems, for any kind of load, the three-phase current vector can be defined [22-24] in terms of line currents \( i_R, i_S \) and \( i_T \) as follows:
where the boldfaced lower case \( i \) denotes a current vector. The thermal effects of asymmetrical currents on a symmetrical three-phase device shown in Figure 2.2.1(a) are the same as in a single-phase device shown in Figure 2.2.1(b), if instead of these three currents \( i_R, i_S \) and \( i_T \), an equivalent current rms value is defined as follows:

\[
||i|| = \sqrt{\frac{1}{T} \int_0^T i^* i dt} = \sqrt{||i_R||^2 + ||i_S||^2 + ||i_T||^2}.
\]  

(5)

Throughout this dissertation, \( ||i|| \) will be called as the rms value of a three-phase current vector. Similarly, the rms value of a three-phase voltage vector, \( ||u|| \), can be defined as follows:

\[
||u|| = \sqrt{\frac{1}{T} \int_0^T u^* u dt} = \sqrt{||u_R||^2 + ||u_S||^2 + ||u_T||^2}.
\]  

(6)

The most noticeable advantage of using such a notation is that most equations defined for single-phase systems can be easily converted to equations for three-phase systems.

As discussed in Ref. [24], the equivalent admittance \( Y_{en} \), the equivalent conductance \( G_{en} \), the equivalent susceptance \( B_{en} \) and the unbalanced admittance \( A_n \) of the \( n \)-th harmonic for three-phase systems with unbalanced linear loads can be introduced and defined as follows:
\[ Y_{en} = \frac{S_{en}^*}{||u_n||^2} = Y_{R_{Sn}} + Y_{S_{Sn}} + Y_{T_{Rn}}, \]  
\[ G_{en} = \text{Re}\{Y_{en}\} = G_{R_{Sn}} + G_{S_{Sn}} + G_{T_{Rn}}, \]  
\[ B_{en} = \text{Im}\{Y_{en}\} = B_{R_{Sn}} + B_{S_{Sn}} + B_{T_{Rn}}, \]

and

\[
A_n = \begin{bmatrix} A_{Rn} & 0 & 0 \\ 0 & A_{Sn} & 0 \\ 0 & 0 & A_{Tn} \end{bmatrix} = A_n \begin{bmatrix} 1 & 0 & 0 \\ 0 & \beta & 0 \\ 0 & 0 & \beta^* \end{bmatrix},
\]

where

\[
A_n = -(Y_{S_{Sn}} + \beta Y_{T_{Rn}} + \beta^* Y_{R_{Sn}}),
\]

\[
\beta = -\frac{1}{2} - j \frac{\sqrt{3}}{2} = e^{j2\pi/3},
\]

and \( s \) is a 'sequence index' defined as follows:

\[
s = \begin{cases} 
+1 & \text{for harmonics of positive sequence, } n = 3k + 1 \\
-1 & \text{for harmonics of negative sequence, } n = 3k - 1
\end{cases}
\]

Under the assumption that the supply voltage is symmetrical, the current can be decomposed as follows:

\[
i = \sum_{n \in N} (i_{an} + i_{rn} + i_{un}),
\]

where

\[
i_{an} = \begin{bmatrix} i_{Ran} \\ i_{San} \\ i_{Tan} \end{bmatrix} = G_{en}U_{an} = \sqrt{2}\text{Re} \begin{bmatrix} G_{en}U_{Rn} \\ G_{en}U_{Sn} \\ G_{en}U_{Tn} \end{bmatrix} e^{j\alpha t},
\]

\[
i_{rn} = \begin{bmatrix} i_{Rrn} \\ i_{Srn} \\ i_{Trn} \end{bmatrix} = \sqrt{2}\text{Re} \begin{bmatrix} jB_{en}U_{Rn} \\ jB_{en}U_{Sn} \\ jB_{en}U_{Tn} \end{bmatrix} e^{j\alpha t},
\]

and

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Let us define the equivalent conductance $G_e$ of the unbalanced linear load at the nonsinusoidal voltage $u$, as the conductance of a symmetrical linear load which at the same voltage $u$ has the same active power $P$, i.e.,

$$G_e = \frac{P}{||u||^2}.$$  \hspace{1cm} (18)

Furthermore, one can decompose the current into four components with regard to their physical meanings:

1. active current $i_a$ - associated with active power transmission.
2. reactive current $i_r$ - associated with reciprocating energy transmission.
3. scattered current $i_s$ - caused by the dependency of the load conductance on harmonic frequencies.
4. unbalanced current $i_u$ - caused by load unbalance.

Then the load current can be expressed as follows:

$$i = i_a + i_r + i_s + i_u.$$  \hspace{1cm} (19)

where

$$i_a = G_e u,$$  \hspace{1cm} (20)

$$i_s = \left( \sum_{n \in N} i_{an} \right) - i_a,$$  \hspace{1cm} (21)

$$i_r = \sum_{n \in N} i_{rn},$$  \hspace{1cm} (22)

and

$$i_u = \sum_{n \in N} i_{un}.$$  \hspace{1cm} (23)

Because of the orthogonality between each component, their rms values fulfill the following relation:

$$||i||^2 = ||i_a||^2 + ||i_r||^2 + ||i_s||^2 + ||i_u||^2.$$  \hspace{1cm} (24)
Multiplying both sides of eqn. (24) by $|u|^2$ results in power equations of this form

$$S^2 = P^2 + Q^2 + D_s^2 + D_u^2,$$

where apparent power $S = |u||i|$, active power $P = |u||i_a|$, reactive power $Q = |u||i_r|$, scattered power $D_s = |u||i_s|$ and unbalanced power $D_u = |u||i_u|$. These components can be related to four different phenomena:

(i) active power: the only useful component, the time average of the instantaneous power over one period of the waveform.

(ii) reactive power: the alternating exchange of energy between two areas.

(iii) scattered power: the extra power increase of apparent power, caused by the dependency of load conductance on harmonic frequency.

(iv) unbalanced power: the extra increase of apparent power, caused by the load unbalance.

One should notice that the only active power is necessary for power transmission while the other unwanted components may be associated with this transmission. Power factor $\lambda$ is defined as:

$$\lambda = \frac{P}{S},$$

where $P$ denotes the active power. Since only the active current $i_a$ is needed for active power transmission while the supply source is loaded with current $i$, the power factor can be expressed in terms of these quantities, namely,

$$\lambda = \frac{|i_a|}{|i|} = \frac{|i_a|}{\sqrt{|i_a|^2 + |i_r|^2 + |i_s|^2 + |i_u|^2}} = \frac{1}{\sqrt{1 + \frac{|i_r|^2 + |i_s|^2 + |i_u|^2}{|i_a|^2}}}. \quad (27)$$

One may conclude from eqn. (27) that improving power factor is a matter of the minimization of supply current rms value without noticeably increasing active current. Therefore, the main task for power factor improvement is to eliminate, or at least to reduce, those useless components.
The most important advantage of this current decomposition is that having these mutually orthogonal components one can deal with each of them at a time without affecting others. The reactive and unbalanced currents can be compensated by shunt passive compensators. However, the scattered current cannot be compensated by such compensators. The scattered current is usually relatively small as compared to the other components, therefore, the reduction of this current is usually out of consideration with respect to the economic viewpoint. For unbalanced linear loads, the task can be narrowed down to the reduction of the reactive and unbalanced currents. Although the total compensation of these two current is theoretically possible for fixed loads [24], in practice, loads change with time, so that the structure and parameters of such a compensator would have to be adjusted to this change. In practical applications, only the reduction of the largest components, i.e., the fundamental components of the reactive and unbalanced currents, is technically possible for varying loads. Therefore, its compensation becomes the main subject for power factor improvement.

2.3 Adaptive balancing compensators

Although the minimization of the reactive and unbalanced currents can be done by the reactance circuits described in Ref. [23,24], such compensators only minimize the current of only fixed loads. The unbalanced and reactive powers of a load are variable quantities, usually, so that a fixed compensator can compensate only a fixed portion of these powers. Adaptive compensators are needed for total compensation of these powers.

To compensate the reactive power of an inductive load the compensator has to provide only a capacitive susceptance. However, the compensator has to provide a capacitive and/or inductive susceptance for compensating the unbalanced power. Thyristor controlled susceptance (TCS) circuits are usually used for this purpose [6,28-32]. A thyristor-switched inductor (TSI) connected to a capacitor in parallel, as shown in Figure 2.3.1, forms a basic structure of TCS. Such a TCS is referred to as a "TCS
with a basic structure" (TCS_B). By varying the firing angle of the thyristor, the susceptance, either capacitive or inductive, of the TCS can be varied for an adaptive compensation.

![Figure 2.3.1 Basic structure of a TCS.](image)

The variable susceptance is provided by the one-port composed of the capacitor $C$ and the TSI. The plots of susceptance of this one-port with the thyristor switched completely ON or OFF are shown in Figure 2.3.2. The susceptance of the one-port is of a minimum value equal to

$$ T_a = \omega \frac{1}{\omega L_C} $$

(28)

when the thyristors are switched ON over the entire half of the period, i.e., at the firing angle $\alpha = 90^\circ$. When the thyristors are in the OFF state over the whole period, i.e., at the firing angle $\alpha = 180^\circ$, the susceptance of the one-port has the maximum value

$$ T_b = \omega C. $$

(29)

For $90^\circ \leq \alpha \leq 180^\circ$, the susceptance $T$ for the fundamental frequency has a value between $T_a$ and $T_b$, as shown in Figure 2.3.2.
A set of waveforms of the TSI current $i_{tsi}$ at a sinusoidal supply voltage $u$ are shown in Figure 2.3.3. Because of the current distortion, the thyristor-switched inductor is a source of current harmonics, especially the third harmonic. Thus, the variable susceptance needed for adaptive compensation and balancing is obtained at the cost of waveform distortion.

To illustrate the level of harmonic distortion, it was assumed that a TCS was to be controlled in the range of $T_a = 0.07S$ to $T_b = 1.36S$ in the system with the short-
circuit power 25 times higher than the load power. It was assumed that reactance to resistance ratio of the distribution system $X/R = 5$. The total harmonic distortion of the supply current changes with the firing angle. For the TCS_B, this distortion, as shown in Figure 2.3.4, may be even higher than 50%.

![Figure 2.3.4 Plot of CHD vs. firing angle of the TCS_B.](image)

An adaptive compensator built of three TCSs is referred to as an "adaptive balancing compensator" (ABC). An ABC with three TCS_Bs in a delta configuration is shown in Figure 2.3.5. Such a compensator is referred to as an "ABC_B". Such an ABC

![Figure 2.3.5 Circuit configuration of an ABC_B.](image)
is suitable, however, for compensating of only balanced loads. In such a case, the third 
order harmonics generated in the compensator branches are in phase and canceled 
mutually in the delta structure. Otherwise the third order harmonics generated by the 
TSI branches differ mutually, and their differences are injected into the system. 
Moreover, the shunt capacitor may cause an unexpected increase of current harmonics 
because of the resonance between the capacitance and the system inductance. Therefore, 
a modified ABC with a structure that can eliminate the above drawbacks is desirable.

To reduce the effects of the resonance between the TCS and the system 
inductance, a series inductor can be inserted. To reduce the current harmonic injection, 
in particular, the third harmonic, a harmonic filter tuned to the third harmonic can be 
connected in parallel with the TSI, instead of the shunt capacitor. Several different 
structures of TCSs, as shown in Figure 2.3.6, have been studied [28-32]. Program 
written in C++ were used to study these TCSs. It was concluded that the one-port of the 
structure specified as TCS_SI_PF had outperformed the others.

![Figure 2.3.6 Different structures of TCSs studied.](image)

The susceptance of the TCS_SI_PF can be controlled in the range of \( T_a \) to \( T_b \), if 
inductance \( L \) fulfills the equation

\[
a_3L^3 + a_2L^2 + a_1L + a_0 = 0
\]

(30)

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with coefficients [28]:

\[ a_0 = 9 - \Omega^2, \]  
\[ a_1 = (27 - 1\Omega^2)\omega_1T_b, \]  
\[ a_2 = [2T_b(9 - 5\Omega^2) + 9T_a(1 - \Omega^2)], \]  

and

\[ a_3 = 9T_a(1 - \Omega^2)\omega_1^2T_b^2. \]  

The symbol \( \Omega \) denotes the relative frequency of the series resonance in the circuit with thyristors switched ON completely. It should be chosen below the second harmonic frequency, i.e., \( \Omega < 2 \). Its choice affects the number of solutions of eqn. (30) with real positive \( L \), as well as, the compensator parameters. After inductance \( L \) is calculated from eqn. (30), the remaining parameters of the TCS_SI_PF can be found with the following equations [28]:

\[ L_2 = \frac{(\omega_1L + \frac{1}{T_b})}{8}, \]  
\[ C_2 = \frac{1}{9\omega_1^2L_2}, \]  

and

\[ L_c = \frac{\omega_1^2L^2T_aT_b + \omega_1L(T_a + T_b) + 1}{(T_b - T_a)\omega_1}. \]  

With \( T_a = 0.07S, T_b = 1.36S, \Omega = 1.7 \) and the same assumptions as for TCS_B, the plot of the \textit{CHD} of the TCS_SI_PF versus the firing angle is shown in Figure 2.3.7. The current distortion is below 1.5% in the whole range of the susceptance control.
Figure 2.3.7 Plot of CHD vs. firing angle of the TCS_SI_PF.

The modified circuit configuration of an ABC with three TCS_SI_PFs in a delta configuration, referred to as an "ABC_M," is shown in Figure 2.3.8.

![Circuit configuration of an ABC_M.](image)

Figure 2.3.8 Circuit configuration of an ABC_M.

The control of the balancing compensator can be simplified if the needed susceptances \( T_{RS}, T_{ST} \) and \( T_{TR} \) of the compensator branches are calculated based on the measured values of the unbalanced admittance and the equivalent susceptance of the load. Namely [23],

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\[ T_{RS} = \frac{(\sqrt{3} \text{Re} A_i - \text{Im} A_i - B_{el})}{3}, \]
\[ T_{ST} = \frac{(2 \text{Im} A_i - B_{el})}{3}, \]
and
\[ T_{TR} = -\frac{(\sqrt{3} \text{Re} A_i + \text{Im} A_i + B_{el})}{3}. \]

After the needed susceptances \( T_{RS}, T_{ST} \) and \( T_{TR} \) are calculated, the firing angles of thyristors could be found with a 'look-up' table. However, this is a open-loop approach.

The ABC can be controlled in a closed-loop with a proportional and integral (PI) control algorithm to ensure the accuracy. The signals for the ABC's control are generated by the digital signal processing system (DSPS) of the ABC based on the measurement of two line currents and two line-to-line voltages at the supply terminals. This is discussed in more detail in Chapter 5.

The experimental results obtained for the modified structure of the ABC, ABC\_M, are presented in Reference [31]. The results show that an ABC\_M compensates reactive and unbalanced power of the fundamental frequency and achieves the power factor improvement without injecting an excessive amount of current harmonics.
CHAPTER 3
HARMONIC SUPPRESSION

3.1 Introduction

In recent years, as switching speed and handling capabilities with respect to voltage and current ratings of semiconductor devices have increased, a large number of power electronic devices such as switch-mode dc power supplies, uninterruptible power supplies and AC/DC converters have been used in industrial and commercial applications. Unfortunately, the load current waveforms of such power electronic equipment is far from sinusoidal and power factor is usually poor [1-5] not only because of reactive and unbalanced current components but also because of harmonic contents. Because of the distribution system impedance, distorted current causes voltage distortion. Consequently, with the proliferation of non-linear loads, voltage distortions are more and more often observed in distribution systems. Moreover, harmonic distortion could increase substantially in the presence of capacitor banks because of the series resonance between the bank and the system inductance.

Harmonics cause a number of harmful effects to customers and to distribution system equipment [14]. All of these harmful effects are referred to as the supply quality degradation in distribution systems. Thus, additional equipment may be needed to reduce harmonic distortion if it reaches an unacceptable level. Resonant harmonic filters (RHF) have been installed in distribution systems for reducing harmonic distortion [12-20]. They also compensate the reactive power of the loads. RHF are usually tuned to a few dominating harmonics of the load current, however, and they are sensitive to voltage distortion. Therefore, the effectiveness of RHF may not be sufficient. Harmonic distortion could be also reduced by active filters (AF). Active filters for reactive power compensation and harmonic suppression have been studied [33-40] since their basic principle was proposed in the 1970's. Recent development of IGBTs and GTO thyristors with elevated voltage-current rating and switching speed has
spurred greater interest in the study of active power filters focusing on practical applications. However, it is still difficult to construct a high power active filter with a rapid current response. Moreover, their initial costs and maintenance costs are high. These drawbacks still limit the use of active filters in practical applications.

A novel device for reactive power compensation and harmonic suppression is studied in this dissertation. The properties of the existing devices for harmonic suppression provide a reference for the new device under study. Therefore, after reviewing fundamentals of the power factor improvement in three-phase non-sinusoidal supply, this chapter summarizes properties of resonant harmonic filters and active filters.

3.2 Harmonic distortion

The current of three-phase non-linear and/or time-variant loads can be decomposed similarly as for linear, time-invariant loads. One new component occurs for such loads apart from the current components identified previously. This new component is referred to as a "generated harmonic current", \( i_g \), [22]. Thus, the current can be expressed as follows:

\[
i = i_a + i_r + i_s + i_u + i_g.
\]  

(41)

Because of the generated harmonic current \( i_g \), non-linear and/or time-variant loads are referred to as harmonic generating loads (HGLs). Because of the mutual orthogonality of all components [22], their rms values fulfill the relation:

\[
||i||^2 = ||i_a||^2 + ||i_r||^2 + ||i_s||^2 + ||i_u||^2 + ||i_g||^2.
\]  

(42)

Then, power factor can be expressed in terms of these current components, namely,

\[
\lambda = \frac{P}{S} = \frac{||i_a||}{||i||} = \frac{||i_a||}{\sqrt{||i_a||^2 + ||i_r||^2 + ||i_s||^2 + ||i_u||^2 + ||i_g||^2}} = \frac{1}{\sqrt{1 + \frac{||i_r||^2 + ||i_s||^2 + ||i_u||^2 + ||i_g||^2}{||i_a||^2}}}.
\]  

(43)
Power factor is a measure of the supply source utilization. At a low power factor, the current drawn by the equipment is substantially larger than necessary. This implies increased power losses and increased apparent power ratings of utility equipment such as transformers, transmission lines and generators. Therefore, power factor should be elevated to an economically justifiable value.

A one-line diagram of a three-phase power system, as shown in Figure 3.2.1, is used for the discussion of power factor improvement. The impedance $Z_s$ represents the total equivalent impedance of generators, transformers, transmission lines, etc. as observed from the point of common coupling (PCC) and is referred to as the "distribution system impedance" or "system impedance" in this dissertation. Due to the system impedance, a distorted supply current causes a voltage distortion at the PCC. Consequently, other loads connected to the PCC are operated with a distorted voltage. Therefore, it is desirable that not only the power factor but also the voltage waveform be improved.

![Figure 3.2.1 One-line diagram of simplified power system.](image)

To evaluate the performance of harmonic suppression devices, the supply current harmonic distortion ($CHD$), denoted in percent as $\delta_i$, and voltage harmonic distortion ($VHD$), denoted in percent as $\delta_u$, at PCC can be used. They are defined as follows:
\[
\delta_i = \frac{\sum_{n \neq 1} I_n^2}{I_i} \cdot 100 = \frac{\sqrt{||i||^2 - I_i^2}}{I_i} \cdot 100, \quad (44)
\]

and

\[
\delta_u = \frac{\sum_{n \neq 1} U_n^2}{U_i} \cdot 100 = \frac{\sqrt{||u||^2 - U_i^2}}{U_i} \cdot 100. \quad (45)
\]

To see how the generated harmonic current deteriorates the power factor, one can consider a simplified situation where the supply voltage is sinusoidal and the load is a balanced HGL. Under these assumptions, eqn. (41) can be simplified to:

\[
i = i_I + i_g = i_u + i_r + i_g, \quad (46)
\]

where \(i_u\) and \(i_r\) are the active current and the reactive current, respectively. In this case, power factor can be calculated as follows:

\[
\lambda = \lambda_I = \frac{P}{S} = \frac{||i_u||}{||i||} \cos \phi_I, \quad (47)
\]

where \(\phi_I\) denotes the phase angle between the voltage \(u\) and current \(i_I\). For linear loads, eqn. (46) has the form:

\[
i = i_I = i_u + i_r, \quad (48)
\]

and the power factor can be simplified to:

\[
\lambda = \lambda_I = \frac{P}{S} = \frac{||i_u||}{||i||} = \cos \phi_I. \quad (49)
\]

By comparing eqn. (47) and (49), the power factor of non-linear loads is worse by a factor of \(||i_I||/||i||\), which is the ratio of the rms value of the fundamental frequency component to the rms value of the supply current. These two rms values differ due to the presence of the generated harmonic current \(i_g\). The current \(i_g\) increases the rms value of the supply current similarly as the reactive current, consequently, it increases the power loss, i.e., extra heating, as well.
Since most devices are designed to be operated at sinusoidal supply voltage and energy is generated, transmitted and distributed with sinusoidal waveforms, the power factor can be approximated in systems with non-sinusoidal supply voltage with eqn. (47). Having the definition for \( CHD \), power factor can be expressed in terms of \( \delta_i \) and \( \cos\phi_i \), namely,

\[
\lambda = \frac{\cos\phi_i}{\sqrt{\left(\frac{\delta_i}{100}\right)^2 + 1}}.
\]  \hspace{1cm} (50)

Equation (50) implies that to improve the power factor not only the reactive power should be compensated but also the current harmonic distortion should be reduced.

Conventionally, harmonic distortion measures, i.e., \( CHD \) and \( VHD \), are defined for single-phase systems or on per phase base for three-phase systems. For three-phase balanced systems, coefficients \( CHD \) and \( VHD \) for only one phase are needed to describe the degree of distortion. However, for three-phase unbalanced systems, three \( CHDs \), phase by phase, namely, \( CHD \) for phase R, S and T, are needed. The level of current distortion could be also specified by the highest \( CHD \) value of the three \( CHD \) phase by phase values. Three phase systems in the dissertation are not handled on phase-by-phase basis, but they are handled as three-phase entities, described in terms of three-phase vectors of voltages and currents. Keeping this approach in mind, the three-phase current distortion could be defined with a single measure, namely,

\[
\delta_i = \frac{\sqrt{|i_i|^2 - |i_i'|^2}}{|i_i|} \cdot 100\%.
\]  \hspace{1cm} (51)

where

\[
|i_i'| = \sqrt{I_{Ri}^2 + I_{Si}^2 + I_{Ti}^2},
\]  \hspace{1cm} (52)

and \( |i| \) is as defined with eqn. (5). Observe, eqn. (51) is similar to the definition used for single-phase systems except the current vector rms values are used. Nevertheless, if
the system is balanced, it is the same as calculated by using the conventional definition. The voltage harmonic distortion, $VHD$, can be defined in a similar way.

3.3 **Resonant harmonic filters**

A single-phase system with a common resistive-inductive HGL has the equivalent circuit shown in Figure 3.3.1. Since the system impedance is usually highly inductive, the system impedance can be simplified for theoretical analysis and be represented by an inductance $L_s$ with zero resistance. The current source in this circuit represents current harmonics, without the fundamental, generated by the HGL. Therefore, an equivalent circuit for the fundamental frequency has the form shown in Figure 3.3.2. Since the impedance of the RL branch increases with frequency increase,

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this branch can be neglected if the harmonic band of frequency is considered. Thus, it can be assumed that the equivalent circuit for harmonic frequencies could be simplified as shown in Figure 3.3.3. In general, the bus voltage $u$ and the supply current distortion is caused not only by the HGL but also by the distortion of the distribution voltage $e$.

![Figure 3.3.3 Equivalent circuit of distribution system with HGL at harmonic frequencies.](image)

Reactive power in systems with sinusoidal waveforms can be compensated simply by connecting a capacitor bank with the appropriate capacitance in parallel with the load. In systems with nonsinusoidal waveforms, such a compensator may cause a resonance in a band of harmonic frequencies. The equivalent circuit of a compensated system is shown in Figure 3.3.4. The resonant frequency $\omega_r$ in such a system is approximately equal to:

$$\omega_r = \frac{1}{\sqrt{L_s C}},$$

(53)

where $C$ is the capacitance needed for full compensation of reactive power. The system inductance $L_s$ depends on the supply system configuration and is beyond the control of the designing engineer. Therefore, the resonant frequency may occur at any frequency.
To see how supply voltage harmonics and load generated current harmonics affect the voltage at PCC and the supply current, the following transmittances:

\[ A(j\omega) = \frac{U(j\omega)}{E(j\omega)} \]  

(54)

and

\[ B(j\omega) = \frac{I(j\omega)}{J(j\omega)} \]  

(55)

are defined. The plots of the magnitudes of transmittances \( A(j\omega) \) and \( B(j\omega) \) are shown in Figures 3.3.5 and 3.3.6, respectively. Apart from resonance at frequency \( \omega_r \), amplifications of the distribution voltage harmonics and the load generated current

Figure 3.3.5 Plot of the magnitudes of transmittances \( A(j\omega) \).
harmonics may occur in the presence of the compensating capacitor. These harmonic amplifications appear not only at one frequency, the resonant frequency, but also appear in a wide band of harmonic frequencies. If such a resonance occurs, the supply current may contain unexpectedly large harmonics and the power factor may decline.

Resonant harmonic filters (RHFs) have been installed in distribution systems under nonsinusoidal conditions to protect the system against the injection of load generated current harmonics. They compensate the reactive power as well. Each branch of a RHF provides a low impedance path for a single load generated current harmonic, thus such a current harmonic can bypass the supply source without distorting the supply current and the supply voltage. As long as non-linear loads are the only cause of waveform distortion and RHFs have a sufficient number of branches tuned to frequencies of the load generated harmonics, RHFs are usually effective devices for reducing harmonics. However, RHF effectiveness depends not only on the filter parameters but also on the distribution system impedance. The use of RHFs results in parallel resonances between the filter and the system inductance at a frequency below each tuning frequency. Moreover, each branch of a RHF forms a low impedance sink for the supply voltage harmonics, thus, while reducing the load originated supply current distortion, the filter increases the supply voltage originated distortion of this
current. Since resonant harmonic filters are usually tuned to only a few dominating harmonics, the remaining are not attenuated but even amplified.

To simplify the analysis, the properties of RHFs are discussed below with an example of a two branch RHF tuned to frequencies $\omega_a$ and $\omega_b$. These tuned frequencies are equal to:

$$\omega_a^2 = \frac{1}{L_a C_a}$$  \hspace{1cm} (56)

and

$$\omega_b^2 = \frac{1}{L_b C_b}.$$  \hspace{1cm} (57)

The equivalent circuit of a distribution system with such a RHF at harmonic frequencies is shown in Figure 3.3.7. The plot of reactance as seen from the harmonic current source versus relative frequency is drawn in Figure 3.3.8. It can be seen that there is a resonant frequency below each tuned frequency. These resonant frequencies of the filter with the system inductance $L_s$ are denoted as $\omega_c$ and $\omega_d$. At each of these frequencies,
a very high impedance is seen from the harmonic current source. This implies that even for a low rms value of a current harmonic, the voltage harmonic may have a high rms value. A plot of reactance as seen from the supply voltage source versus relative frequency is drawn in Figure 3.3.9. At resonant frequencies $\omega_a$ and $\omega_b$, low impedance is seen from the supply voltage source $e$. Consequently, voltage harmonic of low rms value can cause current harmonics of a very high rms value.

\[
Y_{out} = \frac{1}{jn\omega_L L_a} + \frac{1}{jn\omega_L L_a + 1/n\omega_C} + \frac{1}{jn\omega_L L_b + 1/n\omega_C} \\
= -j \frac{n^4 \omega_L^4 (L_a + L_b + 1) - n^2 \omega_L^2 (L_a \omega_a^2 + L_b \omega_b^2 + \omega_a^2 + \omega_b^2) + \omega_a^2 \omega_b^2}{n\omega_L (n^2 \omega_a^2 - \omega_a^2) (n^2 \omega_b^2 - \omega_b^2)}.
\]
Then the resonant frequencies $\omega_c$ and $\omega_d$ can be calculated as follows:

$$\omega_c = \sqrt{A - \sqrt{A^2 - F}}, \quad (59)$$

and

$$\omega_d = \sqrt{A + \sqrt{A^2 - F}}, \quad (60)$$

where

$$A = \frac{L_x \omega_b^2 + L_x \omega_d^2 + \omega_a^2 + \omega_b^2}{2(L_x L_a + L_x L_b + 1)}, \quad (61)$$

and

$$F = \frac{\omega_a^2 \omega_b^2}{L_x L_a + L_x L_b + 1}. \quad (62)$$

According to eqns. (59), (60), (61) and (62), resonant frequencies $\omega_c$ and $\omega_d$ are not only functions of the filter parameters but also the system inductance $L_s$. However, the system inductance may not be known by the designing engineer and may change with the supply system configuration, i.e., with the change of the generation pattern and transmission lines in service. If resonant frequencies $\omega_c$ and $\omega_d$ are located closely to frequencies of the load current harmonics, voltage harmonics of high rms values may be observed at the supply terminals. Therefore, filter design must anticipate the changes in system inductance that might occur in the foreseeable future, e.g., an additional interconnection. This is a noticeable disadvantage of RHF. All of these drawbacks reduce filter effectiveness.

### 3.4 Active filters

Active filters (AFs) enable reactive current compensation and harmonic suppression [33-40]. An active filter is built of a voltage- or current-source PWM inverter and a signal processing system (SPS). Its basic structure is illustrated in Figure 3.4.1. A switch-mode PWM inverter forms a current source and injects a compensating
current to the line to ensure that the supply current is equal to the active current, namely,

\[ i = i_L + i_A \]

\[ = (i_{La} + i_{lr} + i_a) + (-i_{lr} - i_a). \]  

\[ = i_{La} \]

Active filters have, however, two main drawbacks. Recent development of IGBTs and GTO thyristors with elevated voltage-current rating and switching speed has spurred greater interest in the study of active power filters. However, it is still difficult to construct a high power active filter with a rapid current response. Secondly, its initial cost and maintenance cost are relatively high. These drawbacks limit the use of active filters in practical applications.

To overcome the first drawback, an active filter is often combined with a passive compensator (PC) such as a RHF. Such a combination is shown in Figure 3.4.2 and is
called a "hybrid compensator". In such a case, a lower rated current source is needed for a hybrid compensator than for an active filter alone. However, their complexity is still high.
CHAPTER 4
HARMONIC BLOCKING COMPENSATORS

4.1 Introduction

Resonant harmonic filters (RHF) s, commonly installed for reducing waveform distortion in distribution systems [12-20], and active filters (AFs), mainly still under development [33-40], have a number of disadvantages as discussed in the previous chapter. This chapter investigates another alternative for harmonic suppression and reactive power compensation, namely, a harmonic blocking compensator. This newly developed compensator is a passive device of relatively simple structure as compared with RHF s and AFs. It compensates the reactive power and it is more effective in suppressing harmonics than traditional RHF s.

A harmonic blocking compensator (HBC) [41-43], built of a series filter tuned to the fundamental frequency and a shunt capacitor, is developed and discussed in this chapter. The series filter of such a compensator acts as a low impedance path at the fundamental frequency and as a harmonic obstructor at harmonic frequencies, meanwhile the capacitor provides the needed susceptance for reactive power compensation at the fundamental frequency, and a low impedance path for the load generated harmonic current at harmonic frequencies. A capacitor can provide only a fixed reactive power compensation, therefore, a HBC should be applied for compensation of fixed loads.

In the next two sections, the following topics are discussed:
• Fundamentals of harmonic blocking compensator operation.
• Properties and design of harmonic blocking compensators.

4.2 Fundamentals of harmonic blocking compensator operation

The fundamentals of operation of a novel compensator for reactive power compensation and harmonic suppression of fixed loads is discussed in this section. To overcome the disadvantage of capacitor banks regarding the resonance with the system
inductance, an inductor $L_o$ can be inserted, as shown in Figure 4.2.1 (a) and (b), to manipulate the location of the resonance. In (a) and (b) of the Figure 4.2.1, the inductor is connected in series and in parallel with respect to the system inductance, respectively. For both cases, assuming that the system inductance is much smaller than the load impedance, the resonant frequency $\omega_r$ can be approximated by:

$$\omega_r = \frac{1}{\sqrt{(L_s + L_o)C}},$$}

where usually $L_o >> L_s$. Unfortunately, in the case of circuit (b), the load generated current harmonics of frequency above this resonant frequency $\omega_r$ will flow through the supply system. Circuit (a) is advantageous in this respect, since the current harmonics will flow through the capacitor branch. However, such a line inductor, as connected in (a), causes a voltage drop so that its inductance cannot be high, consequently, it will
not improve effectiveness of harmonic suppression. A series filter, tuned to the fundamental frequency, connected instead of the inductor makes it possible to increase the series reactance, as shown in Figure 4.2.2, for harmonic frequencies without causing

\[
X(\alpha/\omega_0)
\]

Figure 4.2.2 Reactance vs. frequency plot of the harmonic filter and capacitor.

a noticeable voltage drop on this series branch at fundamental frequency. Having high series reactances for harmonic frequencies, a shunt resonant harmonic filter may not be needed, since a shunt capacitor may provide a low impedance path for all load generated current harmonics. At the same time, the shunt capacitor provides the needed susceptance for reactive power compensation at the fundamental frequency.

A compensator, as shown in Figure 4.2.3, that consists of a series filter tuned to the fundamental frequency and a shunt capacitor is referred to as a "harmonic blocking compensator" (HBC). The series filter of such a compensator acts as a low impedance

\[
\begin{array}{c}
\text{HGL} \\
\text{C} \\
\text{L} \\
\end{array}
\]

Figure 4.2.3 Circuit configuration with a HBC.
path at the fundamental frequency and as a harmonic obstructer at harmonic frequencies, meanwhile the capacitor provides the needed susceptance for reactive power compensation at the fundamental frequency, and a low impedance path for the load generated harmonic current at harmonic frequencies [46]. To adjust the needed parameters of the series filter to the rated values on the market, a transformer can be inserted between the line and the series filter as shown in Figure 4.2.4. With the secondary winding almost short-circuited by the resonant branch at the fundamental frequency, the transformer is operated as a current transformer. Consequently, it can be built as a current transformer, i.e., using a core without an airgap. Moreover, a high permeability toroidal permalloy could be used for such a transformer. Thus, it can have a high magnetizing inductance with a low number of turns. If it is built in this way, the transformer will have a low winding resistance and low active power loss in the core and in the windings. Consequently, such a transformer can be considered as an ideal transformer. The possibility of choosing the turn ratio of such a transformer provides more flexibility in the HBC design. The coupling transformer, however, is not necessary for the compensator's operation.

The equivalent circuit of a HBC with an ideal transformer is shown in Figure 4.2.5. Defining the turn ratio of the transformer as:
where $N_1$ and $N_2$ are the number of turns on the primary and secondary sides, respectively, the equivalent inductance and capacitance are equal to:

$$L_{eq} = a^2 L_f$$

(66)

and

$$C_{eq} = \frac{C_f}{a^2}.$$  \hfill (67)

![Figure 4.2.5 Equivalent circuit of a HBC w/ transformer.](image)

In cases where a transformer is not used, $L_{eq}$ is equal to $L_f$ and $C_{eq}$ is equal to $C_f$.

The performance of the HBC as shown in Figure 4.2.6 can be determined in terms of the following four transmittances:

$$B(j\omega) = \frac{I(j\omega)}{J(j\omega)} = \frac{Z_n(j\omega)}{Z_x(j\omega)}, \quad (68)$$

$$Y_x(j\omega) = \frac{I(j\omega)}{E(j\omega)} = \frac{1}{Z_x(j\omega)}, \quad (69)$$

$$T(j\omega) = -\frac{U(j\omega)}{J(j\omega)} = -\frac{Z_n(j\omega)Z_q(j\omega)}{Z_x(j\omega)}, \quad (70)$$

and

$$A(j\omega) = \frac{U(j\omega)}{E(j\omega)} = \frac{Z_n(j\omega) + Z_q(j\omega)}{Z_x(j\omega)}, \quad (71)$$

where

$$Z_L(j\omega) = R + j\omega L,$$  \hfill (72)
\[ Z_s(j\omega) = R_s + j\omega L_s, \quad (73) \]
\[ Z_L(j\omega) = R_{eq} + j(\omega L_{eq} - \frac{1}{\omega C_{eq}}), \quad (74) \]
\[ Z_a(j\omega) = \frac{Z_C(j\omega)Z_L(j\omega)}{Z_C(j\omega) + Z_L(j\omega)}, \quad (75) \]

and

\[ Z_x(j\omega) = Z_s(j\omega) + Z_{eq}(j\omega) + Z_a(j\omega). \quad (76) \]

Figure 4.2.6 Circuit configuration with the equivalent circuit of a HBC.

To obtain general results of respective transmittances specified with eqns. (68)-(71), analysis below is in 'per unit'. It is assumed for that purpose that the rms values of the load voltage and load current fundamental harmonic \( U_f \) and \( I_f \) before compensation are reference bases for per unit calculation, i.e., \( U_f = 1 \)pu and \( I_f = 1 \)pu. Consequently, the apparent power at the fundamental frequency is:
\[ S_f = U_f I_f = 1 \]pu \quad (77)\]

and the absolute value of the load impedance for the fundamental frequency is:
\[ |Z_{L_f}| = \sqrt{R^2 + (\omega_f L)^2} = \frac{U_f}{I_f} = 1 \]pu. \quad (78)\]

The angular fundamental frequency \( \omega_f \) is normalized to 1 rad/sec.

To demonstrate the frequency properties of HBCs, a system with the following assumptions has been modeled: the supply source short circuit power \( S_{sc} \) equal to 25pu,
i.e., the voltage loss on the source impedance is of the order of 4%. It is also assumed that the system reactance to resistance ratio is $X_s/R_s = 5$ and the load has the form of a series RL load with power factor 0.71 lagging at the fundamental frequency. A HBC for such a load is designed for total reactive power compensation and assuming that $C_{eq} = C$. It is assumed that the filter inductor has a quality factor of 50, and the power loss in capacitors of the HBC is neglected. The plot of the magnitude of transmittance $B(j\omega)$ versus relative frequency is shown in Figure 4.2.7. It should be noted that the level of $|B(j\omega)| = 1$ separates the bands of frequency where the supply current harmonics are higher than the load generated current harmonics, i.e., the amplification bands, from the bands of harmonic attenuation. Observe, that the values of transmittance magnitude $|B(j\omega)|$ for HBC are substantially reduced as compared to RHF in the whole harmonic band. It implies that a HBC reduces the supply current harmonics caused by the load originated current harmonics without any amplification, unlike the RHF, where some harmonic amplification bands may occur. Moreover, as shown in Figure 4.2.8, drawn for systems with different short-circuit power, the values of transmittance magnitude $|B(j\omega)|$ for the HBC are nearly independent of the distribution system, thus any reconfiguration in the distribution system does not affect HBC performance.
Figure 4.2.8 Plots of $|B(j\omega)|$ with different system impedance: $S_{sc} = 12.5$ and $100$.

The plot of the magnitude of transmittance $T(j\omega)$ versus relative frequency is shown in Figure 4.2.9. Due to low values of $|T(j\omega)|$ at all frequencies, the voltage at PCC is not likely to be affected by the load generated current harmonics.

Figure 4.2.9 Magnitude plot of transmittance $T(j\omega)$ with specified parameters.

The plot of the magnitude of admittance $Y_x(j\omega)$ versus relative frequency is shown in Figure 4.2.10. The values of $|Y_x(j\omega)|$ for the system with the HBC are much lower than for the system with the RHF. This implies that the supply voltage harmonics do not deteriorate the HBC's performance.
Figure 4.2.10 Magnitude plot of transmittance $Y_x(j\omega)$ with specified parameters.

The plot of the magnitude of transmittance $A(j\omega)$ versus relative frequency is shown in Figure 4.2.11. Since the values of $|A(j\omega)|$ are less than one in the band of harmonic frequencies, any amplification of the distribution voltage harmonics cannot occur.

4.3 Properties and design of harmonic blocking compensators

Harmonic blocking compensators are very effective devices for the suppression of harmonic distortion caused by non-linear loads and they do not cause resonances with the distribution system inductance. These advantages are obtained, however, at the price of the higher power ratings of compensator components than traditional RHFs, an
extra active power loss on the inductor of the series filter and an extra voltage drop across the series filter branch. The design of harmonic blocking compensators is a trade off between the effectiveness of HBCs and the cost, which can be decomposed into investment cost and operating cost, and is discussed in detail in this section.

The parameters for a transformerless HBC include the capacitances of the shunt capacitor $C$ and the series filter capacitor $C_{eq}$ and the inductance of the series filter inductor $L_{eq}$. As mentioned earlier, the capacitance $C$ needed should fully compensate the load reactive power to minimize the supply current. If the power factor of the load for the fundamental frequency $\lambda_i$ is expressed as $\lambda_i = P_I/S_I$, where $P_I$ and $S_I$ are the fundamental components of the load active and apparent powers before compensation, respectively. Then, the capacitance $C$ needed is equal in per unit to

$$C = \frac{Q_I}{\omega_l U_I^2} = \sqrt{1 - \lambda_i^2}. \quad (79)$$

Therefore, this capacitance is not chosen by the designer but is determined rather by the load reactive power. The shunt capacitor reduces the supply current fundamental harmonic $i_I$ to its active component $i_{aI}$ of the rms value

$$i_{aI} = \lambda_i i_I. \quad (80)$$

Since the series filter is tuned to the fundamental frequency, i.e., in per unit $L_{eq}C_{eq} = 1/\omega_l^2 = 1$, thus only parameter, either $L_{eq}$ or $C_{eq}$, can be chosen at the discretion of the designer. The filter capacitance $C_{eq}$ can be such a parameter. Let it be proportional to the shunt capacitance. Namely,

$$C_{eq} = \delta C. \quad (81)$$

The parameters used in the previous section to obtain the transmittance plots, for instance, were chosen as $\delta = 1$. This $\delta$ value, the ratio of capacitances $C_{eq}$ and $C$, is important for the design of HBCs. A set of transmittance plots, shown in Figure 4.3.1,
are obtained with the same parameters as previously used except that the $\delta$ is increased from 1 to 1.2. The increase in $\delta$ value shifts resonances of the HBC towards higher frequencies and consequently it increases the magnitude of transmittance $|B(j\omega)|$ and others. There resonances occur at frequencies where the impedance $Z_q(j\omega)$, as defined in eqn. (75), has the minimum value. At the assumption that the system inductance is much less than the series filter inductance, i.e., $L_s << L_e$, and neglecting all resistances of the circuit, the values of the resonant frequencies $\omega_p$ can be calculated approximately by solving the following equation:

$$\left(\frac{\omega_p}{\omega_1}\right)^4 - \left(1 + \delta + \frac{1}{1 - \lambda_1^2}\right) \left(\frac{\omega_p}{\omega_1}\right)^2 + \frac{1}{1 - \lambda_1^2} = 0.$$  

(82)

For $\lambda_1 = 0.71$ and $\delta = 1$, eqn. (82) results in the resonant frequencies $\omega_p$ equal $0.76\omega_1$ and $1.85\omega_1$. However, these values are an approximation of the actual values, which are slightly higher than the approximated values.

The power ratings of HBCs include the power ratings of shunt capacitor and elements in the resonant filter. These ratings are approximately equal to the apparent powers of the fundamental frequency. The per unit fundamental apparent power $S_{Cl}$ of the shunt capacitor can be calculated, under previous assumptions, as follows:

$$S_{Cl} = U_{Load}I_{Cl} = U_1I_{Cl} = U_1^2\omega_1C = \sqrt{1 - \lambda_1^2}.$$  

(83)
The per unit fundamental apparent power $S_{Ceql}$ of the filter capacitor can be calculated as follows:

$$S_{Ceql} = U_{Ceql} I_{al} = \frac{I_{al}}{\omega_1 C_{eq}} I_{al} = \frac{\lambda^2_i}{\delta \sqrt{1 - \lambda^2_i}}.$$  \hspace{1cm} (84)

Since the rms values of the filter inductor and capacitor voltages at the fundamental frequency are mutually equal, eqn. (84) also specifies the per unit fundamental apparent power $S_{Leql}$ of the filter inductor.

Equation (84) shows that the required power ratings of the series filter increase with power factor increase. This implies that a HBC with lower power ratings, i.e., less expensive, is required for loads with lower power factor. However, eqns. (83) and (84) provide only an approximation of the required power ratings of the HBC components. To calculate these ratings more accurately, harmonic components of the voltage and current have to be taken into account, i.e., the power rating of the shunt capacitor has to be calculated with following formula

$$S_C = ||u_C|| ||i_C||.$$  \hspace{1cm} (85)

At a fixed value of the load power factor, the power ratings of the series filter can be reduced, according to eqn. (84), only by increasing the ratio of capacitances, $C_{eq}/C = \delta$. However, this reduces the effectiveness of the HBC. The change in the transmittance magnitude $|B(j\omega)|$ for the same circuit parameters as considered previously but with $\delta = 1.2$ is shown in Figure 4.3.2. Since $|B(j2)| > 1$, an amplification of the second order harmonic of the load current occurs. The second order harmonic usually has relatively low value in distribution systems, thus its amplification is not a substantial disadvantage for HBCs.
In systems with a low value for the third order harmonic, the power rating of the series filter can be reduced by shifting the resonant frequency to the band between the second and the third order harmonic, or even above the third harmonic. If the resonant frequency \( \omega_p \) and power factor \( \lambda_i \) are given, then the approximated value of the ratio \( \delta \) can be calculated from eqn. (82), namely,

\[
\delta = \left( \frac{\omega_L}{\omega_i} \right)^2 - \frac{1}{1 - \lambda_i^2} \left[ 1 - \left( \frac{\omega_L}{\omega_p} \right)^2 \right] - 1.
\]

For example, for \( \lambda_i = 0.71 \) the resonant frequency \( \omega_p \) is located at \( 2.5 \omega_i \) if \( \delta = 3.6 \).

The choice of a higher value of the capacitance ratio \( \delta \) is also beneficial for reducing the active power loss of the compensator, in particular, the power loss of the series filter inductor. Assuming that the load reactive power of the fundamental frequency is entirely compensated and all load current harmonics flow through the shunt capacitor, the active power loss in the series inductor with resistance \( R_{eq} \) is equal to

\[
\Delta P = R_{eq}(I_{al})^2 = R_{eq}(\lambda_i I_i)^2 = \frac{\omega_i L_{eq} (\lambda_i I_i)^2}{q},
\]

where \( q \) denotes inductor's quality factor, i.e., \( q = \omega_i L_{eq} / R_{eq} \). Taking into account eqns. (79) and (81), this power loss in per unit value can be expressed as
This equation shows that for a given load, i.e., the power factor $\lambda_j$, and a specified capacitance ratio $\delta$, the power loss $\Delta P$ can be reduced only by increasing the quality factor of the inductor.

An important feature of HBCs that has to be taken into account at their design and implementation is the voltage drop $\Delta U_t$ across the series filter since it reduces the fundamental harmonic of the load voltage. At total compensation of the load reactive power, the relative voltage drop in per unit is equal to

$$\frac{\Delta U_t}{U_t} = \frac{R_{eq}I_{al}}{U_t} = R_{eq} \lambda_j.$$  \hspace{1cm} (89)

It can be expressed in terms of inductor quality factor $q$ and the capacitance ratio $\delta$ as follows:

$$\frac{\Delta U_t}{U_t} = \frac{\omega L_{eq}}{q} \lambda_j = \frac{\lambda_j}{q C_{eq}} = \frac{\lambda_j}{q \delta \sqrt{1 - \lambda_j^2}}.$$  \hspace{1cm} (90)

Thus the voltage drop declines when the load power factor $\lambda_j$ for the fundamental frequency declines and when the inductor's quality factor $q$ increases. Moreover, it can be reduced at the designer's discretion by a choice of higher value of the capacitance ratio $\delta$. However, the choice of a higher value of this ratio increases the resonant frequency $\omega_r$, and this reduces the effectiveness of the harmonic suppression of HBCs. The only other possibility of the reduction of the voltage drop as well as the reduction of active power loss is the elevation of the inductor's quality factor $q$. To avoid ferroresonance phenomenon caused by the inductor core saturation, the inductor associated with HBCs has to be built as a linear inductor, i.e., with a sufficiently large air gap. This air gap should be optimized to obtain the quality factor as high as possible.

An inductor's quality factor $q$ declines due to an increase in the power loss in the winding and in the magnetic core. For a given core, these losses depend on the number
of turns and the air gap of the core. By a proper choice of the turn number and a proper adjustment of the air gap, the quality factor $q$ can be enhanced to its maximum value. To find the optimum parameters of the inductor, the equivalent circuit of the resonance branch, as shown in Figure 4.3.3, will be used [44,45]. In this figure, resistance $R_w$ is the winding resistance and resistance $R_c$ is the equivalent resistance with respect to the active power dissipated in the core of the inductor. The impedance of the resonance branch is equal to

$$Z(j\omega) = \frac{j\omega(R_wR_c + L_{eq}) + \frac{R_c}{C_{eq}} - \omega^2(R_w + R_c)L_{eq}}{j\omega(R_c + j\omega L_{eq})}.$$  \hfill (91)

In the dominator, reactance $\omega L_{eq}$ to resistance $R_c$ ratio for the fundamental frequency, i.e., $f_j = 60\text{Hz}$, could be of the order of $10^{-2}$ and the winding resistance $R_w$ could be of the order of $10^{-3}R_c$. Therefore, if the branch is tuned to the fundamental frequency,

$$\omega = \omega_j = \frac{1}{\sqrt{L_{eq}C_{eq}}},$$  \hfill (92)

its impedance at this frequency is approximately equal to

$$Z(j\omega_j) = R_w + \frac{(\omega_j L_{eq})^2}{R_c} = R_{eq},$$  \hfill (93)

where $R_{eq}$ is referred to as resonance resistance. The magnitude of the branch impedance around the fundamental frequency can be expressed as

$$|Z(j\omega)| = R_{eq}\sqrt{1 + q_{eq}^2\left(\frac{\omega}{\omega_j} - \frac{\omega_j}{\omega}\right)^2},$$  \hfill (94)
where

\[ q_{eq} = \frac{\omega I L_{eq}}{R_{eq}} = \frac{q_w q_c}{q_w + q_c} \]  \hspace{1cm} (95)

is referred to as the quality factor of the resonance branch, and

\[ q_w = \frac{\omega I L_{eq}}{R_w} \]  \hspace{1cm} (96)

and

\[ q_c = \frac{R_c}{\omega_1 L_{eq}} \]  \hspace{1cm} (97)

are referred to as the quality factors of the winding and the core, respectively.

To roughly estimate the resonance resistance \( R_{eq} \) of the branch and its quality factor \( q_{eq} \), as well as their dependence on the technical features of the inductor, the following assumptions have been made. It is assumed that the inductor has a very simple geometry, as shown in Figure 4.3.4, with the same core and the same winding

window square cross-sections, namely, \( s = b^2 \). This assumption simplifies the relation between quality factors \( q_w, q_c, q_{eq} \) and the inductor's dimensions. The inductor of the resonance branch should be built as a linear inductor, so that the reluctance of the air gap is much higher than the reluctance of the core. Therefore, the reluctance of the core is neglected for the inductance calculation.

At the inductor voltage rms value \( U_L \) and the maximum value of the magnetic flux density \( B_m \) the number of turns of the winding is equal to
\[ N = 2.25 \times 10^3 \frac{U_L}{f_i B_m b^2}. \]  

(98)

If \( \Delta \) denotes the width in millimeters of the core air gap, its reluctance is approximately equal to

\[ \varepsilon = 10 \frac{\Delta}{\mu_0 b^2} = 7.96 \times 10^6 \frac{\Delta}{b^2}, \quad (1/\text{H}) \]

(99)

and hence

\[ L_{eq} = \frac{N^2}{\varepsilon} = 0.636 \frac{U_L^2}{f_i^2 B_m^2 b^2} \frac{1}{\Delta}. \quad (\text{H}) \]

(100)

Assuming that the winding has two coils, the average length of a single turn of the winding is \( 6b \), so that the winding conductor length is \( l_w = 6bN10^{-2} \) (m). The conductor has cross-section \( s_w = kb^210^2\) (mm²) where coefficient \( k \) is referred to as the 'winding factor', which specifies utilization of the winding window. At the copper conductor resistivity equal to 0.018 (\( \Omega \text{mm}^2/\text{m} \)), the winding resistance is equal to

\[ R_w = 0.018 \frac{l_w}{s_w} = 55 \frac{U_L^2}{f_i^2 B_m^2 kb^5}. \]

(101)

The active power loss on resistance \( R_c \) at the inductor voltage rms value \( U_L \) is equal to its loss in the magnetic core, i.e., \( P_c = U_L^2/R_c \). It is assumed that the loss of the active power in the core changes with the square of the maximum value of the magnetic flux density \( B_m \), namely, for a core of the weight \( W \) it is equal to

\[ P_c = p_r (\frac{B_m}{B_r})^2 W, \]

(102)

where \( p_r \) is the active power loss in the core per one kg of its weight at the flux density \( B_r \) and frequency \( f_i \). For M-5 grain-oriented steel of the thickness 0.3 (mm) as used in transformers and at frequency \( f_i = 60\text{Hz} \), the value of \( p_r \) at \( B_r = 1.0 \) (Wb/m²) is approximately equal to 0.55 (W/kg). Since the core volume is \( 8b^2 \), thus its weight in kg is \( W = 7640 \times 10^{-6}b^3 \), hence, the resistance \( R_c \) is equal to
\[ R_c = \frac{U_i^2}{P_c} = 16.4 \frac{U_i^2}{P_r b^3} \text{B_m}^2. \]  

(103)

Having the values of the winding resistance \( R_w \) and its inductance \( L_{eq} \), the quality factor \( q_{w} \) can be expressed as follows

\[ q_w = \frac{\omega L_{eq}}{R_w} = 0.073 f_{r} k b^3 \frac{1}{\Delta}. \]  

(104)

Similarly, having resistance \( R_c \) given by eqn. (103), the quality factor \( q_{c} \) is equal to

\[ q_c = \frac{R_c}{\omega L_{eq}} = 4.1 \frac{f_{i}}{P_r b} B_{r}^2 \Delta. \]  

(105)

The quality factor \( q_{eq} \) of the resonance branch is lower than quality factors \( q_{w} \) and \( q_{c} \). It is maximum when quality factors \( q_{w} \) and \( q_{c} \) are mutually equal, i.e., \( q_{w} = q_{c} \). Since \( q_{c} \) increases with the air gap increase while \( q_{w} \) declines, there is an optimum length of the air gap, denoted by \( \Delta_0 \), when this condition is fulfilled. Namely, the optimum length of the air gap is equal to

\[ \Delta_0 = 0.133 \frac{b^2}{B_r} \sqrt{kp_r}. \]  

(106)

The quality factor of the resonance branch at such an air gap has the maximum value, referred to as the 'optimized \( q \)-factor' and denoted by \( q_o \). It is equal to

\[ q_o = \frac{q_w}{2} = \frac{q_c}{2} = \frac{1}{2} (4.1 \frac{f_{i}}{P_r b} B_{r}^2 \Delta_0) = 0.27 f_{r} B_{b} \frac{k}{P_r}. \]  

(107)

Thus, the optimized \( q \)-factor of the resonant link, \( q_o \), increases with the value of \( b \), i.e., with the physical dimensions of the core.

The inductance at this optimum air gap, referred to as the 'optimized inductance' and denoted by \( L_o \) has the value

\[ L_o = 4.78 \frac{B_r U_i^2}{f_r B_{m}^2 b^4 \sqrt{kp_r}}. \]  

(108)

The resistance at this optimum air gap, referred to as the 'optimized resistance' and denoted by \( R_o \) has the value
Although the term "optimized" is used above, it should be noted that optimization is usually defined as the adjustment of parameters to achieve the most favorable or advantageous design with respect to some special limits, for example, in this case, the length of the inductor air gap. Of course, other optimization can still be applied to such an inductor, for instance, a change in the proportion of the core cross-section area to the area of the winding window may still increase the quality factor $q_\circ$.

The coupling transformer associated with HBCs does not affect the operation principle of HBCs. It only enables HBCs to separate the resonant filter from line voltages and to ground this filter for safety purpose. Also, this transformer makes it possible to adjust the voltage rating and the needed capacitance of the series capacitor to market value. The only difference in frequency properties of the HBC with the coupling transformer as compared with the transformerless HBC is the presence of a path for dc current which might be generated in the load non-linearity. Consequently, the coupling transformer affects the properties of a HBC in a sub-harmonic band of frequency.

Since the resonant filter branch is tuned to the fundamental frequency, the coupling transformer operates in a short-circuit mode at this frequency. In this sense, such a transformer can be designed similarly as a current transformer. As a current transformer, the coupling transformer may have the same kind of magnetic core, i.e., toroidal without an air gap. A high quality permalloy core with permeability of the order of $10^5$ makes it possible to reduce the stray inductance of the winding well below the magnetizing inductance. Since the coupling transformer operates in an almost short-circuit mode at the fundamental frequency, with low voltage drop across the primary winding, it may have a low number of turns. Thus, the coupling transformer has a low resistance and, consequently, a low active power loss which is much lower than the

$$R_\circ = \frac{\omega_i L_o}{q_\circ} = 111 \frac{B^2 U_o^2}{f_i^2 B_m^2 b^2 k}.$$  (109)
active power loss on the series inductor as specified in eqn. (88). It can be considered as an ideal transformer.

The design of a HBC with transformer remains the same as compared with a transformerless HBC. For both cases, the designer determines the capacitance ratio, \( \delta = \frac{C_{eq}}{C} \). The only difference is that in the latter case,

\[
C_f = C_{eq}, \tag{110}
\]

and

\[
L_f = L_{eq}, \tag{111}
\]

and in the former case,

\[
C_f = a^2 C_{eq}, \tag{112}
\]

and

\[
L_f = \frac{L_{eq}}{a^2}. \tag{113}
\]

However, the coupling transformer increases the power loss and the compensator cost. Thus the usage of this transformer should be a matter of individual assessment of benefits and the cost.

The design of a HBC for a particular HGL are concluded as follows: First of all, the voltage and current harmonic spectra of the load should be known. The capacitance of the shunt capacitor should provide total compensation of reactive power to minimize the supply current. At first, a transformerless HBC should be considered. The choice of parameters of a HBC is a matter of the trade-off between the cost of components related to their power ratings and effectiveness of harmonic suppression. To have a higher harmonic attenuation, a lower resonant frequency \( \omega_p \), or a lower capacitance for the filter capacitor, is needed. However, this implies that a higher ratings on the series components are needed and higher loss on the series inductor is expected. After choosing an acceptable current harmonic distortion, the resonant frequency \( \omega_p \),
parameters and ratings of the series filter can be calculated approximately. To adjust the needed parameters to the rated values on the market, a transformer may be needed. If a transformer is needed, the actual parameters of the series filter can be calculated according to the transformer turn ratio.

The following numerical example demonstrates that a HBC can outperform a capacitor compensator and a RHF.

Numerical illustration: A single-phase ac/dc converter is connected to a distorted supply voltage with $U_{st}=120V$, $U_{sf}=3.6V$, $U_{sf}=3.6V$ and 0.3mH system inductance, and the harmonic measurements of voltage and current at PCC are listed as $U_i=118.3V$, $U_j=2.2V$, $U_k=5.3V$, $U_l=5.2V$, $U_m=2.2V$, $U_n=1.3V$, $U_{ij}=2.2V$, $I_j=19.7A$, $I_k=6.5A$, $I_l=3.7A$, $I_m=2.7A$, $I_n=2.1A$, $I_{ij}=1.7A$, $I_{ik}=1.4A$, and $\phi_j = 45^\circ$ lagging.

The voltage and current waveforms obtained from PSpice simulation for the uncompensated load are shown in Figure 4.3.5. The current harmonic distortion of the supply current is 43.2%, the voltage harmonic distortion is 7.2% at PCC and the power factor $\lambda_i$ is 0.65. If a capacitor of 313μF alone is used for total reactive power compensation, it results in a resonance that occurs at 8.7$\omega_i$, causes current harmonic

![Figure 4.3.5 Voltage and current waveforms of circuit without compensator.](image-url)
amplifications around the resonant frequency and causes the current and voltage harmonic distortions to increase to 188% and 22.1%, respectively. Also, the power factor becomes worse, 0.47. If a double tuned resonant harmonic filter tuned to the third and the fifth order harmonics is used to compensate the reactive power and to eliminate the two dominating load generated current harmonics, 33% and 19% with respect to the fundamental. While reducing load generated current harmonics, it introduces an increase of the fifth order current harmonic caused by the supply voltage harmonic. In this case, the voltage harmonic distortion is improved to 4.1% and the power factor is improved to 0.91, however, the current harmonic distortion is increased to 44.3%. Then a HBC with $\delta = 0.5$ is used for reactive power compensation and harmonic suppression. The voltage and current waveforms obtained from PSpice simulation for a compensated load with such a HBC are shown in Figure 4.3.6. The current and voltage harmonic distortions are improved to 4.4% and 4.1%, respectively, and the power factor is almost 1. Observe, not only the CHD is improved but the VHD is also improved.

![Figure 4.3.6 Voltage and current waveforms of circuit with HBC.](image)

The supply current spectra for an uncompensated ac/dc single-phase converter, a converter compensated with a capacitor, a RHF and a HBC as discussed above are
shown in Figure 4.3.7. The values of \( CHD \), \( VHD \) and power factor are compiled in Table 4.3.1.

![Figure 4.3.7 Spectra of current harmonic components.](image)

Table 4.3.1 Comparison of different compensators.

<table>
<thead>
<tr>
<th></th>
<th>w/o Comp.</th>
<th>w/ Cap.</th>
<th>w/ RHF</th>
<th>w/ HBC</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \delta_i )</td>
<td>43.2</td>
<td>188.0</td>
<td>44.3</td>
<td>4.2</td>
</tr>
<tr>
<td>( \delta_u )</td>
<td>7.2</td>
<td>22.1</td>
<td>4.1</td>
<td>4.1</td>
</tr>
<tr>
<td>( \lambda_i )</td>
<td>0.65</td>
<td>0.47</td>
<td>0.91</td>
<td>1.0</td>
</tr>
</tbody>
</table>

As shown in Table 4.3.1, a capacitor compensator does not improve the power factor because it causes a resonance and it increases the \( CHD \) and \( VHD \). A RHF reduces load generated current harmonics but increases current harmonics caused by the supply voltage. Therefore, a RHF may improve the power factor but not the \( CHD \). The harmonic blocking compensator provides remarkably better performance without having the disadvantages of RHFs. It improves the power factor of the system by compensating the reactive power, reducing the current distortion, and eliminating the effect of the system inductance.

For the same HGL, with the capacitance ratio changed, compensator parameters, ratings, power loss, voltage drop, \( VHD \) and \( CHD \) are obtained using PSpice, calculated
using associated equations and compiled in Table 4.3.2. As discussed in the previous section, the power ratings of the resonant filter components, the active power loss of the inductor and the voltage drop across the resonant filter are inversely proportional to the capacitance ratio. However, the current harmonic distortion of the supply current increases with the capacitance ratio increase.

Table 4.3.2 Comparison with increasing $\delta$.

<table>
<thead>
<tr>
<th>$\delta$</th>
<th>$C_{eq}$ (µF)</th>
<th>$S_{Ceq}$ (VA)</th>
<th>$S_{Ceql}$ (VA)</th>
<th>$L_{eq}$ (mH)</th>
<th>$S_{Leq}$ (VA)</th>
<th>$S_{Leql}$ (VA)</th>
<th>$\Delta P$ (W)</th>
<th>$\frac{\Delta U_l}{U_l}$ (%)</th>
<th>$\delta_u$ (%)</th>
<th>$\delta_i$ (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.5</td>
<td>156.5</td>
<td>3131</td>
<td>3126</td>
<td>44.96</td>
<td>3162</td>
<td>3129</td>
<td>62.9</td>
<td>0.039</td>
<td>4.1</td>
<td>4.2</td>
</tr>
<tr>
<td>0.6</td>
<td>187.8</td>
<td>2645</td>
<td>2642</td>
<td>37.46</td>
<td>2682</td>
<td>2644</td>
<td>52.6</td>
<td>0.032</td>
<td>4.1</td>
<td>5.1</td>
</tr>
<tr>
<td>0.7</td>
<td>219.1</td>
<td>2288</td>
<td>2285</td>
<td>32.11</td>
<td>2331</td>
<td>2286</td>
<td>45.4</td>
<td>0.028</td>
<td>4.1</td>
<td>6.0</td>
</tr>
<tr>
<td>0.8</td>
<td>250.4</td>
<td>2018</td>
<td>2013</td>
<td>28.1</td>
<td>2067</td>
<td>2015</td>
<td>40.1</td>
<td>0.024</td>
<td>4.1</td>
<td>6.9</td>
</tr>
<tr>
<td>1</td>
<td>313</td>
<td>1633</td>
<td>1626</td>
<td>22.48</td>
<td>1697</td>
<td>1627</td>
<td>32.9</td>
<td>0.02</td>
<td>4.1</td>
<td>8.8</td>
</tr>
<tr>
<td>1.2</td>
<td>375.6</td>
<td>1456</td>
<td>1364</td>
<td>18.73</td>
<td>1453</td>
<td>1365</td>
<td>27.4</td>
<td>0.016</td>
<td>4.1</td>
<td>11.0</td>
</tr>
</tbody>
</table>

From eqns. (84), (88) and (90), one may notice that the power ratings of the resonant filter components, the active power loss of the inductor and the voltage drop across the resonant filter decline dramatically with the decreasing of the load power factor $\lambda_1$, as shown in Figure 4.3.8. Due to a lower power rating, a lower active power loss and a lower voltage drop, loads with a lower power factor could be a prospective area for HBCs.

![Figure 4.3.8 Change of ratings with respect to power factor.](image)
Moreover, distortion of the load current may change in a broad range, and only in some short intervals of time this distortion is unacceptably high, so that, there is no need for a permanent harmonic reduction. In such a case, the series resonant filter could be by-passed by a switch connected in parallel as shown in Figure 4.3.9. This could reduce the active power loss on its inductor. Also, systems with existing capacitor banks could be equipped with the series resonant filter and converted into a harmonic blocking compensator.

![Figure 4.3.9 Modification of HBC.](image)

For three-phase systems, three HBCs with shunt capacitors connected in a delta configuration, as shown in Figure 4.3.10, can be used for reactive power compensation.

![Figure 4.3.10 Circuit configuration of a 3-phase HBC.](image)
and harmonic suppression. The results from computer simulation using PSpice are obtained. The supply voltage and current waveforms of a three-phase harmonic generating load and the compensated system with a three-phase HBC are shown in Figure 4.3.11 and 4.3.12, respectively. It improves the CHD from 28.9% to 0.7% and the VHD from 5.1% to 4.2%

Figure 4.3.11 Supply voltage and current waveforms of 3-phase AC/DC converter.

Figure 4.3.12 Supply voltage and current waveforms of the system with 3-phase HBC.

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CHAPTER 5
ADAPTIVE HARMONIC BLOCKING COMPENSATORS

5.1 Introduction
The dissemination of controlled power electronic equipment causes higher variability of reactive power and waveform distortions than before. This variability may require that compensators and harmonic suppression equipment have an adaptive property. Traditionally used resonant harmonic filters [12-20] and newly developed harmonic blocking compensators [41-43] do not have adaptive properties. Also, RHFs usually have too many components to convert them into adaptive devices.

The loads could be also unbalanced. Therefore, it may be desirable for an adaptive compensator to have the capability for reducing reactive current, unbalanced current and harmonic distortion at the same time. Inverter based active filters (AFs) or hybrid compensators, built of AFs and passive compensators, can be used for that purpose [33-40], however, they are more complex than passive compensators. Entirely passive devices can be used for that purpose as well. The development of a passive compensator with adaptive capability is the subject of this chapter.

The harmonic blocking compensator discussed in the previous chapter does not have adaptive properties. Moreover, it cannot provide any inductive susceptance at the fundamental frequency. Consequently, a HBC cannot be used as a balancing compensator which requires negative susceptance. The conversion of the harmonic blocking compensator into an adaptive device which can be used for compensation of reactive and unbalanced powers and harmonic suppression is discussed in this chapter. Such a compensator will be referred to as an "adaptive harmonic blocking compensator" (AHBC).

The adaptive harmonic blocking compensator integrates a harmonic blocking compensator with an adaptive balancing compensator (ABC) [28-32]. Such a compensator can be used for compensating reactive power, for symmetrizing load
currents and for protecting the system against load generated current harmonics. It can reduce not only current harmonic distortion but also the voltage harmonic distortion caused by harmonic generating loads.

5.2 Fundamentals of adaptive harmonic blocking compensator operation

As shown in the previous chapter, harmonic blocking compensators have higher effectiveness in harmonic suppression than resonant harmonic filters. Also, they do not cause resonances with distribution system inductance. Such resonances are common in systems with capacitor banks. To have an adaptive compensator with the capability of not only reactive power compensation and harmonic suppression but also load balancing, the HBC could be equipped with a thyristor-switched inductor (TSI) in parallel with the shunt capacitors as shown in Figure 5.2.1. The shunt one-port, built of a capacitor and a TSI, was discussed in Chapter 2. It is called the "thyristor controlled susceptance (TCS) one-port with the basic structure" (TCS_B). Therefore, this compensator is referred to as an "adaptive harmonic blocking compensator with basic structure of TCS" (AHBC_B). It can be analyzed as a device built of three single-phase

Figure 5.2.1 Circuit configuration of a three-phase AHBC_B.
compensators. To simplify the analysis, a single-phase circuit configuration is used, as shown in Figure 5.2.2.

![Figure 5.2.2 Circuit configuration of a single-phase AHBC_B.](image)

The variable susceptance of the compensator is provided by the one-port, built of a capacitor \( C \) and a TSI. The frequency properties of such a compensator can be specified in terms of the susceptance as seen from the load terminals, as shown in Figure 5.2.3. As discussed in Chapter 2, the TSI generates a third order harmonic current which is injected into the supply system. This is an essential drawback of the structure discussed.

![Figure 5.2.3 Plots of susceptance seen from load terminals vs. frequency.](image)
To reduce the third harmonic current injected by the TSI of the one-port into the system, a third harmonic filter can be added and connected in parallel with the TSI as shown in Figure 5.2.4. Such an AHBC is referred to as an "adaptive harmonic blocking compensator with third harmonic filter" (AHBC_3HF).

![Figure 5.2.4 Single-phase circuit configuration of an AHBC_3HF.](image)

The plots of susceptance of this one-port with the thyristor switched completely ON and OFF are shown in Figure 5.2.5.

The susceptance of the one-port is of a minimum value equal to

\[ T_a = \omega_1 C + \frac{9\omega_1 C_2}{8} - \frac{1}{\omega_1 L_c}, \]

(114)

when the thyristors are switched ON over the entire half of the period, i.e., at the firing angle \( \alpha = 90^\circ \). When the thyristors are in the OFF state over the whole period, i.e., at the firing angle \( \alpha = 180^\circ \), the susceptance of the one-port has the maximum value

\[ T_b = \omega_1 C + \frac{9\omega_1 C_2}{8}. \]

(115)

For \( 90^\circ \leq \alpha \leq 180^\circ \), the susceptance \( T \) for the fundamental frequency has a value between \( T_a \) and \( T_b \), as shown in Figure 5.2.5.
The compensator susceptance as seen from the load terminals is shown in Figure 5.2.6.

Figure 5.2.5 Plot of susceptances of the one-port in AHBC_3HF vs. frequency in ON and OFF states.

Figure 5.2.6 Plot of susceptances of AHBC_3HF seen from load terminals vs. frequency in ON and OFF states.

The third harmonic filter not only provides a low impedance path for the third harmonic currents of the TSI but also for the third harmonic current of the load.

5.3 Properties and design of adaptive harmonic blocking compensators

The equivalent circuit of the AHBC_B, as shown in Figure 5.3.1, is used to study its properties. The thyristor-switched inductor is equivalent to an inductor, denoted as $L_{eff}$, and a harmonic current source $j'$ in parallel.
Figure 5.3.1 Equivalent circuit of a single-phase AHBC_B.

For instance, $L_{\text{eff}} = L_c$ at $\alpha = 90^\circ$, which implies that the TSI branch is switched ON all of the time, and $L_{\text{eff}} = \infty$ at $\alpha = 180^\circ$, which implies that the TSI branch is switched OFF completely from the circuit. When $\alpha$ changes within the range of $90^\circ$ to $180^\circ$, the inductance $L_{\text{eff}}$ is controlled in the range of $L_c$ to $\infty$. The inductance of such an inductor, which is a function of the thyristor firing angle $\alpha$, is called the effective inductance of the TSI, and it is equal to

$$L_{\text{eff}} = \frac{L_c}{2 - \frac{2}{\pi} \alpha + \frac{1}{\pi} \sin 2\alpha}.$$

The frequency characteristics of an AHBC_B can be illustrated by using the plot of susceptance as seen from the load terminals versus frequency in two different states, i.e., at thyristors switched completely ON or OFF. Denoting

$$Z_{\text{eq}} = j n \omega L_{\text{eq}} + \frac{1}{j n \omega C_{\text{eq}}} = j \frac{n^2 - 1}{n \omega L_{\text{eq}}},$$

and

$$Z_{\text{scs}} = \frac{(j n \omega L_{\text{eff}}) (1/j n \omega C)}{j n \omega L_{\text{eff}} + 1/j n \omega C} = j \frac{n \omega L_{\text{eff}}}{1 - n^2 \omega^2 L_{\text{eff}} C},$$

where $L_{\text{eq}} = a^2 L_f$ and $C_{\text{eq}} = C_f/a^2$ if a transformer is present, the total admittance as seen from the load terminals is equal to:
\[ Y = \frac{1}{Z_{eq}} + \frac{1}{Z_{tcs}} = j \frac{n^2 \omega I^2 L_{eff} C - 1}{n \omega I L_{eff} (1 - n^2)} \]  

The admittance has pole at \( \omega = \omega_1 \), i.e., \( n = 1 \), and two zeros. With parameters of \( L_{eq} = 33.5 \text{mH}, C_{eq} = 210 \mu \text{F}, C = 70 \mu \text{F} \) and \( L_c = 13.75 \text{mH} \), the susceptance plot is shown in Figure 5.3.2. The location of the second pole is within the range between the minimum resonant frequency, \( \omega_{min} \), and the maximum resonant frequency, \( \omega_{max} \), which are equal to

\[ \omega_{min} = \sqrt{1 + \frac{C_{eq}}{C}}, \]  

and

\[ \omega_{max} = \frac{\left(\omega_1^2 L_c (C + C_{eq}) + 1\right) + \sqrt{\left(\omega_1^2 L_c (C + C_{eq}) + 1\right)^2 - 4 \omega_1^2 L_c C}}{2 \omega_1^2 L_c C}. \]  

Figure 5.3.2 Plot of susceptance of AHBC_B seen from the load terminals vs. frequency in ON and OFF states.

The values of \( \omega_{min} \) and \( \omega_{max} \) affect the compensator performance substantially. As shown in Figure 5.3.2, there is no harmonic frequency between \( \omega_{min} \) and \( \omega_{max} \). If
there is such a frequency, i.e., the second and/or the third harmonics, large voltage harmonics may occur at these frequencies while controlling the firing angle. Therefore, any harmonic frequency can be in the range between $\omega_{\text{min}}$ and $\omega_{\text{max}}$. To satisfy this requirement, the parameters of an AHBC_B have to be carefully chosen. Harmonic suppression increases with the frequency $\omega_{\text{min}}$ reduction. Therefore, it should be chosen as close to $\omega_f$ as possible.

The AHBC consists of two parts. The series part is a filter tuned to the fundamental frequency, and it is referred to as a “harmonic blocking filter” due to the nature of this filter’s behavior. The load is compensated and balanced by the shunt one-port with the thyristor controlled susceptance (TCS). In the case of an AHBC_B, the TCS consists of a capacitor and a thyristor-switched inductor (TSI). The TCS one-port parameters have to be calculated in such a way that the one-port provides the susceptance in the required control range between $T_a$ and $T_b$. By changing the firing angle, the susceptance provided by the TCS can be calculated and tabulated as shown in Table 5.3.1. The plot of susceptance versus the firing angle is shown in Figure 5.3.3.

<table>
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<th>$\alpha$</th>
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<th>$T$ (siemens)</th>
</tr>
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</tr>
<tr>
<td>$110^\circ$</td>
<td>24.99</td>
<td>-0.03141</td>
</tr>
<tr>
<td>$120^\circ$</td>
<td>35.17</td>
<td>0.00374</td>
</tr>
<tr>
<td>$130^\circ$</td>
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<td>238.44</td>
<td>0.06804</td>
</tr>
<tr>
<td>$160^\circ$</td>
<td>780.59</td>
<td>0.07577</td>
</tr>
<tr>
<td>$170^\circ$</td>
<td>6135.6</td>
<td>0.07874</td>
</tr>
<tr>
<td>$180^\circ$</td>
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<td>0.07917</td>
</tr>
</tbody>
</table>
The results from a computer simulation using PSpice for a single-phase AHBC_B were obtained to study its performance. To study the performance of a single-phase AHBC_B, it was simulated using PSpice along with the single-phase AC/DC converter, specified in Chapter 4. The waveforms of the supply voltage and the load current are shown in Figure 5.3.4. It was assumed that the AHBC_B had the following parameters $C_{eq} = 339\, \mu F$, $L_{eq} = 20.8\, mH$, $C = 505\, \mu F$ and $L_c = 13.5\, mH$. The waveforms of the supply voltage and current in the circuit with such an AHBC_B are shown in
Figure 5.3.5. The simulation results show that the supply voltage harmonic distortion is improved from 7.2% to 4.1% and the supply current harmonic distortion is improved from 43.2% to 7.7%. The waveforms and the harmonic spectra of the supply, TSI and load currents of the circuit with the AHBC_B were obtained and were shown in Figures 5.3.6 and 5.3.7, respectively. Observe, that apart from the third order harmonic of the load current, the TSI also generates a third order harmonic current and injects it into the
supply system. This is an essential drawback of the AHBC_B. However, an AHBC_B can still be used in three-phase balanced systems for reactive power compensation and harmonic reduction. In this case, three TCSs should be connected in a delta configuration. In such a case, the third harmonic currents generated by TSIs are in phase and canceled mutually in the delta structure. The waveforms of the supply voltage and the load current of a three-phase AC/DC converter are shown in Figure 5.3.8. The waveforms of the supply voltage and current of the three-phase balanced system with an AHBC_B obtained using PSpice are shown in Figure 5.3.9. It reduces the CHD from

![Figure 5.3.7 Harmonic spectra of the supply, TSI and load currents in circuit compensated with AHBC_B.](image)

![Figure 5.3.8 Waveforms of supply voltage and load current of 3-phase AC/DC converter before compensation.](image)
28.9% to 0.6% and the VHD from 5.1% to 4.2%. The current waveforms of the supply current, the load current and the TSI current are shown in Figure 5.3.10 and their harmonic spectra are shown in Figure 5.3.11. It can be seen in Figure 5.3.10 that the current harmonics generated by the TSI are relatively high, especially the third order harmonic, but the third order harmonic currents do not appear in the supply current. However, the AHBC_B should not be used for single-phase or three-phase unbalanced systems because an excessive amount of current harmonics, in particular the third

Figure 5.3.9 Supply voltage and current waveforms of circuit compensated with 3-phase AHBC_B.

Figure 5.3.10 Waveforms of the supply, TSI and load currents in circuit compensated with 3-phase AHBC_B.
harmonic, generated by TSIs may appear in the supply current. For such applications, an AHBC_3HF should be used. The equivalent circuit of an AHBC_3HF is shown in Figure 5.3.12.

![Figure 5.3.12 Equivalent circuit of a single-phase AHBC_3HF.](image)

To show how it improves the compensator performance, the admittance of the controlled one-port, $Y_{ecs}$, have to be calculated. It is equal to

$$Y_{ecs} = jn\omega C + \frac{1}{jn\omega L + 1/jn\omega C_3} + \frac{1}{jn\omega L_{eff}}$$

$$= j(n\omega C + \frac{9n\omega C_3}{9 - n^2} - \frac{1}{n\omega L_{eff}})$$

(122)
The total admittance seen from the load terminals is equal to

\[
Y = Y_{eq} + Y_{sc} = j \frac{an^6 + bn^4 + cn^2 + d}{n\omega_1L_{eff}(9 - n^2)(1 - n^2)},
\]

(123)

where

\[
a = \omega_1^2CL_{eff},
\]

(124)

\[
b = -(1 + \omega_1^2L_{eff}(10C + 9C_3 + C_{eq})),
\]

(125)

\[
c = 10 + 9\omega_1^2L_{eff}(C + C_3 + C_{eq}),
\]

(126)

and

\[
d = -9.
\]

(127)

According to eqn. (122), the admittance as seen from the load terminals has two poles at \(\omega = \omega_1\) and \(\omega = 3\omega_1\) and three zeros. For the compensator with parameters \(C_{eq} = 339\mu\) F, \(L_{eq} = 20.8\) mH, \(C = 505\mu\) F, \(C_3 = 35\mu\) F, \(L_3 = 22.3\) mH and \(L_c = 11.4\) mH, the plot of susceptance seen from the load terminals versus relative frequency, \(\omega/\omega_1\), is shown in Figure 5.3.13. The plot shows that the first zero is located between 0 and \(\omega_1\), the second zero can be located between \(\omega_1\) and \(3\omega_1\), and the third zero can be located anywhere.
above $3\omega_f$. The susceptance provided by the TCS at the fundamental frequency is equal to

$$T = \omega_f C + \frac{9\omega_f C_3}{8} \frac{1}{\omega_f L_{ef}}. \quad (128)$$

At the fundamental frequency, the capacitor provides a fixed capacitive susceptance, the third harmonic filter provides a fixed capacitive susceptance and the TSI provides a variable inductive susceptance which changes with the firing angle. The plot of the susceptance of the one-port versus the firing angle is shown in Figure 5.3.14.

![Plot of susceptance of the TCS one-port vs. firing angle.](image)

To study the performance of a single-phase AHBC_3HF, it was modeled with PSpice. The waveforms of the supply voltage and the load current before compensation are shown in Figure 5.3.15. The waveforms of the supply voltage and current in the circuit compensated with the AHBC_3HF are shown in Figure 5.3.16. The results obtained from PSpice show that the CHD was reduced from 28.9% to 2.7%. The waveforms and the harmonic spectra of the supply, TSI, load and the third harmonic filter currents in the circuit with the AHBC_3HF are shown in Figures 5.3.17 and 5.3.18, respectively.
Figure 5.3.15 Waveforms of supply voltage and load current of single-phase AC/DC converter before compensation.

Figure 5.3.16 Waveforms of the supply voltage and current in circuit compensated with AHBC_3HF.

Figure 5.3.17 Current waveforms of the supply, TSI, load and 3HF currents in circuit compensated with AHBC_3HF.
Figure 5.3.18 Harmonic spectra of the supply, TSI, load and 3HF currents in the circuit compensated with AHBC_3HF.

The supply current contains only a very low third order harmonic since the third harmonic filter of the TCS one-port absorbs most of the third harmonics generated by the load and the TSI.

An AHBC_3HF instead of an AHBC_B should be used for three-phase unbalanced systems. The waveforms of the load currents for a three-phase unbalanced system are shown in Figure 5.3.19, and the waveforms of the supply currents in the circuit compensated with an AHBC_3HF are shown in Figure 5.3.20. The results obtained from PSpice simulation show that the \( CHD \) was reduced from 16.1% to 2.9% and the supply currents were symmetrical with respect to their rms values and phase shifts.

Figure 5.3.19 Waveforms of load currents of 3-phase unbalanced system before compensation.
Figure 5.3.20 Waveforms of supply currents of 3-phase unbalanced system compensated with a 3-phase AHBC_3HF.

An AHBC_3HF is built of two parts, the HBC and the TCS one-port, as shown in Figure 5.3.21.

Parameters of the compensator depend on the required boundary of the control range, i.e., the minimum susceptance, $T_a$, and the maximum susceptance, $T_b$, and the locations of the second and the third maximum resonant frequencies, $\omega_{1\text{max}}$ and $\omega_{2\text{max}}$, respectively. Having $T_a$, $T_b$, $\omega_{1\text{max}}$ and $\omega_{2\text{max}}$, all the parameters $L_c$, $C_{eq}$, $L_{eq}$, $C_3$, $L_3$ and $C$ can be found, namely,

$$L_c = \frac{1}{\omega_j(T_b - T_a)},$$  \hspace{1cm} (129)
\[ C_{eq} = \frac{\frac{9D''E'' - N'' - D''}{8} T_b - K''E''}{M'' - \frac{9D''E''}{8} + K''E''} \]  
(130)

\[ L_{eq} = \frac{1}{\omega_i^2 C_{eq}} \]  
(131)

\[ C_3 = E''E'' C_{eq}, \]  
(132)

\[ L_3 = \frac{1}{9\omega_i^2 C_3}, \]  
(133)

and

\[ C = \frac{1}{\omega_i} T_b - \frac{9}{8} C_3, \]  
(134)

where

\[ D'' = \omega_{2\text{max}}^2 \omega_i^2 L_c (\omega_{2\text{max}}^4 - 10\omega_{2\text{max}}^2 + 9), \]  
(135)

\[ D' = \omega_{1\text{max}}^2 \omega_i^2 L_c (\omega_{1\text{max}}^4 - 10\omega_{1\text{max}}^2 + 9), \]  
(136)

\[ K'' = 9\omega_{2\text{max}}^2 \omega_i^2 L_c (1 - \omega_{2\text{max}}^2), \]  
(137)

\[ K' = 9\omega_{1\text{max}}^2 \omega_i^2 L_c (1 - \omega_{1\text{max}}^2), \]  
(138)

\[ M'' = \omega_{2\text{max}}^2 \omega_i^2 L_c (9 - \omega_{2\text{max}}^2), \]  
(139)

\[ M' = \omega_{1\text{max}}^2 \omega_i^2 L_c (9 - \omega_{1\text{max}}^2), \]  
(140)

\[ N'' = -(\omega_{2\text{max}}^4 - 10\omega_{2\text{max}}^2 + 9), \]  
(141)

\[ N' = -(\omega_{1\text{max}}^4 - 10\omega_{1\text{max}}^2 + 9), \]  
(142)

\[ E'' = -\frac{M' C_{eq}}{9K' - \frac{9}{8} D'}, \]  
(143)

and

\[ E' = -\frac{N' + \frac{D'}{\omega_i} T_b}{9K' - \frac{9}{8} D'}. \]  
(144)
It should be noted that positive compensator parameters may not exist for any given design specifications.

To have a better understanding of how the performance is related to certain input data, one of these data may be varied in order to obtain a particular set of compensator parameters and a computer simulation may be run. The single-phase AC/DC convert specified in Chapter 4 was the load which needed to be compensated. The values of the $CHD, \delta_i$, along with compensator parameters were tabulated in Tables 5.3.2 and 5.3.3. For Table 5.3.2, the following input data were used: $T_a = -0.077S$, $T_b = 0.25S$, $\omega_{l_{\text{max}}} = 1.5 \text{ rad/sec}$. The frequency $\omega_{2_{\text{max}}}$ was varied. For Table 5.3.3, the following input data were used: $T_a = -0.077S$, $T_b = 0.25S$, $\omega_{2_{\text{max}}} = 3.5 \text{ rad/sec}$ and frequency $\omega_{l_{\text{max}}}$ was varied.

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<th>$C_j$ ($\mu$F)</th>
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The dependence of the $CHD$, $\delta_i$, on the chosen frequencies $\omega_{1\text{max}}$ and $\omega_{2\text{max}}$ are shown in Figures 5.3.22 and 5.3.23.

![Figure 5.3.22 Plot of dependence between CHDs and frequencies $\omega_{1\text{max}}$ and $\omega_{2\text{max}}$.](image)

![Figure 5.3.23 Plot of dependence between CHDs and frequencies $\omega_{1\text{max}}$ and $\omega_{2\text{max}}$.](image)

Conclusions of this dependence from computer simulations:

1) Figure 5.3.22 shows that to improve harmonic suppression, the frequency $\omega_{2\text{max}}$ should be located as close to the third harmonic frequency as possible.

2) Figure 5.3.23 shows that to improve harmonic suppression, the frequency $\omega_{1\text{max}}$ should be located as close to the fundamental frequency as possible.
To estimate the cost of reactive elements of the compensator, it was assumed, according to various catalog prices, that the price of an inductor was 20 unit dollars ($) per mH and the price of a capacitor was 1 unit dollar per μF. The compensator was designed, for various values of frequencies \( \omega_{l_{\text{max}}} \) and \( \omega_{z_{\text{max}}} \), to compensate the single-phase AC/DC converter specified in Chapter 4 to have the \( CHD \) below 3\%. The results of respective frequencies \( \omega_{l_{\text{max}}} \) and \( \omega_{z_{\text{max}}} \), the total required capacitance, the total required inductance, the \( CHD \), \( \delta_i \), and the cost are tabulated in Table 5.3.4. It is obvious that the first three options in Table 5.3.4 will cost much more as compared with others although they provide a very impressive harmonic suppression, \( \delta_i = 0.2\% \). The fourth option with \( \delta_i = 0.7\% \) seems to be a reasonable solution for the given specification. However, the least expensive one is the tenth option with \( \omega_{l_{\text{max}}} = 1.5\text{rad/sec} \) and \( \omega_{z_{\text{max}}} = 3.7\text{rad/sec} \). Of course, the total cost should include the costs for thyristors, transducers

<table>
<thead>
<tr>
<th>( \omega_{l_{\text{max}}} )</th>
<th>( \omega_{z_{\text{max}}} )</th>
<th>( C_{\text{total}} ) (μF)</th>
<th>( L_{\text{total}} ) (mH)</th>
<th>( \delta_i ) (%)</th>
<th>Cost ($)</th>
</tr>
</thead>
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<tr>
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<td>680.4</td>
<td>253.85</td>
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<tr>
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<td>671.0</td>
<td>229.29</td>
<td>0.2</td>
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</tr>
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<td>4.5</td>
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<td>66.06</td>
<td>2.4</td>
<td>2070</td>
</tr>
<tr>
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<td>5.5</td>
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<td>62.84</td>
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<tr>
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</tr>
<tr>
<td>1.5</td>
<td>3.3</td>
<td>1022.7</td>
<td>35.5</td>
<td>2.0</td>
<td>1732</td>
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<td>31.87</td>
<td>2.3</td>
<td>1668</td>
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<td>3.7</td>
<td>1035.5</td>
<td>30.08</td>
<td>2.7</td>
<td>1635</td>
</tr>
</tbody>
</table>

Table 5.3.4 Cost table.

and all associated circuits. It was assumed that these costs were the same for each AHBC_3HF, therefore, these costs were not considered while calculating the costs in Table 5.3.4.
5.4 Dynamic properties of adaptive harmonic blocking compensators

Adaptation can be considered to be a process of modifying the parameters of the system to reach some goals. The word 'adaptive' in the name of adaptive harmonic blocking compensators implies that TCSs of the ABC associated with an AHBC adjust their susceptances at any change of the load parameters to compensate the reactive current and the unbalanced current at the fundamental frequency. The simplest control algorithm is an open-loop control. The flow diagram of an open-loop control of an AHBC is shown in Figure 5.4.1. The susceptance needed for each phase is calculated by eqn. (38), (39) and (40), respectively. Then, the corresponding firing angles are found and the thyristors are triggered. In this case, the measurements of two currents and two voltages can be done at the cross-section in front of load terminals. However, the accuracy with such an open-loop control could be low, since the results of compensation are not used in the control algorithm.

\[ Y_{sr}, Y_{st}, B, \text{ and } A \]

Figure 5.4.1 Flowchart of an open-loop control of AHBCs.

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Usually, the primary objective of the control system is to obtain a high steady-state accuracy. For this purpose, a feedback control has to be used. The steady-state accuracy of many feedback systems can be increased by increasing the amplifier gain in the forward path. However, the resulting transient response may not be acceptable. The system may not even be stable. For the feedback system, the measurements are taken at the cross-section between the harmonic blocking filters and the ABC as shown in Figure 5.4.2. For AHBC experiments, a proportional control with a gain of $K$ was the first approach. However, this approach failed because if the proportional gain $K$ was too large, the system became unstable. If $K$ was too small, the steady-state accuracy was not sufficient. Therefore, such a feedback control with a proportional gain was abandoned.

![Control system of AHBC](image)

**Figure 5.4.2** Diagram of control system associated with AHBC.
If the steady-state accuracy is not sufficient, one can introduce an integration-type network to compensate for the lack of integration in the original transfer function. One form of controller widely used is the proportional plus integral (PI) controller, which has a transfer function

\[ G_c(s) = K_p + \frac{K_i}{s} = \frac{K_p}{s} (s + \frac{K_i}{K_p}), \]

where \( K_p \) is the proportional constant and \( K_i \) is the integral constant. In this case the PI controller is equivalent to adding a zero at \( s = -K_i/K_p \) and a pole at \( s = 0 \) to the open-loop transfer function. One obvious effect of the integral control is that it increases the order of the system by one. More importantly, it increases the type of the system by one. Therefore, the steady-state error of the original system without integral control is improved by one order; that is, if the steady-state error to a given input is constant, the integral control reduces it to zero and ensures the stability of the final system. However, because the system is now one order higher, it may be less stable than the original system or even become unstable if the parameters \( K_p \) and \( K_i \) are not properly chosen.

For AHBC application, a PI control algorithm is used. As mentioned in Chapter 2, all the calculations are based on the measurements of two load currents, \( i_R \) and \( i_S \), and two line to line voltages, \( u_{RT} \) and \( u_{ST} \). It should be mentioned that in three-phase three-wire systems the third current \( i_T \) is dependent on the others, e.g., \( i_T = -(i_R + i_S) \). The samples of these four signals are used to calculate their complex rms (CRMS) values, \( I_R, I_S, U_{RT}, \) and \( U_{ST} \), respectively. The concept of the fictitious admittances of the load [21] is used, which implies that one of the load admittances can be chosen arbitrarily. In this dissertation, the load admittance \( Y_{RS} = 0 \) is assumed. To calculate the needed susceptances, the equivalent admittance \( Y_{el} \) and the unbalanced admittance \( A_l \) of the load at the fundamental frequency are needed and are simplified, namely,

\[ Y_{el} = Y_{STI} + Y_{RTI}, \]
and

\[ A_i = -(Y_{STI} + \alpha Y_{RTI}), \]  

(147)

where

\[ Y_{RTI} = \frac{I_{RI}}{U_{RTI}}, \]  

(148)

and

\[ Y_{STI} = \frac{I_{SI}}{U_{STI}}. \]  

(149)

If the reactive and unbalanced components of the load current fundamental harmonic are compensated, then \( B_{ei} = 0 \) and \( A_i = 0 \). If not, the balancing compensator susceptances \( T_{RS}, T_{ST} \) and \( T_{TR} \) should be changed by the values of

\[ \Delta T_{RS} = (\sqrt{3} Re A_i - Im A_i - B_{ei}) / 3, \]  

(150)

\[ \Delta T_{ST} = (2 Im A_i - B_{ei}) / 3, \]  

(151)

\[ \Delta T_{TR} = - (\sqrt{3} Re A_i + Im A_i + B_{ei}) / 3. \]  

(152)

Then, the needed susceptances can be calculated in discrete time as follows:

\[ T_{RSk} = IT_{RSk} + K_p \Delta T_{RSk}, \]  

(153)

\[ T_{STk} = IT_{STk} + K_p \Delta T_{STk}, \]  

(154)

and

\[ T_{TRk} = IT_{TRk} + K_p \Delta T_{TRk}, \]  

(155)

where

\[ IT_{RSk} = IT_{RSk-1} + K_i \Delta T_{RSk}, \]  

(156)

\[ IT_{STk} = IT_{STk-1} + K_i \Delta T_{STk}, \]  

(157)

and

\[ IT_{TRk} = IT_{TRk-1} + K_i \Delta T_{TRk}. \]  

(158)

The thyristor firing angles can be determined according to these susceptances \( T_{RS}, T_{ST} \) and \( T_{TR} \). The block diagram of the AHBC control system is represented in Figure 5.4.3.
It should be mentioned that if the needed susceptance is over the range of the minimum susceptance $T_a$ to the maximum susceptance $T_b$, then the thyristor is triggered with a firing angle of $90^\circ$ or $180^\circ$, respectively.
CHAPTER 6

EXPERIMENTAL RESULTS

6.1 Introduction

Theoretical analysis and computer simulation using PSpice are the main tools in previous chapters for studying harmonic blocking compensators and adaptive harmonic blocking compensators. However, theoretical analysis and computer simulation may not provide full information on devices under development. Laboratory experiments are needed for that. Experimental results provide more reliable verification of the performance of the harmonic blocking compensators and adaptive harmonic blocking compensators than computer modeling.

This chapter presents some experimental results obtained from measurements on laboratory prototypes of a HBC and an AHBC. However, the built prototypes were relatively low power devices with low inductors' quality factor $q$, which usually increases with the increase of an inductor's power rating. Consequently, the effectiveness of the devices constructed in the laboratory was lower than that which could be obtained with high power compensators. In spite of this, the experimental results confirm the effectiveness of HBCs and AHBCs.

This chapter presents experimental results obtained with a HBC prototype in single-phase and three-phase systems. In the case of the AHBC, only the results obtained in a three-phase system, where adaptive balancing could be a matter of the concern, are presented.

The following experiments are described in the next two sections:

Experiment 1: A single-phase system with a single-phase HBC.

Experiment 2: A three-phase system with a three-phase HBC.

Experiment 3: A three-phase system with an AHBC under steady-state conditions.

Experiment 4: A three-phase system with an AHBC under transient conditions.

The experiments are specified as follows:

88
1) The goal of the experiment.
2) Load specification.
3) Compensator specification.
4) The description of the digital signal processing system (DSPS) and the algorithm.
5) The results.
6) Discussion of the results obtained.

6.2 Experiments with harmonic blocking compensators

The HBC performance was studied in the two following experiments.

Experiment 1. A single-phase system with a HBC.

1) Goal. Verification of the effectiveness of a single-phase HBC.

2) Load. A controlled single-phase AC/DC converter with a resistive and inductive load, as shown in Figure 6.2.1. The load current and the supply voltage waveforms are shown in Figure 6.2.2 and their harmonic spectra are shown in Figures 6.2.3 and 6.2.4, respectively.

Figure 6.2.1 A single-phase system with a controlled AC/DC converter.
Figure 6.2.2 Supply voltage and load current waveforms.

Figure 6.2.3 Harmonic spectrum of the load current.

Figure 6.2.4 Harmonic spectrum of the supply voltage in the circuit without compensator.

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3) **Compensator.** A single-phase HBC with $L_f = 33.5\, \text{mH}$, $C_f = 210\, \text{\mu F}$ and $C = 180\, \text{\mu F}$, as shown in Figure 6.2.5, was used for reactive power compensation and harmonic suppression.

![Figure 6.2.5 Circuit configuration with HBC.](image)

4) **DSPS.** The DSPS used in this experiment consists of two transducers, one for voltage and one for current, an analog-to-digital converter (Appendix A) installed in a computer and a signal analysis program written in C++. The transducers provide analog signals for the A/D converter. The signal analysis program performs a Fast Fourier Transform (FFT) to provide harmonic spectra of the voltage and the current and then calculates their rms values, active power, apparent power, power factor, $CHD$ and $VHD$.

5) **Results.** The results of compensation are shown in Figures 6.2.6, 6.2.7 and 6.2.8, where Figure 6.2.6 presents the supply current and the supply voltage waveforms and Figures 6.2.7 and 6.2.8 show their harmonic spectra, respectively.
Figure 6.2.6 Supply voltage and current waveforms of the circuit with a HBC.

Figure 6.2.7 Harmonic spectrum of the supply current of the circuit with a HBC.

Figure 6.2.8 Harmonic spectrum of the supply voltage of the circuit with a HBC.
Discussion. The $VHD$ before compensation was of the order of $\delta_u = 6.1\%$, the $CHD$ was of the order of $\delta_i = 40.9\%$ and the power factor was $\lambda = 0.64$. The HBC reduced the $CHD$ to 7.6\% and the $VHD$ to 2.7\% and the power factor was increased to 0.99.

Experiment 2. A three-phase system with a three-phase HBC.

1) Goal. Verification of the effectiveness of a three-phase HBC.

2) Load. A controlled three-phase AC/DC converter with resistive and inductive load, as shown in Figure 6.2.9. The load current and the supply voltage waveforms are shown in Figure 6.2.10 and their harmonic spectra are shown in Figures 6.2.11 and 6.2.12, respectively.

![Figure 6.2.9 A three-phase controlled AC/DC converter.](image)

![Figure 6.2.10 Supply voltage and load current waveforms of the AC/DC converter.](image)
3) **Compensator.** A three-phase HBC with transformer with $L_f = 33.5\, \text{mH}$, $C_f = 210\, \mu\text{F}$, $C = 35\, \mu\text{F}$ and transformer turn ratio $a = 1.1$ in each phase, as shown in Figure 6.2.13, was used for reactive power compensation and harmonic suppression.

4) **DSPS.** The same DSPS was used except the active power and the apparent power were calculated for three-phase quantities. Since the three-phase system was balanced, only one current and one voltage were recorded.
5) **Results.** The supply current and the supply voltage waveforms of the circuit compensated with the three-phase HBC are shown in Figure 6.2.14 and their harmonic spectra are shown in Figures 6.2.15 and 6.2.16, respectively.
6) **Discussion.** The *VHD* before compensation was of the order of $\delta_u = 9.1\%$, the *CHD* was of the order of $\delta_i = 30.9\%$ and the power factor was $\lambda = 0.65$. The HBC reduced the *CHD* to 2.7%, the *VHD* to 2.7% and the power factor was increased to 0.99. The third order harmonic currents do not exist in three-phase balanced systems. Because of this, such a three-phase HBC has a higher effectiveness than the single-phase HBC. Experiments with the same parameters except the transformer turn ratio was $a = 0.9$ and 1.0 have been done as well, to demonstrate how the turn ratio $a$ affects the performance of HBCs. The *CHDs*
and the voltages across the capacitor of the harmonic blocking filters were measured and calculated for different values of the turn ratio. The results are complied in Table 6.2.1.

Table 6.2.1 Table of \( CHD, \delta_i \), and voltage \( U_{cf} \) with different turn ratio \( a \).

<table>
<thead>
<tr>
<th>( a )</th>
<th>( \delta_i ) (%)</th>
<th>( U_{cf} ) (V)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.9</td>
<td>4.7</td>
<td>56.4</td>
</tr>
<tr>
<td>1.0</td>
<td>3.1</td>
<td>63.2</td>
</tr>
<tr>
<td>1.1</td>
<td>2.5</td>
<td>70.4</td>
</tr>
</tbody>
</table>

Observe, that the impedance of each series branch of the HBC at harmonic frequencies increases with the turn ratio \( a \) increase. Consequently, the \( CHD, \delta_i \), decreases. It implies that the turn ratio can be manipulated to achieve a higher harmonic attenuation if the other parameters are fixed. However, the voltage across the capacitor of the filter increases with the transformer turn ratio \( a \) increase.

The experiments with HBCs show that harmonic blocking compensators provide a remarkable suppression of the load generated harmonics. HBCs improve the power factor by compensating the reactive power and reducing harmonics independently of its spectrum. The \( CHD \) can be reduced to a desired level and the \( VHD \) can be reduced to its original level in the distribution system.

6.3 Experiments with adaptive harmonic blocking compensators

The AHBC performance was studied in the two following experiments.

Experiment 3. A three-phase system with an AHBC under steady-state conditions.

1) Goal. Verification of the effectiveness of an AHBC under steady-state condition.
2) **Load.** A three-phase controlled AC/DC converter with resistive and inductive load and an inductor of inductance of 120mH connected in between phases R and T, as shown in Figure 6.3.1.

![Figure 6.3.1 Unbalanced harmonic generating load used in Experiment 3.](image)

3) **Compensator.** A three-phase AHBC as shown in Figure 6.3.2, was used for reactive power and unbalanced power compensation and harmonic suppression.

![Figure 6.3.2 Circuit configuration of system with an AHBC.](image)
This laboratory AHBC prototype consists of three main components:

a) Harmonic blocking filters - including three harmonic filters tuned to the fundamental frequency with $L_f = 33.5\,\text{mH}$ and $C_f = 210\,\mu\text{F}$.

b) Adaptive balancing compensator - including three thyristor controlled susceptance circuits (TCSs) in delta configuration, as shown in Figure 6.3.3, with $C = 70\,\mu\text{F}$, $L_3 = 3.72\,\text{mH}$, $C_3 = 210\,\mu\text{F}$ and $L = 13.75\,\text{mH}$.

![Figure 6.3.3 Structure of three TCSs in delta configuration.](image)

Figure 6.3.3 Structure of three TCSs in delta configuration.

c) Digital signal processing system of AHBC - This DSPS consists of signal processing hardware and a control algorithm. The hardware includes transducers for three-phase voltages and currents, a microcontroller (8XC196MC by Intel) and some associated circuits for digital signal processing (Appendix B). The transducers provide analog signals for microcontroller evaluation board. The control algorithm provides the thyristor firing angles calculated based on the current orthogonal decomposition method as discussed in Chapter 5. A firing pulse generator generates firing pulses for the thyristor drivers of TCSs.

4) DSPS. The DSPS used in this experiment consists of four current transducers, to monitor the load currents and the supply currents, an analog-to-digital converter...
and an instrumentation software, LabWindows/CVI (Appendix C). Then the data generated by LabWindows/CVI is saved into the format for MATLAB. The file with .mat extension is used as the input file by a MATLAB m-file which generates the plots of the load currents and the supply currents and performs a Fast Fourier Transform (FFT) to obtain current harmonic spectra. Since this experiment is for steady-state, only waveforms of two cycles are shown.

5) **Results.** The waveforms of the load currents and the supply currents in the circuit with the AHBC under steady-state condition are shown in Figure 6.3.4. The harmonic spectra and rms values of the load currents and supply currents are tabulated in Tables 6.3.1 and 6.3.2, respectively.

![Figure 6.3.4 Load currents and supply currents in Experiment 3.](image-url)
# Table 6.3.1 Load current harmonic spectrum and rms values.

<table>
<thead>
<tr>
<th></th>
<th>Phase R</th>
<th></th>
<th>Phase S</th>
<th></th>
<th>Phase T</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>rms (A)</td>
<td>%</td>
<td>rms (A)</td>
<td>%</td>
<td>rms (A)</td>
<td>%</td>
</tr>
<tr>
<td>$I_1$</td>
<td>8.49</td>
<td>100.0</td>
<td>5.73</td>
<td>100.0</td>
<td>10.28</td>
<td>100.0</td>
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<td>0.06</td>
<td>0.7</td>
<td>0.02</td>
<td>0.4</td>
<td>0.07</td>
<td>0.7</td>
</tr>
<tr>
<td>$I_3$</td>
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<td>0.9</td>
<td>0.02</td>
<td>0.3</td>
<td>0.09</td>
<td>0.9</td>
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<tr>
<td>$I_4$</td>
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<td>0.1</td>
<td>0.03</td>
<td>0.5</td>
<td>0.04</td>
<td>0.4</td>
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<tr>
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<td>1.21</td>
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<td>$I_7$</td>
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<td>$I_{13}$</td>
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<td>0.43</td>
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<td>0.26</td>
<td>2.5</td>
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<td></td>
<td>5.98</td>
<td></td>
<td>10.43</td>
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# Table 6.3.2 Supply current harmonic spectrum and rms values.

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<th>Phase S</th>
<th></th>
<th>Phase T</th>
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<td>rms (A)</td>
<td>%</td>
<td>rms (A)</td>
<td>%</td>
<td>rms (A)</td>
<td>%</td>
</tr>
<tr>
<td>$I_1$</td>
<td>5.07</td>
<td>100.0</td>
<td>5.05</td>
<td>100.0</td>
<td>5.26</td>
<td>100.0</td>
</tr>
<tr>
<td>$I_2$</td>
<td>0.03</td>
<td>0.5</td>
<td>0.03</td>
<td>0.7</td>
<td>0.02</td>
<td>0.3</td>
</tr>
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<td>0.10</td>
<td>0.2</td>
<td>0.01</td>
<td>0.2</td>
<td>0.02</td>
<td>0.4</td>
</tr>
<tr>
<td>$I_4$</td>
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<td>0.2</td>
<td>0.01</td>
<td>0.2</td>
</tr>
<tr>
<td>$I_5$</td>
<td>0.17</td>
<td>3.3</td>
<td>0.18</td>
<td>3.6</td>
<td>0.17</td>
<td>3.2</td>
</tr>
<tr>
<td>$I_7$</td>
<td>0.17</td>
<td>3.4</td>
<td>0.22</td>
<td>4.4</td>
<td>0.13</td>
<td>2.5</td>
</tr>
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<td>0.0</td>
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<td>0.01</td>
<td>0.1</td>
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<td>0.01</td>
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<td>0.2</td>
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<td>0.01</td>
<td>0.1</td>
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6) **Discussion.** The supply current rms value $||i||$ was 8.89A while the load current rms was 14.81A calculated by eqn. (5). The CHD of the supply current was 4.9% while the load current CHD was 20.4% calculated by eqn. (51).

**Experiment 4.** A three-phase system with an AHBC under transient conditions.

1) **Goal.** Observation of the AHBC performance in a transient state.

2) **Load.** A resistor was switched on, as shown in Figure 6.3.5, in the same circuit as described in Experiment 3 to initiate transients in the system with AHBC.

3) **Compensator.** The same AHBC used in Experiment 3 was used.

4) **DSPS.** The same DSPS as described in Experiment 3 was used except the MATLAB m-file generates the plots of the load currents and the supply currents and the plot of harmonic rms values of the supply currents for eight cycles to show the transient.

5) **Results.** The load currents and the supply currents of the AHBC are shown in Figure 6.3.6. To see the transient response, waveforms of eight periods are recorded. The rms values of the supply current harmonics in those periods were shown in Figure 6.3.7.
6) **Discussion.** After the resistive load was connected between phases R and T, the supply current was asymmetrical. After approximately four cycles delay, the supply current had been symmetrized. Also, the CHD of the supply currents was reduced to the order of 4%.

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The experiments show that adaptive harmonic blocking compensators provide remarkable performance in suppressing harmonics and adaptive compensation of reactive and unbalanced power. The control algorithm of the AHBC control system introduced in Chapter 5 is verified. It provides the thyristors' firing angles needed for reactive and unbalanced power compensation at the fundamental frequency. The transient response with a settling time of four cycles and a relatively small overshoot are acceptable for a laboratory prototype under development. Therefore, there is no further study concerning the possible effects of the transient response of the AHBC's operation.
CHAPTER 7
CONCLUSIONS

Harmonic blocking compensators (HBCs) are very effective devices for suppression of harmonic distortion caused by non-linear loads and/or time variant loads, and HBCs do not cause resonances with the distribution system inductance. Since harmonic blocking compensators do not have shunt branches tuned to harmonic frequencies, such devices are particularly suitable for distribution systems with loads that generate a large number of current harmonics and even current components of non-harmonic frequencies, referred to sometimes as inter-harmonics, where resonant harmonic filters cannot be effectively used. However, these advantages are obtained at the cost of higher power ratings of the compensator over traditional resonant harmonic filters (RHFs), a voltage drop on the series branch and extra loss of active power on the series filter inductor. As shown in Chapter 4, the power ratings, the loss of active power and the voltage drop decline if the HBC is applied for compensating loads with a low power factor for the fundamental frequency, $\lambda_f$. Moreover, the relative loss of active power and the voltage drop can be reduced by increasing the inductor's quality factor, $q$, which can be elevated along with an increase in the inductor's power rating. Therefore, it seems that a HBC could be a reasonable option if compensation of high power loads with low power factor and rich current harmonic spectrum is a matter of concern.

Harmonic blocking compensators have other advantages over traditional RHFs. A HBC can be easily converted into an adaptive compensator, which is practically not possible in the case of resonant harmonic filters. This is because the value of the capacitance $C$ can be adjusted to the variability of the load reactive power without any substantial effect on the harmonic attenuation. Moreover, distortion of the load current may change in a broad range, and only in some short intervals of time this distortion is unacceptably high, so that, there is no need for a permanent harmonic reduction. In such a case, the series resonant filter could be by-passed by a switch connected in parallel as
shown in Figure 4.3.9. This could reduce the active power loss on its inductor. Also, systems with existing capacitor banks could be equipped with the series resonant filter and converted into a harmonic blocking compensator.

The adaptive harmonic blocking compensator (AHBC) integrates a harmonic blocking compensator with an adaptive balancing compensator (ABC). Such a compensator can be used for compensating reactive power, for symmetrizing load currents and for protecting the system against the load generated current harmonics. It reduces not only current harmonic distortion but also the voltage harmonic distortion caused by harmonic generating loads.

The ABC is built of three thyristor controlled susceptance (TCS) one-ports in a delta configuration. The structure of the TCS has to be chosen in such a way that current harmonics generated by the TCS's thyristor-switched inductor (TSI) will not inject into the system. A TCS structure consisting of a TSI, a capacitor and a harmonic filter tuned to the third harmonic frequency in parallel is the recommended structure to be used with an AHBC.

The choice of parameters of an AHBC is a matter of the trade-off between the cost of components and the effectiveness of harmonic suppression. Parameters of the compensator depend on the required boundary of the control range, i.e., the minimum susceptance, $T_a$, and the maximum susceptance, $T_b$, and the locations of the second and the third maximum resonant frequencies, $\omega_{1\text{max}}$ and $\omega_{2\text{max}}$, respectively. To improve the harmonic suppression of an AHBC, the second resonant frequency $\omega_{1\text{max}}$, which has a range between $\omega_f$ and $3\omega_f$, should be located as close to the fundamental frequency as possible. Also, the third resonant frequency $\omega_{2\text{max}}$ should be located as close to the third order harmonic frequency as possible.
REFERENCES


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APPENDIX A

DSPS ASSOCIATED WITH THE HBC EXPERIMENTS

One voltage transducer and one current transducer (OHIO SEMITRONICS, INC.) were used for the DSPS associated with the HBC experiments. The connection diagram for both transducers is shown in Figure A.1. Both transducers require a 85-135 VAC instrument power supply and have an accuracy of the order +/- 0.5% full scale.

![Connection diagram of voltage and current transducers.](image)

The output signals are then connected to an analog-to-digital conversion board (RTI-860, ANALOG DEVICES). The RTI-860 is a high-speed data acquisition board that plugs into one of the long expansion slots in an IBM PC. The RTI-860 provides 16 single-ended analog input channels that measure analog signals in the range of ±10 V or ±5 V. The analog-digital converter has 12-bit resolution (2 maximum of 4096 counts), which provides a least significant bit (LSB) value of 4.88 mV in the ±10 V range and 2.44 mV in the ±5 V range. The analog-to-digital conversion time is 3μs (typical).

A signal analysis program written in C++ was used for the HBC experiments. The signal analysis program performs a Fast Fourier Transform (FFT) to provide harmonic spectra of the voltage and the current and then calculates their rms values, active power, apparent power, power factor, \( CHD \) and \( VHD \).
APPENDIX B

DSPS OF THE AHBC

A digital signal processing system (DSPS) for three-phase quantities has been developed for the control of the AHBC. The DSPS associated with the AHBC includes two current and two voltage transformers, instrumentation amplifier circuit, 8XC196MC microcontroller and evaluation board, filter and zero detector circuit, firing pulse generator and a control algorithm. The circuit configuration of such a DSPS is shown in Figure B.1. The 8XC196MC is a 16-bit microcontroller, designed especially for the high-speed event control required for inverter and electric motor applications. The 8XC196MC has a 64 Kbyte address space. Its main components include a CPU,
several types of memory, seven I/O ports and several on-chip peripheral devices. The peripherals include an A/D converter, an event processor array (EPA), two timers (Timer1 and Timer2), a three-phase waveform generator (WG) and a pulse width modulation unit (PWM). The transformers reduce the input voltage and current signals and the instrumentation amplifiers, as shown in Figure B.2, provide analog signals for the microcontroller evaluation board. The filter and zero detector circuit, as shown in Figure B.3, is needed for signal synchronization. The control algorithm provides the thyristor firing angles calculated based on the current orthogonal decomposition method as discussed in Chapter 5. A firing pulse generator, as shown in Figure B.4, generates firing pulses for the thyristors' drivers, as shown in Figure B.5, of TCSs. The flowchart of the control algorithm is shown in Figure B.6.
Figure B.2 Circuit configuration of instrumentation amplifiers.
Figure B.3 Circuit configuration of the filter and zero detector circuit.
Figure B.5 Circuit configuration of the TCS's thyristor driver circuit.
Figure B.5 Circuit configuration of the TCS's thyristor driver circuit.
Main routine

Initialize
port2, timer1, timer2,
A/D converter and
enable interrupt

Use CRMS values to
calculate $Y_{STI}$, $Y_{ST}$, $B_{cl}$
and $A_i$

Get errors of balancing
compensator
susceptances
$\Delta T_{RS}, \Delta T_{ST}, \Delta T_{TR}$

$IT_{RS} = IT_{RS,i} + K_p \Delta T_{RS}$
$T_{RS} = IT_{RS} + K_p \Delta T_{RS}$
$IT_{ST} = IT_{ST,i} + K_p \Delta T_{ST}$
$T_{ST} = IT_{ST} + K_p \Delta T_{ST}$
$IT_{TR} = IT_{TR,i} + K_p \Delta T_{TR}$
$T_{TR} = IT_{TR} + K_p \Delta T_{TR}$

Find corresponding
thyristor firing angles

Figure B.6 Main routine of the control algorithm.
APPENDIX C

DSPS ASSOCIATED WITH THE AHBC EXPERIMENTS

Four current transducer were used for the DSPS associated with the AHBC experiments. The output signals were then connected to an analog-to-digital converter (EISA-A2000, NATIONAL INSTRUMENTS). The EISA-A2000 has a 12-bit resolution, which is equivalent to an analog voltage resolution of 2.44 mV, A/D plug-in board for the EISA bus with a 1MS/s sampling rate. The board has four analog input channels, each with its own T/H circuitry. This board can sample one channel at 1MS/s, two channels simultaneously at 500KS/s or four channels simultaneously at 250KS/s. The input range for all channels is ±5 V.

The instrumentation software, LabWindows/CVI, is an integrated programming environment for building instrumentation systems for test and measurement, data acquisition and process monitoring and control applications using the ANSI C programming language. Data generated by LabWindows/CVI was saved into the format for MATLAB. The files with .mat extension were used as the input file by a MATLAB m-file which generates the plots of the load currents and the supply currents and performs a Fast Fourier Transform (FFT) to obtain current harmonic spectra.
VITA

Shih-Min Hsu was born on June 29, 1963, in Taipei, Taiwan, Republic of China. With an interest in electrical engineering, he chose Electrical Engineering as his major at Kuang Wu Junior College of Technology and National Taiwan Institute of Technology, Taipei, Taiwan. He received his bachelor of science in electrical engineering degree from National Taiwan Institute of Technology in June 1989.

He immigrated to the United States in July 1989, and entered the Department of Electrical and Computer Engineering of Louisiana State University, Baton Rouge, Louisiana, in August 1990. He earned his master of science in electrical engineering from Louisiana State University in December 1992. He continued his doctoral studies in the Department of Electrical and Computer Engineering of Louisiana State University. He became a citizen of the United States by naturalization in May 1995. He is going to receive his doctor of philosophy degree in December 1996.
DOCTORAL EXAMINATION AND DISSERTATION REPORT

Candidate: Shih-Min Hsu

Major Field: Electrical Engineering

Title of Dissertation: Adaptive Harmonic Blocking Compensators

Date of Examination: October 22, 1996

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