A New Adaptive Detection Algorithm for Power Quality Improvement

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A NEW ADAPTIVE DETECTION ALGORITHM FOR POWER QUALITY IMPROVEMENT

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To my Mom Exian Qian, Dad Zhongbao Qian and my brother Leyi Qian for their strong support from the other side of the earth.

This dissertation would be the best gift I have ever had for them.
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ABSTRACT

In this dissertation, a new adaptive harmonic selective detection algorithm is proposed for power quality improvement applications. The adaptive gains of the proposed method can be chosen relatively large to obtain faster convergence. The stability of the proposed method is guaranteed. The adaptive harmonic selective algorithm is analyzed then compared to a popular d-q method. This proposed adaptive method is simple and effective in extracting fundamental and harmonic current information from harmonic load currents. The extracted fundamental or harmonic currents therefore can be used as the reference signals for power quality improvement applications such as harmonic selective cancellation or reactive power compensation. The proposed adaptive algorithm can estimate time varying power system frequency and it also can identify a dc offset in a load current. These are two conditions that the d-q transformation based detection algorithm is incapable of. This adaptive detection method is phase independent and therefore it can be easily applied to three phase systems. Simulation and experimental results verify the good performance of the proposed new adaptive detection method.
1 INTRODUCTION

With the rapid development and application of power electronics devices, there has been a significant decrease in the power quality of most distribution systems worldwide. Power electronic devices provide power to many of today’s high tech products such as adjustable speed drives (ASDs), computer power supplies, medical equipment, uninterruptible power supplies (UPSs), etc., These devices generate significant harmonic currents into power systems. The main effects of these harmonics within the power systems are [1]:

1. Local system series or parallel resonances and these resonances cause consequent amplification of harmonic current levels to excessive magnitudes, which can lead to overloaded distribution and damage to capacitors in the system,
2. A reduction in the efficiency of the generation, transmission and utilization of electric energy, such as losses and torque disturbances in rotating machines,
3. Increased reactive power burden and aging of the insulation of electrical components, with consequent shortening of their life time,
4. Malfunctioning of the system or electrical components, such as failure of ripple control systems and protection relay systems, electromagnetic interference, and
5. Interference with communications, especially with telephone systems and instrumentation.

To improve the power quality of a power system, passive or active filters are usually used to cancel existing current harmonics to meet specific power quality related standards. With advances in power electronics devices, more active filters are applied. The key to control an active filter properly is the harmonic detection, which means identifying the harmonic information of a nonlinear load current. These harmonic currents should be identified quickly and precisely to provide the control references for the cancellation current controller. This dissertation proposes a new adaptive harmonic detection algorithm for power quality improvement, which has good steady state and dynamic performance compared to the state of the art popularly used d-q (transformation) based detection algorithm. In addition, this new adaptive harmonic detection algorithm has some advantages over popular d-q based harmonic detection algorithms, which will be explained later.
1.1 Background

IEEE standard 519-1992 [2] clearly defines the acceptable harmonic levels in the power systems. The summarized requirements of this standard are introduced in Chapter 2. To achieve this standard, passive filters were used for handling reactive power and harmonic problems. However, passive filters have disadvantages such as bulky space requirements, difficult frequency tuning processes, sensitive filter parameters, and thermal management issues. Active filters solve or prevent these disadvantages of the passive filters and can handle harmonic and reactive power problems well to improve a system’s power quality. Additionally, active filters can provide voltage support and solve other no periodic power quality conditions.

A warship possesses high power, complex distribution systems, thus the US Navy is interested in reconfigurable shipboard power systems (SPS) for its next generation ships. Therefore, the Navy desires to mitigate the main effects of increased power quality, because its ships have many sensitive devices. In summary, the power quality of a shipboard power network needs power improvement. Any solution applicable to ships can be broadly applied to many power systems.

An example of the proposed power quality improvement application in this dissertation is shown in Figure 1-1. This system has one IGBT converter, one nonlinear load and other linear loads. After the nonlinear load starts, it consumes nonlinear currents. Thus the grid side current will be distorted and the distorted current will also affect other loads in the system. When this happens, the power converter is controlled to have the function of active filtering or/and reactive power compensation to increase the power quality of the grid. In this dissertation, this means canceling the current harmonics of the grid (source) current. After the cancellation takes effect, the specific current harmonics in the source current can be attenuated. If more current harmonics need cancellation, more detected current harmonics from nonlinear load currents will be added as control references and the source current will become more sinusoidal. The power quality of the power network therefore can be improved.

1.2 Objectives and Contributions
To analyze, design and control such a power quality improvement system for harmonic cancellation, the following problems must be solved first. You need ways to characterize or detect harmonic sources. Also, you need the control methods to inject the canceling harmonics and fundamental currents into the power network in a flexible way. You need a stable dc bus voltage control method; the simplified control system is shown in Figure 1-2. The power quality application control algorithm gets the dc link voltage information and current harmonic information from the harmonic detection algorithm, then according to the power quality standard requirements and active filter power rating, proper harmonics are injected into the power network to cancel the harmonics existing in the source current. Meanwhile, fundamental current is also controlled to maintain the targeted dc link voltage. The main functionality for power quality improvement relies on the right operation of harmonic detection, which is the focus in this dissertation.

Figure 1-1 A simplified block diagram of a power quality improvement application

Figure 1-2 Simplified proposed control system
The work presented in this dissertation succeeds in solving these problems associated with harmonic detection; and, it provides a basis to the realization of the power quality improvement systems. Therefore, the overall objective of this dissertation work is to present solutions associated with harmonic detection for power quality improvement. Additionally, an experimental setup and experimental verification of the proposed algorithms is presented.

Specifically, The main contribution of this dissertation is to present a new adaptive harmonic detection algorithm for power quality improvement applications. The proposed new adaptive harmonic detection algorithm has advantages such as:

1. Easy tuning of adaptive gains for harmonic detection;
2. Better convergence properties, which makes the harmonic selective detection have better dynamic and steady state performance;
3. Solution of stability problems associated with adaptive detection algorithms;
4. Improved flexibility and easy applications to future multiple coordination control with more active filtering devices.
5. Easy applications to three-phase systems;
6. Fast frequency tracking ability when power system has big frequency change;
7. Works well when system currents have a dc offset.

In addition, a flexible and reconfigurable experimental apparatus is completed for power quality improvement research validation. It is a significant achievement that experimental results are available with developed advanced algorithms for the power quality application cases. Practical problems can be identified during experimental tests for future advanced research.

1.3 Outline

This dissertation is organized as follows:

Chapter 2 reviews the state of art in power quality improvement. This includes harmonic sources, passive filters, active filters and harmonic detection algorithms. Widely used harmonic detection algorithms for power quality improvement are compared and the shortcomings with existing algorithms are classified.

Chapter 3 introduces the popular used d-q transformation based harmonic detection algorithm and its working mechanism.
Chapter 4 introduces the new adaptive harmonic detection algorithm for power quality improvement. The nonlinear load current is modeled and Lyapunov based analysis is used to derive the proposed algorithm. This method has advantages such as phase independence, fast convergence speed and frequency estimation.

Chapter 5 gives the simulation results of the proposed new adaptive detection algorithm and also shows the effectiveness of using a tuning method for cases when choosing adaptive gains by hand tuning is difficult.

Chapter 6 introduces the experimental setup for testing the proposed adaptive detection algorithm. Then, the algorithm is tested experimentally both in hardware in the loop and electrical experiments.

Chapter 7 concludes the dissertation and discusses the future work associated with the adaptive detection algorithm.

The Appendix of this dissertation includes the introduction of the tuning method and supporting experimental results of a hardware in the loop power quality improvement application.

References are given at the end of this dissertation.
2 STATE OF ART

This chapter gives an overview of power quality research, especially for current harmonics related power quality research. Terminologies used in this dissertation are first introduced and then power quality issues, such as passive filters, active filters and harmonic detection, are discussed. These reviews help provide detailed background information leading to an understanding of the proposed harmonic detection’s application in power quality improvement.

2.1 Terminologies

There are several standards that provide power quality limits and mitigation applications. They include IEEE 519 [2], MIL1399 (Navy) [3] and IEC 61000-3-2 standard [4]. IEC 61000-3-2 standard is used in European countries, MIL1399 is used in US Navy and IEEE 519 is widely used as a guide in the USA. Standard IEEE 519 is the IEEE Guide for Harmonic Control and Reactive Compensation of Static Power Converters. The IEEE 519 harmonic limits are based on the ratio of the fundamental component of the load current, $I_L$, to the short circuit current, $I_{sc}$, at the point of the common coupling (PCC) with another supply, such as a utility. The odd harmonic limits for general distribution systems at voltages of 120V to 69kV are listed in Table 2-1. Stricter limits are imposed on larger loads than on small loads and it is the load side owner’s responsibility to meet this standard. The limits below TDD are in percent total harmonic distortion in applications to be explained later.

Power quality is defined as the concept of powering and grounding sensitive equipment in a manner that is suitable to the operation of that equipment [5]. However it still has narrower or no fixed definition for all applications. In this dissertation, a good power quality means that the current waveform is sinusoidal at the fundamental frequency has as few harmonics as possible, and that the current waveform has a small, expected linear phase relationship with its corresponding voltage. A current waveform with a large phase relationship is said to have low power factor. The fundamental current is controlled at or nearly in phase with voltage to unit or near unit power factor.
Fundamental (component) means the component of an order 1 (50Hz or 60Hz) of the Fourier series of a periodic quantity [5], such as voltage or current.

Harmonic (component) means a component of order greater than one of the Fourier series of a periodic quantity [5], such as voltage or current.

Harmonic content is the quantity obtained by subtracting the fundamental component from an alternating quantity [5], such as voltage or current.

Power factor means the ratio of the total power input in watts, to the total volt-ampere input to the converter [2].

In harmonic control applications, the total harmonic distortion (THD) factor and the total demand distortion (TDD) factor are used to describe the harmonic component of a power system in power quality. THD is defined as the ratio of the root sum square value of all the harmonics to the root mean square value of the fundamental [2]. THD can be expressed as shown in the following equation:

\[
THD(\text{current}) = \frac{\sqrt{I_3^2 + I_5^2 + I_7^2 + \ldots}}{I_1}
\] (2-1)

However, in real applications, a more practical TDD, which is defined as the total root sum square harmonic current distortion, in percent of the maximum demand load current \(I_{tr}\) [2], is used.

\[
TDD(\text{current}) = \frac{\sqrt{I_3^2 + I_5^2 + I_7^2 + \ldots}}{I_{tr}}
\] (2-2)

These harmonics usually are caused by nonlinear loads. Their possible solutions include canceling these harmonics using series filters or shunt filters.

<table>
<thead>
<tr>
<th>(\frac{I_n}{I_t})</th>
<th>(n &lt; 11)</th>
<th>(11 \leq n \leq 17)</th>
<th>(17 \leq n \leq 23)</th>
<th>(23 \leq n \leq 35)</th>
<th>(35 \leq n)</th>
<th>TDD</th>
</tr>
</thead>
<tbody>
<tr>
<td>&lt; 20</td>
<td>4.0%</td>
<td>2.0%</td>
<td>1.5%</td>
<td>0.6%</td>
<td>0.3%</td>
<td>5.0%</td>
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<tr>
<td>20 – 50</td>
<td>7.0%</td>
<td>3.5%</td>
<td>2.5%</td>
<td>1.0%</td>
<td>0.5%</td>
<td>8.0%</td>
</tr>
<tr>
<td>50 – 100</td>
<td>10.0%</td>
<td>4.5%</td>
<td>4.0%</td>
<td>1.5%</td>
<td>0.7%</td>
<td>12.0%</td>
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<td>100 – 1000</td>
<td>12.0%</td>
<td>5.5%</td>
<td>5.0%</td>
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<td>1.0%</td>
<td>15.0%</td>
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<tr>
<td>&gt; 1000</td>
<td>15.0%</td>
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</table>

Nonlinear load is a load that draws a nonsinusoidal current wave when supplied by a sinusoidal voltage source [5].
Series filter is a filter that reduces harmonics by putting a high impedance (coil) path in series between the harmonic source and the system to be protected [5].

Shunt filter is a filter that reduces harmonics by providing a low impedance capacitance path to ground of the system to be protected to shunt the harmonics from the source away from the system to be protected [5].

Passive filter is a filter built using passive components such as capacitors, inductances and resistance [5].

Active filter is a filter built by using controllable devices to overcome disadvantages of passive filters [5]. In this dissertation, the active filter is referred to as a power electronics device.

Harmonic cancellation is obtained by injecting a canceling current of the same amplitude and opposite phase as the harmonics into the power network to eliminate the harmonics.

Reactive power compensation is obtained by injecting a current that reduces the phase difference between the fundamental current and voltage.

2.2 Power Quality Issues

As stated in Chapter 1, due to a nonlinear load’s existence, harmonic currents flow in the power network and cause significant problems. This subsection introduces the harmonic sources, their effects and possible solutions for power quality improvement.

2.2.1 Harmonic Sources

There are a significant number of harmonic sources in a power network and these harmonic sources can be classified into the following groups [1] [6] [8] [11] [12]:

1. Transformer magnetization. The primary current will not be purely sinusoidal during the excitation process of the transformer because the flux is not linearly proportional to the magnetizing current. For example, the inrush current of a transformer usually contains significant harmonic content and a transformer experiences significant frequency variation and may lead to flux saturation and a nonlinear output.

2. Rotating machine harmonics. These harmonics are usually caused by motor pole unbalance, rotor phase unbalance and other forms of electro mechanical resonance.

3. Arcing devices. Electric arc furnace, discharge type lighting with magnetic ballasts and arc welders fall into this group. These devices are highly nonlinear and they produce significant harmonics.
(4) Rectification. These harmonics mainly can be divided into DC and AC power supplies, DC and AC motor drives, which are widely used in industry, consumer products and commercial power networks. These rectifiers are usually made by diode rectifiers, thyristor rectifiers, GTO and IGBT rectifiers. They generate significant current harmonics into the power networks and are of great concerns.

(5) Thyristor Controlled Reactors. These reactor applications include Static VAR Compensation and Thyristor Controlled Series Compensation (TCSC). They are normally used in high voltage transmission systems to provide reactive power compensation and voltage stability. However, they are also a source of harmonics.

2.2.2 Harmonic Effect

Harmonics (voltage or current) negatively affect a power system in many areas. These effects are classified into the following groups [1] [9] [12] [13] [14] [18] [20] [10]:

(1) Effects of resonances. Mainly, resonances occur due to the coupling of installed stray capacitance with installed stray inductance existing in transmission and distribution or load systems’ system capacitance and inductance.

(2) Effects on rotating machines. Harmonic currents cause electro mechanical power losses in rotating machines, which will heat the machine and cause reduction in the service life of the machine. Harmonics can also cause harmonic torques, which have little effect on mean torque, however can produce significant torque pulsations. These significant torque pulsations or cogging can cause significant noise and mechanical damage.

(3) Effects on static power plant and transmission systems. These effects apply to transmission systems, transformers and capacitor banks. Harmonics cause excessive losses due to heating. Harmonics also increase the system’s components ratings.

(4) Effects on measuring instruments. Measuring instruments work well on purely sinusoidal alternating current and subsequently a distorted electrical supply is prone to error.

(5) Effects on power system protection. Harmonics distort or degrade the operating characteristics of protective relays, especially digital relays and algorithms that rely on sample data or zero crossings. Protective devices are practically prone to error when harmonic distortion is present.
(6) Other Effects. Other bad effects associated with harmonics are interference and reduced reliability of end user equipment such as power supplies, television receivers, computers and communication devices.

2.2.3 Harmonic Elimination

There are two basic ways to achieve harmonic cancellation or reactive power compensation and improve the power quality of a power system. They are passive filtering and active filtering [1]. Passive filters are made only from passive elements and do not require an external power source. Most passive filters are linear and composed of just three basic linear elements -- resistors, capacitors, and inductors. In power quality applications, passive filters are usually shunt connected and tuned to the frequency that makes it’s the filter’s inductive and capacitive reactances equal. Thus, the harmonic currents are normally shunted to ground and prevented from entering the rest of the system. A shunt connection means that the filter provides a path to ground of low impedance at the harmonic frequencies.

Although passive filters have several advantages such as: no power consumption, cheap to build, suitable for high power rating applications and more linear than active filters, the design and construction of a proper passive filter that captures several harmonics without affecting the fundamental is very complicated and the designed filter is often bulky and possesses a high degree of parametric sensitivity [6] [8] [18] [20].

Active filters are another effective way of providing harmonic elimination and reactive power compensation. Due to recently developed power electronics technology and an expanding power electronics industry, high power active components are becoming cheaper and offering improved performance. Power electronics based active filters are made of power electronic devices such as GTOs and IGBTs. These filters are fully controllable for applications such as reactive power compensation, harmonic current or voltage elimination, neutral current compensation, terminal voltage regulation, voltage flicker suppression and voltage balance improvement [6] [8] [11] [12] [26]. The next subsection introduces active filters in detail and classifies the active filter topology used in this dissertation for selective harmonic cancellation.

2.3 Active Filters

Active filters are essentially converters (inverters) made of power electronics switches. They are divided into two types, current source converters (CSC) or voltage source converters (VSC).
Due to the VSC’s advantages and overall performances, most active power filters are VSC. According to the topologies of the converters, active filters can be classified into shunt active filters as shown in Figure 2-1, series active filters shown in Figure 2-2, universal active filters as in Figure 2-3 and hybrid active filters in Figure 2-4.

The shunt active filter is the filter that draws a compensating current from a power line to cancel harmonic currents on the source side, a grid location where power quality becomes important [6] [29] [37]. It is widely used to eliminate current harmonics, compensate reactive power and balance unbalanced currents by injecting (drawing) additional current, $i_{AF}$. The series active filter is the filter that is connected in series with the utility through a matching transformer, so it is controlled to eliminate voltage harmonics and regulate the terminal voltage of the load or line through the controllable voltage, $v_{AF}$ [6] [15] [16]. The universal active filter (Unified Power Quality Conditioner, UPFC)) is the filter that combines a shunt and a series active filters as shown in Figure 2-3. It is controlled for both voltage and current harmonics cancellation [17] [18]. The hybrid active filter is the filter that combines an active filter and a passive filter to reduce costs and improve efficiency [19] [20]. Due to the passive components, it often suffers significant parameter and frequency sensitivity.

![Figure 2-1 Topology of a shunt active filter](image-url)
Figure 2-2 Topology of a series active filter

Figure 2-3 Topology of a universal active filter

Figure 2-4 Typology of a hybrid active filter
Take a shunt active filter application for example as shown in Figure 2-5. In this system, a nonlinear load is powered by a voltage source $V_s$ and a VSC is shunt connected to the system. Resistance and inductance $R_s, R_L, R_{inv}$ and $L_s, L_L, L_{inv}$ are with the source, nonlinear load and inverter side, respectively. The current $i_s$ is the source current or grid side current, the current $i_L$ is the nonlinear current drawn by the nonlinear load, $i_{inv}$ is the drive active front end current and $i_{IM}$ is the motor side current. The voltage $v_{grid}$ is the grid side voltage and $v_{dc}$ is the dc link voltage. Therefore, for the success operation of such a system, the control should have the following modules.

Harmonic Detection: The nonlinear load current $i_L$ is measured to be sent to the detection algorithm. In this algorithm, the fundamental frequency of the current and different frequency component of the nonlinear load current are extracted. Those quantities are sent to a harmonic percentage calculation and harmonic selection module, where the individual harmonic’s percentage to fundamental current is calculated. According to harmonic selection standards, small percentage harmonics can be ignored and dominant harmonics can be classified. After this process, the fundamental current and harmonic currents can be used as reference currents for reactive power compensation and selective harmonic cancellation.

DC voltage control: the measured dc link voltage $v_{dc}$ is compared with reference dc voltage $v_{dc}^*$ and the error signal is controlled by a PI controller to get the reference current signal. A PLL (Phase Locked Loop) synchronizes the grid voltage $v_{grid}$ and multiply the reference current signal to get a fundamental three phase current reference signal. This signal is added with the reference current signal from harmonic detection to form a summarized current reference for the current control module.

Current control: The active front end current $i_{inv}$ is measured and compared with the reference sinusoidal current from harmonic detection and dc voltage control. The error signals are sent to a current controller (in the figure, it is a PI+Resonant controller). The outputs of the controller are sent to PWM modulation block to generate the required gate signals for the IGBT bridge.

This dissertation deals with the important harmonic detection part only to make the control reference be obtained precisely and quickly for the current control module. Successful harmonic detection algorithm is a must for the proper operation of the whole system.
Figure 2-5 Control block diagram of a shunt active filter
2.4 Harmonic Detection

This section introduces harmonic detection in power quality improvement applications. This is a control action and depends on the performance of harmonic extraction and identification algorithms. Without harmonic information, active power quality improvement application systems will not operate properly. Different methods have been proposed to measure or estimate current or voltage harmonics information. In this dissertation we deal only with current harmonics.

In the frequency domain, Discrete Fourier Transform (DFT) and Fast Fourier Transform (FFT) are widely used [6] [9] [10]. However, these methods have a very large computational burden and lack selective detection ability. Prony analysis was used in power system signal response analysis [21]. Prony analysis is also used for harmonic detection in power quality improvement applications by Li Qi and the author of this dissertation [22] [23]. It shows its ability to differentiate different frequency harmonic extracting ability and in addition, this method is also very good for decaying harmonics identification. However, real time implementation of this Prony analysis method is also computationally intense.

In the time domain, many methods have been proposed to detect, filter or estimate current or voltage information. Notch filters or band pass filters are used to extract harmonics by J. Sevensson [24] and M. Rastogin [25]. Resonant filters were also used by Pee-Chin Tan in [29] to get current references. However, it is difficult to implement the ideal amplitude and phase characteristics of a required filter. The performance of such a filter is sensitive to the parameters and operating conditions. Improved or adaptive notch filters are used for harmonic reference generation in active filter systems by Michael John Newman, etc. [26], Ribeiro M.V. etc.[27] and Byung-Moon Han [28]. However, they are very complicated and still sensitive to power system parameter variations.

A popular method for active filter reference generation is based on p-q theory, first proposed in the early 1980’s by Akagi, in which the instantaneous active and reactive power were computed in terms of transformed voltage and current signals via the $\alpha - \beta$ transformation. This can be found in the work by H. Akagi and F. Z. Peng [30] [31] [32] [33]. In the 1990s, the synchronous d-q reference frame method gained popularity. The d-q method was widely used for active filter harmonic detection and control in the work of V. Soares, H. Fujita, F. Z. Peng, and P. Jintakosonwit, etc.[34] [35] [36] [37] [38] [39]. D-q theory based filtering is popular because
fundamental current or harmonic currents can be converted to dc quantities in different rotating frames and low pass filters can easily extract these dc quantities with little phase delay consideration. Other methods that can be used in active filter applications are phase locked loops (PLL) methods and extended PLL methods, which has been investigated by J. Sebastian Tepper and M. Karimi-Ghartemani [40] [41] [42] [43].

The harmonic selectivity of active filters has been proposed by Po-Tai Cheng and his colleagues, Paolo and Ramadan [44] [45] [47] [48] [52]. Harmonic selection implies that different frequency components can be individually separated and identified, which means that any extracted harmonic is independent and can be used as a control reference according to control requirement. This concept is demonstrated by Figure 2-6, in which a distorted input current passes through a selective harmonic detection algorithm and the outputs are the separated different frequency components. Those components therefore can be used or combined as control references. A harmonic selective feature has advantages that allow for controllable selective harmonic cancellation, lower filter rating, and bandwidth requirements reductions.

In [44] [45] [47], a three-phase active filter (Dominant Harmonic Active Filter) can be used for specific harmonic cancellation, e.g. the 5th or the 7th. The control method proposed in Po-Tai Cheng’s work [48] uses d-q transformations in respective harmonic synchronous frames, which is complicated and only good for a three-phase system. More information about d-q theory is introduced in Chapter 3 of this dissertation. Other harmonic selective detection algorithms include the Kalman filter method proposed by Julio Barros [49], adaptive filters used by Siguo Luo 0 and Houshang Karimi[51] and an adaptive neuron network method applied by R. E. Shatshat[52], L.H. Tey [53], S. Osowski [54] and P.K. Dash [55] [56]. All these methods are summarized in Figure 2-7.

Adaptive detection algorithms have also been proposed for signal processing, especially in acoustic, See the research by Marc Bodson, Alexei Sacks and Douglas S.C [57] [58] [59] and research by Lyndon J. Brouwn [60]. The similar adaptive detection algorithm is successfully extended to power quality improvement application by the author of this dissertation [61] [62] [63]. The adaptive detection algorithm is found to have flexibility for harmonic selective detection, phase independent, frequency tracking ability. However, the stability and convergence are major concerns of using those methods. Therefore, this dissertation proposes a
new harmonic selective detection algorithm for power quality improvement with proved stability and quick convergence.

In most harmonic detection methods, the load current is modeled to represent the instantaneous current of the harmonic load as:

\[ i_L = I_{L1} \sin(\omega t + \phi_{L1}) + \sum_{h=3,5,\ldots}^{\infty} I_{lh} \sin(h\omega t - \phi_{lh}) \]

\[ = I_{L1} \sin(\omega t) \cos \phi_{L1} + I_{L1} \cos(\omega t) \sin \phi_{L1} + \sum_{h=3,5,\ldots}^{\infty} I_{lh} \sin(h\omega t - \phi_{lh}) \]

\[ = i_{L1p} + i_{L1q} + i_{lh} \quad (2-3) \]

In equation (2-3), \( i_{L1p} \) is the fundamental current corresponding to active power and \( i_{L1q} \) is the fundamental current corresponding to reactive power. Current \( i_{lh} \) represents the harmonic current with all orders \( h \). From equation (2-3), it is easy to see that the reference for a compensator providing power quality improving current injection should be \( -i_{L1q} - i_{lh} \).

![Figure 2-6 Harmonic selective detection concept](image)
Figure 2-7 Summary of harmonic detection methods
3 D-Q BASED HARMONIC DETECTION

The chapter reviews the d-q transformation used in popular transformation based harmonic detection algorithms. Section 3.2 therefore introduces the popular d-q transformation based harmonic selective detection algorithm. This method is proved to be quick and effective in three phase power quality improvement applications.

3.1 D-Q Transformation

The d-q transformation is a transformation of coordinates from the three phase stationary coordinate system to the d-q rotating coordinate system [64] [65]. This transformation is made of two steps:

1. Transformation from the three phase stationary coordinate system to the two phase \(\alpha - \beta\) stationary coordinate system;
2. Transformation from the \(\alpha - \beta\) stationary coordinate system to the d-q rotating coordinate system.

The first step is shown in Figure 3-1, where a three phase quantities \(u_a, u_b\) and \(u_c\) in \(abc\) stationary frame is transformed to a two axis system with \(\alpha\) axis in line with axis \(a\). From the relationships in projections and unchanged magnitude of formed vectors, the transformation matrix is obtained as:

\[
\begin{bmatrix}
1 & -\frac{1}{2} & -\frac{1}{2} \\
0 & \frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} \\
\frac{1}{2} & \frac{1}{2} & \frac{1}{2}
\end{bmatrix}
\]

The last row is used to obtain the zero sequence component of a three phase quantity and for a balanced system the last transform quantity becomes zero and only \(\alpha\) and \(\beta\) axis quantities exit. The second step is two axis stationary to two axis rotating frame transformation as shown in Figure 3-2, and the rotating axis \(d\) is with the speed of \(\omega\) with respect to axis \(\alpha\). Therefore the transformation matrix is obtained as:
Through this transformation, the ac component in $\alpha - \beta$ plane at speed of $\omega$ is transformed to dc component. The two steps are combined together to form a complete d-q transformation and the whole transformation matrix is obtained as:

\[
T = \frac{2}{3} \begin{bmatrix}
\cos \theta & \cos(\theta - \frac{2\pi}{3}) & \cos(\theta + \frac{2\pi}{3}) \\
\sin \theta & \sin(\theta - \frac{2\pi}{3}) & \sin(\theta + \frac{2\pi}{3}) \\
\frac{1}{2} & \frac{1}{2} & \frac{1}{2}
\end{bmatrix}
\] (3-3)

where $\theta = \omega t$.

As an example, assume that the three phase stator voltages are sinusoidal and balanced and are given by:

\[
v_a = V_m \cos(\omega t + \varphi) \\
v_b = V_m \cos(\omega t - \frac{2\pi}{3} + \varphi) \\
v_c = V_m \cos(\omega t + \frac{2\pi}{3} + \varphi)
\] (3-4)

Take the abc to d-q transformation to obtain:

\[
v_q = V_m \cos \varphi \\
v_d = -V_m \sin \varphi
\] (3-5)

The two quantities $v_q$ and $v_d$ are dc components now, which proves that the three phase ac quantities can be transformed to dc components in a rotating reference frame with the speed of the ac quantities. This transformation can also be applied to currents. An inverse transformation can transfer the dc components back to ac components by transformation matrix $T^{-1}$. 
Figure 3-1 $\alpha - \beta$ transformation

Figure 3-2 d-q transformation
3.2 D-Q Transformation Based Harmonic Detection

In this method, different rotating reference frame d-q transformations are used for extracting fundamental and harmonic currents. Through these transformations, corresponding fundamental current or harmonics become dc components and other untargeted frequency components still become ac component in the frame. Therefore, these components can be filtered out by low pass filtering (LPF). After an inverse d-q transformation in the respective frame, the unfiltered dc components are transformed back to corresponding harmonic. A targeted frequency components can then be separated from other frequency components in harmonic load currents. This process can be explained by Figure 3-3.

For harmonic selective detection, take the 5th harmonic extraction for example as shown in Figure 3-4. Load currents \( i_{La} \), \( i_{Lb} \) and \( i_{Lc} \) are measured and sent to the 3phase-to-2phase transformation (abc/dq transformation) in 5th harmonic rotating frame (300Hz). If the synchronous signals \( \sin \theta_s \) and \( \cos \theta_s \) are well provided through a PLL, the 5th harmonic currents become DC components in this d-q rotating frame while other frequency components still keep their AC form but with different frequencies. Through a well designed low pass filter, all the other AC components are filtered out except the dc component. Through an inverse d-q transformation (2phase to 3phase) in the same rotating reference frame, the dc components are transformed back to 5th harmonic (300Hz ac form). Therefore, the 5th harmonic is extracted from a distorted load current.

The same method can be applied to any frequency component. Different Phase Locked Loops (PLLs) are required to synchronize each rotating reference frame frequency. The \( \sin \theta_i \) and \( \cos \theta_i \) through \( \sin \theta_n \) and \( \cos \theta_n \) are required for the transformations. The whole control block diagram is shown in Figure 3-4, where different d-q transformations are used to get both reactive power compensation reference in Figure 3-4 (b) and selective harmonic compensation reference in Figure 3-4 (a). This method is widely used in harmonic cancellation and reactive power compensation applications.
Figure 3-3 Transformation procedure
Figure 3-4 Simplified control block diagram of d-q detection method
4 PROPOSED ADAPTIVE DETECTION

This section introduces the proposed adaptive detection algorithm for power quality improvement. This algorithm is derived and analyzed by using Lyapunov stability theory. Section 4.1 introduces the basis of Lyapunov theory for better understanding of the proposed algorithm. Section 4.2 derives the proposed harmonic detection algorithm in detail. Finally, some practical issues related to the application of the proposed algorithm are discussed.

4.1 Lyapunov Theory

The stability study of dynamic systems has a rich history. Lyapunov theory (Lyapunov stability theory) is introduced for better understanding of the proposed adaptive detection algorithm. The introduction of this theory is from the references written by Shankar Sastry [66], Jean Jacques E. Slotine and Weiping Li [67], and Petros A. Ioannou and Jing Sun [68].

Considering systems described by an ordinary differential equation of the form:

\[ \dot{x} = f(t, x), \ x(t_0) = x_0 \]  

(4-1)

Where \( x \in \mathbb{R}^n \), \( f : \mathbb{R} \times B(r) \mapsto \mathbb{R} \). The stability properties of the equilibrium state or solution of equation (4-1) can be found by the direct method of Lyapunov. The objective of this method is to answer questions of stability by using the form of \( f(t, x) \) rather than the explicit knowledge of the solutions. The Lyapunov theory is summarized as:

**Theorem**: Suppose there exists a positive definite function \( V(t, x) : \mathbb{R}^+ \times B(r) \mapsto \mathbb{R} \) for some \( r > 0 \) with continuous first order partial derivatives with respect to \( x, t \) and \( V(t,0) = 0 \ \forall \ t \in \mathbb{R}^+ \). Then the following statements are true:

(i) If \( \dot{V} \leq 0 \), then \( x_e = 0 \) is stable.

(ii) If \( V \) is decrescent and \( \dot{V} \leq 0 \), then \( x_e = 0 \) is uniformly stable

(iii) If \( V \) is decrescent and \( \dot{V} < 0 \), then \( x_e = 0 \) is uniformly asymptotically stable.
4.2 Adaptive Detection Algorithm

The proposed adaptive detection algorithm is an online estimation algorithm. The basic idea is to compare the observed system response \( y(t) \) with the output of a parameterized model \( \hat{y}(\theta, t) \), whose structure is the same as that of the plant model. The parameter vector \( \theta(t) \) is expecting continuous adjustment so that \( \hat{y}(\theta, t) \) approaches \( y(t) \) as time increases. Under certain input conditions, \( \hat{y}(\theta, t) \) will be close to \( y(t) \), which implies that \( \theta(t) \) is close to the unknown parameter vector \( \theta^* \) of the plant model. The algorithm development procedure includes two main steps: In the first step, an appropriate parameterization of the plant model is selected. The second step involves the selection of the adaptive law for generating or updating \( \theta(t) \). The adaptive law is usually a differential equation whose state is \( \theta(t) \) and is designed using stability considerations or optimization techniques to minimize the difference between \( y(t) \) and \( \hat{y}(\theta, t) \) with respect to \( \theta(t) \) at each time \( t \). The derivation of the proposed adaptive algorithm follows the two main steps.

4.2.1 Nonlinear Current Modeling

Assume the nonlinear load current has the following model:

\[
i_L = I_{L1} \sin(\omega_1 t + \phi_1) + I_{L3} \sin(\omega_3 t + \phi_3) + \ldots + I_{Ln} \sin(\omega_n t + \phi_n) + I_{dc} e^{-\sigma t}
\]  

(4-2)

In equation (4-2), \( n = 2k + 1, k = 0, 1, 2, 3, \ldots \) However for a three-phase system, the current harmonic component with order of \( n = 3k, k = 1, 2, 3, \ldots \) may not exist and other order harmonics may exist. If this happens, corresponding changes will occur at this specific frequency component. Additionally, a decaying dc current may exist for some cases. The fundamental has frequency \( \omega_1 \) is assumed to known. The phase \( \phi_n \) and amplitude \( I_{Ln} \) are unknown. The current \( i_L \) is measured through a sensor with transfer function \( W(s) \) to attenuate any possible higher frequency noise present. Therefore, we have:

\[
z = W(s)i_L = W(s)[I_{L1} \sin(\omega_1 t + \phi_1) + I_{L3} \sin(\omega_3 t + \phi_3) + \ldots + I_{Ln} \sin(\omega_n t + \phi_n) + I_{dc} e^{-\sigma t}]
\]  

(4-3)

Using the first two terms of the Taylor series expansion of dc component, equation (4-3) becomes:

\[
i_L = I_{L11} \sin \omega_1 t + I_{L12} \cos \omega_1 t + \ldots + I_{Ln1} \sin \omega_n t + I_{Ln2} \cos \omega_n t + A_{dc1} - A_{dc2}t
\]  

(4-4)
\[ I_{L11} = I_{L1} \cos \phi_1, I_{L12} = I_{L1} \sin \phi_1, \ldots, I_{Ln1} = I_{Ln} \cos \phi_n, I_{Ln2} = I_{Ln} \sin \phi_1 \] (4-5)

Equation (4-3) therefore can be rewritten in the form of \( z = W(s) \theta^T \phi \), where
\[
\theta^* = [I_{L11}, I_{L12}, \ldots, I_{Ln1}, I_{Ln2}, A_{dc1}, A_{dc2}]^T
\]
and
\[
\phi = [\sin \omega_1 t, \cos \omega_1 t, \ldots, \sin \omega_n t, \cos \omega_n t, 1, -1]^T.
\]
It becomes a parameter identification problem and from the estimate \( \theta = [\hat{I}_{L11}, \hat{I}_{L12}, \ldots, \hat{I}_{Ln1}, \hat{I}_{Ln2}, \hat{A}_{dc1}, \hat{A}_{dc2}] \) of \( \theta^* \), the estimates can be calculated by the following equations:

\[
\hat{I}_{Ln}(t) = \sqrt{\hat{I}_{Ln1}^2(t) + \hat{I}_{Ln2}^2(t)}, \hat{\phi}_n(t) = \cos^{-1}\left(\frac{\hat{I}_{Ln1}(t)}{\hat{I}_{Ln}(t)}\right), \left(\hat{I}_{Ld} e^{-\sigma t}\right) = \hat{A}_{dc1} - \hat{A}_{dc2} t
\] (4-6)

### 4.2.2 Direct Lyapunov Designed Adaptive Law

The adaptive law’s derivation is based on the procedure introduced in [68]. Therefore, let \( \theta(t) \) be the estimate of \( \theta^* \) at time \( t \), and the estimate \( \hat{z} \) of \( z \) at time \( t \) is constructed as:

\[
\hat{z} = W(s) \theta^T \phi
\] (4-7)

The estimation error \( \varepsilon_1 \) is therefore obtained as:

\[
\varepsilon_1 = z - \hat{z}
\] (4-8)

A careful look at \( \phi = [\sin \omega_1 t, \cos \omega_1 t, \ldots, \sin \omega_n t, \cos \omega_n t, 1, -1]^T \) shows that \( \phi \notin \ell_\infty \), therefore, normalization is required to get the proper adaptive algorithm. Normalizing the estimation error yields:

\[
\varepsilon = z - \hat{z} - W(s) \phi n_s^2 = \varepsilon_1 - W(s) \phi n_s^2
\] (4-9)

Where \( n_s \) is the normalizing signal, which satisfies the following condition:

\[
\frac{\phi}{m} \in \ell_\infty, m^2 = 1 + n_s^2
\] (4-10)

Therefore, \( n_s^2 = \phi \phi^T \) or \( n_s^2 = \phi P \phi^T \) for any \( P = P^T \) is chosen to satisfy equation (4-10).

Since \( z = W(s) \theta^T \phi \) and \( \hat{z} = W(s) \theta^T \phi \), equation (4-9) can be expressed as equation (4-11) by substituting \( z \) and \( \hat{z} \).

\[
\varepsilon = W(-\phi^T \phi - \phi n_s^2)
\] (4-11)

where
\[
\bar{\theta} = \theta - \theta' \tag{4-12}
\]

A state space representation of equation (4-11) is then obtained as:
\[
\begin{align*}
\dot{e} &= A_c e + B_c (-\bar{\theta}^T \phi - s \varepsilon s) \\
\varepsilon &= C_c^T e \tag{4-13}
\end{align*}
\]

Where \(A_c, B_c\) and \(C_c\) are the matrices associated with a state space representation that has a transfer function \(W(s) = C_c^T (sI - A_c)^{-1} B_c\). The error equation (4-13) relates \(\varepsilon\) with the parameter error \(\bar{\theta}\) and is used to construct an appropriate Lyapunov type function for designing the adaptive law of \(\theta\). However, the state error \(e\) cannot be measured or generated because of the unknown input \(\bar{\theta}^T \phi\).

Considering the following Lyapunov-like function:
\[
V(\bar{\theta}, e) = \frac{e^T P_c e}{2} + \frac{\bar{\theta}^T \Gamma^{-1} \bar{\theta}}{2} \tag{4-14}
\]

In equation (4-14), \(\Gamma = \Gamma^T > 0\) is a constant matrix and \(P_c = P_c^T > 0\) satisfies the algebraic equations:
\[
\begin{align*}
P_c A_c + A_c^T P_c &= -q q^T - \nu L_c \\
P_c B_c &= C_c \tag{4-15}
\end{align*}
\]

for some vector \(q\), matrix \(L_c = L_c^T > 0\) and \(\nu > 0\).

Considering the time derivative \(\dot{V}\) along the solution of equation (4-15), the following equation is obtained as:
\[
\dot{V}(\bar{\theta}, e) = -\frac{1}{2} e^T q q^T e - \frac{1}{2} v e^T L_c e - e^T P_c B_c (\bar{\theta}^T \phi - s \varepsilon s) + \bar{\theta}^T \Gamma^{-1} \hat{\theta} \tag{4-16}
\]

Since \(P_c B_c = C_c \Rightarrow e^T P_c B_c = e^T C_c = \varepsilon\), equation (4-16) can be rewritten as:
\[
\dot{V}(\bar{\theta}, e) = -\frac{1}{2} e^T q q^T e - \frac{1}{2} v e^T L_c e - \varepsilon \bar{\theta}^T \phi - \varepsilon s^2 n_s + \bar{\theta}^T \Gamma^{-1} \hat{\theta} \tag{4-17}
\]

Choose
\[
\dot{\theta} = \hat{\theta} = \Gamma \varepsilon \phi \tag{4-18}
\]

the time derivative \(\dot{V}\) becomes:
\[ V(\tilde{\theta}, e) = -\frac{1}{2} e^T q q^T e - \frac{1}{2} v e^T L e - e^2 n_s^2 \leq 0 \]  

(4-19)

Therefore equation (4-18) is the adaptive law chosen for stability and the adaptive law can be written as:

\[
\begin{align*}
\dot{I}_{L11} &= \gamma_{11} e \sin \omega t, \quad \dot{I}_{L12} = \gamma_{12} e \cos \omega t \\
\dot{I}_{L31} &= \gamma_{31} e \sin \omega_3 t, \quad \dot{I}_{L32} = \gamma_{32} e \cos \omega_3 t \\
& \cdots \\
\dot{I}_{Ln1} &= \gamma_{n1} e \sin \omega_n t, \quad \dot{I}_{Ln2} = \gamma_{n2} e \cos \omega_n t \\
\dot{A}_{dc1} &= \gamma_{d1} e, \quad \dot{A}_{dc2} = \gamma_{d2} e.
\end{align*}
\]  

(4-20)

Equation (4-20) gives the detailed form of different harmonic estimation formulas, where \( \gamma_{n1}, \gamma_{n2} \) and \( \gamma_{dc} \) are adaptive gains. The adaptive law of equation (4-20) guarantees that

(i) \( \theta, e \in \ell_\infty \)

(ii) \( e, \omega n, \tilde{\theta} \in \ell_2 \)  

(4-21)

Independent of the bounded properties of \( \phi \).

Setting \( \dot{e} = 0 \) in equation (4-13) and solve the “quasi” steady state response \( e_{ss} \) of \( e \), we have:

\[ e_{ss} = \frac{\alpha (-\tilde{\theta}^T \phi)}{1 + \alpha n_s^2} = \frac{e_{1ss}}{1 + \alpha n_s^2} \]  

(4-22)

Where \( \alpha = -C_c^T A_c^{-1} B_e \) is positive because of the SPR property of \( W \) and \( e_{1ss} \) is the “quasi” steady state response of \( e_1 \). Because \( n_s, e_{ss} \) cannot become unbounded as a result of a possibly unbounded signal \( \phi \), large \( e_{ss} \) implies that \( \tilde{\theta} \) is large and large \( e \) carries information about \( \tilde{\theta} \), which is less affected by \( \phi \). Since \( \phi \) contains the frequency information of \( \omega \), it also means that \( e \) is less affected by \( \omega \) than \( \theta \). In some cases, the fundamental frequency of the power network does change, therefore the fundamental frequency should be updated too. With the estimated components from equation (4-20), the instantaneous means square error is constructed as:

\[ e(t) = 0.5 \left( \sum_{k=1}^{N} (I_{L1k} \sin n\omega t + I_{L2k} \cos(n\omega t)) - i_e(t) \right)^2 \]  

(4-23)

Therefore, the derivative of the mean square error with respect to the fundamental frequency \( \omega_1 \) is derived as:
\[ \dot{\omega}_1 = -\alpha \varepsilon \left[ \sum_{n=1}^{N} (\hat{I}_{Ln1} \cos n\omega_1 t - \hat{I}_{Ln2} \sin n\omega_2 t) \right] \]  

(4-24)

where \( \alpha \) is an adaptive gain.

The control block diagram using this method is shown in Figure 4-1. The load current \( i_L \) is measured and compared with the estimated load current \( i_{Le} \) to generate the error signal. Using the weights updating algorithm, the corresponding coefficients of sine/cosine vectors can be updated to get accurate fundamental and harmonics information of the load current. The sine/cosine vectors can be generated by an adaptive way by using equation (4-24).

4.2.3 Gradient Designed and Lyapunov Analyzed Adaptive Law

However, the introduced above adaptive detection algorithm cannot guarantee the pattern of convergence speed with high adaptive gains. The convergence is not promised in applications when more harmonics need to be identification. An improved new adaptive detection algorithm is proposed in this dissertation by using an integral cost function and gradient search optimization. The cost function criterion is chosen for available measurements, which leads to sensitivity functions that are implementable. The relationship of the estimation error with the estimated parameters \( \theta \) is chosen to make sure that the cost function is convex. Therefore its gradient with respect to the estimated parameters is implementable and the gradient method is used here to get the adaptive laws. Thus, an adaptive law can be quickly derived. Afterwhich, Lynamunov theory is used to analyze the derived algorithm’s properties such as stability and convergence.

![Figure 4-1 Simplified control block diagram of adaptive neuron method](image_url)
The integral cost function is expressed as:

\[ J(\theta) = \frac{1}{2} \int_0^\tau e^{-\beta(t-\tau)} e^2(t, \tau) m^2(\tau) d\tau \]  

(4-25)

Where \( \beta > 0 \) is a design constant and the error \( \varepsilon \) is defined as:

\[ \varepsilon(t, \tau) = \frac{z(\tau) - \theta^T(t) \phi(\tau)}{m^2(\tau)}, \varepsilon(t, t) = \varepsilon \]  

(4-26)

Equation (4-25) is the normalized estimation error at time \( \tau \) based on the estimate \( \theta(t) \) of \( \theta^* \) at time \( t \geq \tau \). The design constant \( \beta \) acts as a forgetting factor, and as time \( t \) increases the effect of the old data at time \( \tau < t \) is discarded exponentially. Therefore, the estimation parameter \( \theta(t) \) is updated at each time to minimize the integral square of the error on all past data that are discounted exponentially. Note that the exponential term of (4-27) may go to 0 much faster than \( \varepsilon \) or \( m \), making (4-28) an invalid Lyapunov function candidate.

Expressing equation (4-25) in terms of the parameter \( \theta \), the following equation cost function can be obtained:

\[ J(\theta) = \frac{1}{2} \int_0^\tau e^{-\beta(t-\tau)} \left( \frac{z(\tau) - \theta^T(t) \phi(\tau)}{m^2(\tau)} \right)^2 d\tau \]  

(4-29)

Since \( J(\theta) \) is convex over the space of \( \theta \) for each time \( t \), the steepest descent method to minimize \( J(\theta) \) with respect to \( \theta \) yields:

\[ \dot{\theta} = -\Gamma \Delta \theta = \Gamma \int_0^\tau e^{-\beta(t-\tau)} \left( \frac{z(\tau) - \theta^T(t) \phi(\tau)}{m^2(\tau)} \right) \phi(\tau) d\tau \]  

(4-30)

In equation (4-30), \( \Gamma = \Gamma^T > 0 \) is the adaptive gain and has the same dimension in (4-20). Rewriting equation (4-30), we have the integral adaptive law for load current identification as:

\[ \dot{\theta} = -\Gamma (R(t) \theta + Q(t)) \]

\[ \dot{R} = -\beta R + \frac{\phi \phi^T}{m^2} \]

\[ \dot{Q} = -\beta Q - \frac{z \phi}{m^2} \]  

(4-31)

Where \( R(0) = 0 \) and \( Q(0) = 0 \).

This adaptive detection method has better convergence properties when compared with an instantaneous objective function. However, this method has more gains to tune, such as \( \beta \).
simplified control block diagram of this algorithm is shown in Figure 4-2. The sine/cosine vector is also required for identify different frequency component, which can be obtained by an adaptive way shown in previous section.

The adaptive law in equation (4-31) guarantees that if $n_s, \phi \in L_\infty$ and $\phi$ is persistently exciting (PE) then $\theta(t)$ converges exponentially to $\theta^*$ and the rate of convergence can be made arbitrarily large by increasing the value of the adaptive gains. This convergence is proved by using the methods found in reference [68]. The convergence proof is summarized as the followings.

Because $\frac{m}{m} \in L_\infty$, so do $R, Q \in L_\infty$ and the differential equation for $\theta$ behaves like a linear time varying differential equation with a bounded input. Substituting $z = \phi^T \theta^*$ in the differential equation for $Q$, the $Q$ is obtained as:

\[
Q(t) = - \int_0^t e^{-\beta(t-\tau)} \phi(\tau)\phi^T(\tau) \frac{d\tau}{m^2} - R(t)\theta^* \quad (4-32)
\]

Then, we further have:

\[
\dot{\theta} = \dot{\tilde{\theta}} = -\Gamma R(t)\tilde{\theta} \quad (4-33)
\]

Using the Lyapunov-like function:

\[
V(\tilde{\theta}) = \frac{\tilde{\theta}^T R(t)\tilde{\theta}}{2} \quad (4-34)
\]

Figure 4-2 Control block diagram of adaptive detection with integral cost function
And the derivative along the solution of equation (4-33) is given by:

\[ \dot{V} = -\tilde{\theta}^T R(t) \tilde{\theta} \]  \hspace{1cm} (4-35)

Because \( R(t) = R^T(t) \geq 0 \) for \( \forall t \geq 0 \), \( \dot{V} \leq 0 \). This proves that the proposed adaptive law is stable. And from equation (4-31), the following \( R(t) \) can be obtained:

\[ R(t) = \int_0^t e^{-\beta(t-\tau)} \frac{\phi(\tau)\phi^T(\tau)}{m^2(\tau)} d\tau \]  \hspace{1cm} (4-36)

If \( \phi \) is PE and \( m \) is bounded, the following equation can be derived:

\[
\begin{align*}
R(t) &= \int_{-T_o}^t e^{-\beta(t-\tau)} \frac{\phi(\tau)\phi^T(\tau)}{m^2(\tau)} d\tau + \int_0^{-T_0} e^{-\beta(t-\tau)} \frac{\phi(\tau)\phi^T(\tau)}{m^2(\tau)} d\tau \\
&\geq \alpha_0 e^{-\beta \lambda_0} \int_{-T_0}^t \phi(\tau)\phi^T(\tau) d\tau \\
&\geq \beta_1 e^{-\beta \lambda_0} I
\end{align*}
\]  \hspace{1cm} (4-37)

for any \( t \geq T_0 \), where \( \beta_i = \alpha_0 \alpha_0 T_0, \alpha_0 = \sup_i \frac{1}{m^2(t)} \) and \( \alpha_0, T_0 > 0 \) are constants given by the definition of PE. Therefore:

\[ \dot{V} \leq -\beta_1 e^{-\beta \lambda_0} \tilde{\theta}^T \tilde{\theta} \leq -2\beta_1 \lambda_{\text{min}}(\Gamma) e^{-\beta \lambda_0} V \]  \hspace{1cm} (4-38)

For \( t \geq T_0 \), which also implies that \( V(t) \) satisfies

\[ V(t) \leq e^{-\alpha(t-T_0)} V(T_0), t \geq T_0 \]  \hspace{1cm} (4-39)

where \( \alpha = 2\beta_1 e^{-\beta \lambda_0} \lambda_{\text{min}}(\Gamma) \). Therefore, \( V(t) \to 0 \) as \( t \to \infty \) exponentially fast with a rate equal to \( \alpha \). The parameter error is therefore satisfies the following condition:

\[ \sqrt{2\lambda_{\text{min}}(\Gamma)} V \leq |\tilde{\theta}| \leq \sqrt{2\lambda_{\text{max}}(\Gamma)} V \]  \hspace{1cm} (4-40)

And the right part of the above condition can be further written as:

\[ |\tilde{\theta}| \leq \sqrt{2\lambda_{\text{max}}(\Gamma)} V(T_0) e^{-\frac{\alpha}{2}(t-T_0)} \leq \sqrt{\frac{\lambda_{\text{max}}(\Gamma)}{\lambda_{\text{min}}(\Gamma)}} \tilde{\theta}(T_0) e^{-\frac{\alpha}{2}(t-T_0)} \]  \hspace{1cm} (4-41)

This equation clearly shows that the estimation \( \theta(t) \) approaches the unknown parameters \( \theta^* \) exponentially fast with a rate of \( \frac{\alpha}{2} \), which can be made large by increasing the value of the adaptive gain \( \Gamma \).
The PE properties of $\phi$ can be proved by ignoring the dc component in the nonlinear load current (common in many applications) and considering the following PE related function:

$$\frac{1}{T_0} \int_{\tau}^{\tau+T_0} \phi(\tau) \phi^T(\tau) d\tau$$

(4-42)

For $T_0 = \frac{2\pi}{\omega_1}$, equation (4-42) becomes:

$$\frac{1}{T_0} \int_{\tau}^{\tau+T_0} \phi(\tau) \phi^T(\tau) d\tau = \left[ \begin{array}{cccc} \pi / \omega_1 & \ldots & \ldots & \ldots \\ \ldots & \pi / \omega_1 & \ldots & \ldots \\ \ldots & \ldots & \pi / \omega_1 & \ldots \\ \ldots & \ldots & \ldots & \pi / \omega_1 \end{array} \right]$$

(4-43)

Choose $0 < \alpha_0 < \frac{\pi}{\omega_1}, \alpha_1 \geq \frac{\pi}{\omega_1}$, the following condition is satisfied:

$$\alpha_1 I \geq \frac{1}{T_0} \int_{\tau}^{\tau+T_0} \phi(\tau) \phi^T(\tau) d\tau \geq \alpha_0 I$$

(4-44)

Therefore $\phi$ is PE. Since $\phi \in \ell_\infty$ and $\phi$ is PE, then $\theta$ converges exponentially to $\theta^*$ and $\hat{I}_{Ln}, \hat{\phi_n}$ converge to $I_{Ln}, \phi_n$ exponentially fast. This proves the important feature of this method is that the adaptive gains can be choose arbitrary large for fast the convergence speed in theory, which can be shown in the simulation part.

This theory is only valid when the current has the exact model used and no dc current is present. This condition and subsequent convergence are not always satisfied. Especially when the gains are not chosen properly. Although the improved adaptive detection algorithm is relatively insensitive to model uncertainty, real world current signals may have dc offset and more high order harmonics. When modeling such a real current, the adaptive gains cannot be chosen arbitrary large. Therefore, a particle swarm optimization method for tuning many adaptive gains is also proposed in appendix 8.1 in this dissertation.
5 SIMULATIONS

This chapter presents the simulation evaluation work of the proposed adaptive harmonic detection algorithm and its comparison with d-q transformation based harmonic detection. This chapter is organized as follows: Section 5.1 introduces the simulation of the proposed new adaptive detection algorithm and its comparison with popular d-q based harmonic detection algorithm, and Section 5.2 evaluates the application of the particle swarm optimization (PSO) tuning method in adaptive gains setting. Detailed information about PSO tuning process can be found in Appendix 8.1.

5.1 Proposed Adaptive Detection Algorithm

This section gives simulation results of the proposed harmonic detection methods for harmonic detection. A random synthesized test current signal is used as the stimulus and has the parameters shown in Table 5-1. The parameters in brackets are used for load current change tests. The d-q based method, and the proposed adaptive method are tested in these simulations for steady state, start transient, load transient and dc offset currents.

<table>
<thead>
<tr>
<th>Table 5-1 Test current parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>Fundamental</td>
</tr>
<tr>
<td>Magnitude (A)</td>
</tr>
<tr>
<td>Phase (rad)</td>
</tr>
</tbody>
</table>

5.1.1 Gain’s Effects on Proposed Adaptive Algorithm

Simulation results in Figure 5-1 show the convergence speed obtained by choosing different adaptive gains to verify the claim that the rate of convergence can be made relatively large by increasing the value of the adaptive gains for real time applications. The forgetting factor $\beta$ is set as a constant 0.2 and the adaptive gains $\Gamma$ are set as 100, 1000 and 10000, respectively. Initially, these gains are arbitrarily chosen for documentation purposes. They are the same for each frequency component. They are not unique and finding the gains is not in this work’s scope. However, in appendix one technique, particle warm optimization can be used for gain selection. Figure 5-1 (a) shows the estimation errors with different gains, where the bigger the gain is, the faster the convergence speed it becomes. Another result from fundamental current estimations in Figure 5-1 (b) clearly demonstrates that with gains up to 10000, the start transient can be limited...
to one cycle, which satisfies most power quality applications. This test fully verifies that the rate of convergence can be made relatively large by increasing the value of the adaptive gains.

Figure 5-1 Convergence speed with different gains
5.1.2 Steady State Test

The first test is to compare the steady state performance of the proposed adaptive algorithm and the popular d-q algorithm. The test current signal’s parameters are shown in Table 5-1 and Figure 5-2 and Figure 5-4 show the simulation results. Only fundamental and 13th harmonic are demonstrated in Figure 5-2. The estimated current waveforms are overlapped with the reference current signal and it is clear that both methods have good steady state performance. The spectrum analysis shown in Figure 5-4 further verifies the good performances of the proposed adaptive detection algorithm and the d-q algorithm.

5.1.3 Start Transient Test

Figure 5-3 shows the start transient results of both methods. A start transient refers to the initial adaptation period of the algorithm and its simulation response is a good indicator of the algorithm’s ability to track. Only fundamental and 13th harmonic results are shown to give a comparison here. Adaptive detection algorithm has good dynamic performance compared with traditional d-q method. An example is shown in Table 5-2, where five sampling points are chosen for comparison between two methods. From sampling points 1000 (10ms), the adaptive method is already convergent while D-Q method still has significant errors compared with reference current. From this comparison, one can assume the proposed adaptive method will have better tracking properties due to its large adaptive gains.

5.1.4 Dynamic Transient Test

Figure 5-5 shows the dynamic transient results of both methods. The input current changes its magnitudes and phases of different frequencies as shown in Table 5-1. Figure 5-5 (a) shows the fundamental current’s results and both can get fast and accurate estimations in one cycle (around 16.7ms). For 13th harmonic’s transient change, both methods can complete fast and accurate estimations within one cycle (around 15ms) as shown in Figure 5-5 (b). A careful look in Table 5-2 shows that the proposed adaptive algorithm has better convergence speed in the transient. The percentage of absolute error is smaller compared with D-Q method.

5.1.5 DC Offset Tests

Figure 5-6 (a) shows that dc offset is tracked well with the proposed adaptive detection method and both dynamic performance and steady state performance are good. The d-q method is not designed to, nor is it capable of estimating dc offset. Figure 5-6 (b) shows the estimated fundamental current results. Estimated fundamental current from d-q method does not match the
reference current and it leaves dc. The low pass filters are not working as well as those without dc offset, because the fundamental and dc offset only have 60Hz frequency difference even in rotating frame. The simulated different frequency components are added together to get the final signal for spectrum analysis in Figure 5-6 (c). D-q method in this situation is not as good as the proposed adaptive method.

Table 5-2 Comparison during start transient (fundamental current)

<table>
<thead>
<tr>
<th>Sampling points (1e-5s)</th>
<th>Reference Current</th>
<th>D-Q Method</th>
<th>Adaptive method</th>
</tr>
</thead>
<tbody>
<tr>
<td>1000</td>
<td>2.322</td>
<td>1.180</td>
<td>2.323</td>
</tr>
<tr>
<td>1100</td>
<td>3.789</td>
<td>2.269</td>
<td>3.789</td>
</tr>
<tr>
<td>1200</td>
<td>4.724</td>
<td>3.102</td>
<td>4.724</td>
</tr>
<tr>
<td>1300</td>
<td>4.995</td>
<td>3.540</td>
<td>4.995</td>
</tr>
<tr>
<td>1400</td>
<td>4.565</td>
<td>3.380</td>
<td>4.565</td>
</tr>
</tbody>
</table>

Table 5-3 Comparison during start transient (fundamental current)

<table>
<thead>
<tr>
<th>Sampling points (1e-5s)</th>
<th>Reference Current</th>
<th>D-Q Method</th>
<th>Percentage of Absolute Error</th>
<th>Adaptive method</th>
<th>Percentage of Absolute Error</th>
</tr>
</thead>
<tbody>
<tr>
<td>26000</td>
<td>-3.4602</td>
<td>-0.85487</td>
<td>72.1%</td>
<td>-2.7298</td>
<td>21.1%</td>
</tr>
<tr>
<td>26100</td>
<td>-0.9772</td>
<td>0.92543</td>
<td>194.2%</td>
<td>-0.67599</td>
<td>30.8%</td>
</tr>
<tr>
<td>26200</td>
<td>1.6431</td>
<td>2.7093</td>
<td>64.9%</td>
<td>1.6064</td>
<td>2.2%</td>
</tr>
<tr>
<td>26300</td>
<td>4.0326</td>
<td>4.3061</td>
<td>8.1%</td>
<td>3.8041</td>
<td>5.6%</td>
</tr>
<tr>
<td>26400</td>
<td>5.8557</td>
<td>5.4641</td>
<td>6.7%</td>
<td>5.6449</td>
<td>3.6%</td>
</tr>
</tbody>
</table>

(a) Fundamental current

Figure 5-2 Steady state results
Figure 5-2 continued

(a) Fundamental current

(b) 13\textsuperscript{th} harmonic current

Figure 5-3 Start transient results
Figure 5-4 Steady state spectrum analysis

(a) Fundamental current

(b) 13th harmonic current

Figure 5-5 Dynamic transient results
(a) DC offset estimation results

(b) Fundamental current results

(c) Spectrum analysis of estimated currents

Figure 5-6 DC offset results
6 EXPERIMENTAL RESULTS

To verify this proposed new adaptive harmonic detection algorithm and other power quality applications, a completed experimental apparatus is established for hardware feasibility, system integration, and rapid control validation purposes at the Center for Advanced Power Systems, Florida State University. This experimental system fulfills the requirements for testing the proposed novel algorithm in short term and long term plans. The establishment of the experimental system is also an important validation component of this dissertation work and contributes to the research on future reconfigurable induction motor drive systems and other related power quality improvement research.

6.1 Experimental Apparatus

This system consists of six 28.8KVA variable frequency drives supplying a variable frequency experimental bus, an 11.3kW specific designed permanent magnetic synchronous motor (PMSM), and an 11.3 kW induction motor (IM). As shown in Figure 6-1 two drives are connected in a back-to-back structure in VS1 to generate a Variable Voltage Variable Frequency (VVVF) bus through LC filter with power provided by CAPS utility bus 1. Subsystems DS1, DS2, NL1 can be connected to CAPS bus 1 or generated experimental bus by VS1. The three subsystems, DS1, DS2, and NL1, are introduced as follows.

DS1 contains two drives. The first one supplies the power and control to the PMSM. A dc motor is coupled with the PMSM mechanically. The second drive’s dc bus is connected with the first drive through a breaker. Therefore, the PMSM can work in both back-to-back connection mode and diode rectifier converter mode by choosing proper switching functions. Based on this setup, the PMSM drives can be configured as back-to-back structure for active filtering cancellation work or using diode rectifier as nonlinear load for harmonic detection algorithm research.

DS2 also contains two drives. The first one supplies power and control to the induction motor (IM). The second drive’s dc bus is also connected with the first drive through a breaker. Hence, the IM can work in both back-to-back connection mode and diode rectifier converter mode.
When IM works in back-to-back connection, reconfigurable IM motor drive research can be performed when it is working in diode rectifier mode, it will be considered as a nonlinear load for harmonic detection algorithm research, which is further supporting the reconfigurable power conversion system research.

NL1 consists of a programmable thyristor converter as a nonlinear load when DS1 and (or) DS2 is working in reconfigurable control mode. This device will be used for active filtering research by providing significant current harmonics into the system.

The system shown in Figure 6-1 is also connected with CAPS RTDS® and dSPACE® controller to form a flexible hardware in the loop (HIL) test apparatus as shown in Figure 6-2. Control algorithm can run in a dSAPCE® controller in real time and it can control a virtual electrical system in RTDS® for the controller-HIL tests. The RTDS can solve electrical system models in real-time at time-steps down to typically 50 micro-seconds, with solutions for portions of the system possible at time-steps down to typically 2 μs utilizing the gigahertz processor card (GPC) cards. The RTDS® system in CAPS has 14 racks. Each rack can solve a 54-node electrical network, for a total of 756 electrical nodes in real-time (plus associated controls and other systems) with all 14 racks. Therefore, the power quality improvement electrical circuit can fully run in RTDS for real time algorithm verification.

The dSPACE® controller also can control all the electronic devices in the electrical system for hardware experiments. The targeted electrical system can also be coupled to a higher level power system in RTDS® through interface VS1 for the Power-HIL tests. A desktop computer running Matlab® and RSCAD® is working as a user interface for algorithm development, algorithm downloading, virtual system molding and downloading and data acquisition, etc. The effect of coupling the proposed reconfigurable power conversion system to an even higher lever system can therefore be evaluated. Based on this structure, the proposed system has plenty of advantages and unique characteristics for algorithm verification and power quality improvement research and more details can be found in the paper written by the author of this dissertation [74]. The establishment of the experimental apparatus provides several unique characteristics and contribution to the dissertation work. It makes the experimental verification of the harmonic detection algorithm and power quality improvement research possible.
Figure 6-1 Experimental System Overview
Figure 6-2 RT-HIL Experimental System Overview
6.1.1 RT-HIL Abilities

The RTDS® system provides powerful ability to simulate a power system in real time. For controller HIL tests, the targeted electrical system is simulated in RTDS®. Through A/D, D/A, D/D and PWM interfaces, the controller can read electrical information from this virtual system, and then it can control the virtual electrical system in real time. The developed control algorithm can hence be verified quickly before it is applied to physical electrical equipment. The power HIL provides researchers ability to check a targeted electrical system’s impact on the power system, to which it would be connected.

6.1.2 Fast and Efficient Prototyping Process

With the proposed experimental setup, the future research cycle of control algorithm development and other related power quality improvement research can be greatly reduced. Figure 6-3 shows an example research development cycle for a control algorithm implementation to an electrical system. At the initial stage, the control algorithm is developed and simulated in Matlab/Simulink®. When it is verified, it is compiled using Real Time Workshop (RTW) and runs in real time using personal computer processors. Proper choice of time steps and other issues related to the real time implementation can be identified at this stage. The verified real time algorithm is then downloaded to dSPACE® controller while the targeted electrical system is simulated in RTDS® in real time.

The controller’s performance can thus be fully tested without damaging any real electrical equipment. The fully tested control algorithm is now ready to be applied to the electrical systems so that their performance can be studied. Later, the developed control algorithm and the targeted system can be combined to an even bigger or higher level power system in order to study the impact of adding new electrical components into the existing power system. Usually, a failed stage means going back to the previous stage for more tests.

6.2 Harmonic Detection

This section introduces the experimental results associated with the proposed novel harmonic detection algorithm. The first part of the experimental results is related to hardware in the loop experimental results and the next subsection of the experimental results is related to electrical experiment results. The proposed method is also compared with the popular used d-q based detection algorithm.
6.2.1 HIL Experiments

To make the proposed algorithm suitable for real time application, proper time step and other hidden issues related to real time implementation must be classified before its application to real experiments. The controller’s performance then can be fully tested without damaging any electrical equipment. Therefore, RT (real time) controller HIL (hardware in the loop) test is first performed and the tested control algorithm is then applied to the electrical systems. The performance of the proposed adaptive algorithm is compared with the common used d-q method. By using RT-HIL tests, the control algorithm can run in real time by a physical controller and nonlinear current is generated by a RTDS (Real time digital simulator) working as signal generator. Through this test, the proposed algorithm can be fully tested by setting suitable time step, sampling rate, etc. Through this test, different frequency components can be analyzed and compared to investigate the accuracy of the proposed method in different frequency components.

Figure 6-3 Design and development procedure
The RT HIL experimental setup is shown in Figure 6-4 by using the experimental apparatus introduced in above subsection. The algorithms are calculated quickly enough to continuously produce output conditions representing real world conditions. Physical devices (controller with algorithms in our applications) can therefore be tested in real-time. The real time digital simulator (RTDS) sends out programmable current signals and through D/A converters and corresponding current signals are generated. A dSPACE controller measures current signals through A/D converters and executes the downloaded d-q and the proposed adaptive algorithms in real time simultaneously. Experimental data are collected through a data acquisition system in dSPACE controller. Both algorithms are digitalized first with sampling time of $50\mu s$ and time step of $50\mu s$. The test current is chosen as shown in Table 6-1. The parameters in brackets are used for load current change tests. And only phase A currents are displayed for comparison.

### 6.2.1.1 Start Transient

The estimated different frequency components are added together to form the estimated signal. The start transient results are shown in Figure 6-5, which clearly demonstrates that both methods can reach steady state within one fundamental cycle. Figure 6-5 (a) shows the input and estimated currents from both methods and the estimated currents match the reference input current very well in about one cycle. Careful look at different frequency components (fundamental and 13th harmonic) in Figure 6-5 (b) and Figure 6-5 (c), respectively, reveals that the proposed adaptive method has better convergence speed. It has significant convergence improvement at higher order harmonics (13th harmonic in our case as shown in Figure 6-5 (b)). A careful investigation the start transient comparison in

Table 6-2 demonstrates that the proposed adaptive method converges quicker than D-Q method and has a clear trend of reducing the difference between estimated and reference currents.
Table 6-1 RT-HIL test current parameters

<table>
<thead>
<tr>
<th>Harmonic</th>
<th>Fundamenta</th>
<th>5th</th>
<th>7th</th>
<th>11th</th>
<th>13th</th>
</tr>
</thead>
<tbody>
<tr>
<td>Magnitude (A)</td>
<td>5 (7)</td>
<td>4 (2)</td>
<td>3 (1.5)</td>
<td>2 (1)</td>
<td>1 (0.5)</td>
</tr>
<tr>
<td>Phase (rad)</td>
<td>3 (2)</td>
<td>0.5 (1)</td>
<td>1 (1.5)</td>
<td>1.5 (3)</td>
<td>2 (2.5)</td>
</tr>
</tbody>
</table>

Table 6-2 Comparison during start transient (13th harmonic)

<table>
<thead>
<tr>
<th>Sampling points (5e-5s)</th>
<th>Reference Current</th>
<th>D-Q Method</th>
<th>Percentage of Absolute Error</th>
<th>Adaptive method</th>
<th>Percentage of Absolute Error</th>
</tr>
</thead>
<tbody>
<tr>
<td>260</td>
<td>-0.49835</td>
<td>-0.82507</td>
<td>65.5%</td>
<td>-0.30941</td>
<td>37.9%</td>
</tr>
<tr>
<td>270</td>
<td>-0.17853</td>
<td>-0.15945</td>
<td>10.7%</td>
<td>-0.21437</td>
<td>20.0%</td>
</tr>
<tr>
<td>280</td>
<td>0.85297</td>
<td>0.92795</td>
<td>8.8%</td>
<td>0.79343</td>
<td>7.0%</td>
</tr>
<tr>
<td>290</td>
<td>-0.95856</td>
<td>-1.3442</td>
<td>40.2%</td>
<td>-0.89539</td>
<td>6.6%</td>
</tr>
<tr>
<td>300</td>
<td>0.85306</td>
<td>1.0659</td>
<td>25.0%</td>
<td>0.84436</td>
<td>1.1%</td>
</tr>
</tbody>
</table>

6.2.1.2 Load change Transient

Nonlinear load currents may change during operation and the proposed algorithm is also tested with different frequency component changes as shown in Figure 6-6. Both d-q and the proposed adaptive methods show good dynamic performance at tracking harmonic changes as shown in Figure 6-6. The proposed adaptive method still shows better convergence speed than the traditional d-q transformation based method by Figure 6-6 (b) and (c). Detailed information about the convergence speed can be found in Table 6-3, where the adaptive method shows smaller absolute error compared with d-q method in the same transient region. Both tests verify the results from simulation and the proposed adaptive method has simplicity at finding adaptive gains and also has great convergence speed.

Table 6-3 Comparison during load change transient (13th harmonic)

<table>
<thead>
<tr>
<th>Sampling points (5e-5s)</th>
<th>Reference Current</th>
<th>D-Q Method</th>
<th>Percentage of Absolute Error</th>
<th>Adaptive method</th>
<th>Percentage of Absolute Error</th>
</tr>
</thead>
<tbody>
<tr>
<td>1250</td>
<td>0.48553</td>
<td>0.3494</td>
<td>28.0%</td>
<td>0.43661</td>
<td>10.2%</td>
</tr>
<tr>
<td>1260</td>
<td>-0.13947</td>
<td>0.020673</td>
<td>85.2%</td>
<td>-0.115341</td>
<td>17.3%</td>
</tr>
<tr>
<td>1270</td>
<td>-0.1947</td>
<td>-0.39428</td>
<td>102.5%</td>
<td>-0.20173</td>
<td>3.6%</td>
</tr>
<tr>
<td>1280</td>
<td>0.50507</td>
<td>0.60343</td>
<td>19.5%</td>
<td>0.48711</td>
<td>3.5%</td>
</tr>
<tr>
<td>1290</td>
<td>-0.41412</td>
<td>-0.51094</td>
<td>23.4%</td>
<td>-0.42865</td>
<td>3.2%</td>
</tr>
</tbody>
</table>
Figure 6-5 RT-HIL start transient results
6.2.1.3 DC Current and Frequency Change Test

An extreme test is performed with current parameters in Table 6-4 together with a fundamental frequency changes from 60 Hz to 70 Hz. Figure 6-7(a) shows that input current can be well followed even when fundamental and harmonic currents parameters change simultaneously with a significant fundamental frequency change. Harmonic components responses are slower compared with previous results because of frequency change. The
frequency response is shown in Figure 6-7(b). If only fundamental frequency changes (5Hz for example) but current parameters don’t change, frequency response will be improved significantly as shown in Figure 6-7(c). Considering the fact that a normal power network’s frequency change is below 0.5Hz, the proposed algorithm is quicker in tracking the system fundamental frequency and further estimates corresponding harmonics in real time. Additionally, this result shows its significant ability to track significant frequency change, which is not applicable to common PLLs.

Figure 6-7 RT HIL tests results with frequency changes
Table 6-4 Test current parameters

<table>
<thead>
<tr>
<th></th>
<th>Fund.</th>
<th>5th</th>
<th>7th</th>
<th>11th</th>
<th>13th</th>
<th>dc</th>
</tr>
</thead>
<tbody>
<tr>
<td>Mag. (A)</td>
<td>4(6)</td>
<td>3 (2)</td>
<td>3 (1.5)</td>
<td>2 (1)</td>
<td>1 (0.5)</td>
<td>1 (0.5)</td>
</tr>
<tr>
<td>Phase (rad)</td>
<td>3 (2)</td>
<td>0.5 (1)</td>
<td>1 (1.5)</td>
<td>1.5 (3)</td>
<td>2 (2.5)</td>
<td></td>
</tr>
</tbody>
</table>

6.2.2 Electrical Experiment

To further test the proposed algorithm, two kinds of nonlinear loads, a diode rectifier front end drive and a thyristor rectifier front end drive, are used for tests. A simplified test system schematic is shown in Figure 6-8. DC drive controls dc motor in toque mode and ac drive controls IM work in speed mode. Measured inputs currents to IM drive $i_{d2}$ (diode rectifier front end) and DC drive $i_{t2}$ (thyristor rectifier front end) are used for testing both algorithms under steady states and load current changes transients.

A dSPACE controller runs the algorithms in real time and collects data to a desktop computer. Ac motor drives and dc motor drives have different nonlinear characteristics, therefore both loads are used for experimental tests. Also, the estimated different frequency components are added together to form the estimated current and to see whether it matches the input nonlinear load current well. With the results from RT-HIL tests, it is reasonable to make the comparison between estimated currents by using traditional d-q and the proposed adaptive algorithm. Sampling time and time step of the control algorithms are both set to 100μS.

Figure 6-8 Electrical experimental setup
6.2.2.1 Steady State

In this test, DC motor is controlled at its 20% torque ability to provide the load for the IM motor. Both input currents of ac drive and dc drive are used to verify the proposed adaptive and popular d-q method. Figure 6-9 shows the steady state results of estimating ac drive’s front end input current, phase A current only.

Both methods show satisfactory performances from experimental results in Figure 6-9 and Figure 6-10. Figure 6-9 shows the experimental results from ac drive. The combined estimated current matches the input load current well. Due to high frequency current harmonics, two curves have some difference. Figure 6-9 (b) shows spectrum analysis of input current and estimated current using both methods (1st, 5th, 7th, 11th, and 13th only). Both methods have good steady state performance. Figure 6-10 shows the experimental results of two methods with dc drive load. Spectrum analysis shown in Figure 6-10 (b) demonstrates that dc current deteriorates d-q method performance. It verifies the results found in simulation tests.

6.2.2.2 Start Transient

Start transient tests are also performed to see the dynamic performance of the proposed method. Only the dc drive’s results are shown in Figure 6-11 because its current is more complicated than ac drive current. This test is to show the results of adaptive algorithm with and without PSO tuning. The proposed PSO tuned adaptive algorithm has smaller time to reach steady state. This conclusion can be further verified by Figure 6-11 (b), where PSO tuned adaptive method has much faster dc offset current estimation. DC offset current estimation doesn’t affect high frequency components’ estimation, but it affects the estimated current (total)’s waveforms.

6.2.2.3 Load Change Transient

This test verifies the good performance of the proposed algorithm in load change transient. The dc motor torque reference is set to 15% first and then jumps to 30% with a 100% positive change. Figure 6-12 shows the load change transient results of both methods with ac drive and dc drive. Both methods have good dynamic performance as shown in RT HIL tests results. However, the proposed adaptive algorithm has better performance when input current has dc offset in DC drive. The estimated current by d-q method clearly has some error from input current. This matches previous HIL and experimental results.
(a) Input and estimated currents

(b) Spectrum analysis

Figure 6-9 Experimental results of AC drive current (Left: D-Q ; Right: Adaptive)
Figure 6-10 Experimental results of DC drive current (Left: D-Q; Right: Adaptive)

Figure 6-11 Experimental start transient results (DC drive)
The advantages of the proposed method are further verified by the experimental results such as easy and efficient tuning of adaptive algorithms for parameter identifications and enhanced steady state and dynamic performances for real time applications. The estimated fundamental and harmonic currents therefore can be reconfigured for specified harmonic cancellation or reactive power compensation applications.

6.2.2.4 Frequency Tracking Result

Figure 6-13 shows frequency estimation results of adaptive detection method. Fundamental frequency of 60 Hz is estimated well. Different frequency harmonics can be accurately estimated to form a matching estimated current. Input current in Ch1 and estimated current in Ch2 are
overlapped, which means the adaptive algorithm with frequency estimation function works well. Ch3 shows the estimated fundamental current waveform.

6.2.3 Summary

The proposed adaptive detection algorithm is found to be satisfactory when compared in real time application with the popular d-q method. However, the adaptive detection algorithm's performance is significantly better when input current has a dc offset. Adaptive detection methods also have some advantages over d-q methods such simple application to single phase system, phase independent characteristic and frequency estimation. For unbalanced system or frequency changing systems, the proposed adaptive detection algorithm is simpler and produces better performance. The summary of the comparison between the proposed adaptive detection algorithm and the traditional d-q method is shown in Table 6-5.

Figure 6-13 Frequency tracking: 10ms/div; 5V/div; Ch1: Input current 1V/A; Ch2: Estimated current 1V/A; Ch3: estimated fundamental current 1V/A; Ch4: Estimated fundamental frequency 100mV/Hz
<table>
<thead>
<tr>
<th></th>
<th>Adaptive method</th>
<th>d-q method</th>
</tr>
</thead>
<tbody>
<tr>
<td>Start Transient</td>
<td>Fast</td>
<td>Faster</td>
</tr>
<tr>
<td>Load change transient</td>
<td>Fast</td>
<td>Fast</td>
</tr>
<tr>
<td>DC component estimation</td>
<td>Good</td>
<td>N/A</td>
</tr>
<tr>
<td>Three phase system</td>
<td>Easy</td>
<td>Easy</td>
</tr>
<tr>
<td>Frequency estimation</td>
<td>Included</td>
<td>PLLs</td>
</tr>
<tr>
<td>Flexibility</td>
<td>High</td>
<td>Limited</td>
</tr>
<tr>
<td>Computational burden</td>
<td>Small</td>
<td>Small</td>
</tr>
<tr>
<td>Three phase unbalanced systems</td>
<td>Phase independent</td>
<td>Complicated</td>
</tr>
</tbody>
</table>
7 CONCLUSION AND FUTURE WORK

In this dissertation, a new adaptive harmonic selective detection method is proposed for power quality improvement. In theory, this adaptive method has better convergence properties compared to previous adaptive neuron detection method. The adaptive gains of this method can be relatively large to obtain faster convergence. The stability of the proposed method is guaranteed and its property is analyzed when compared to popular d-q method. In summary, the proposed adaptive method achieves the following contributions for power quality improvement applications. The proposed new adaptive harmonic detection algorithm has advantages such as:

1. Better convergence properties, which makes harmonic selective detection have better dynamic and steady state performance;
2. Solution of stability problems associated with adaptive detection algorithms;
3. Improved flexibility and easy applications to future multiple coordination control with more active filtering devices.
4. Easy applications to three-phase systems;
5. Fast frequency tracking ability when power system has big frequency change;
6. Works well when system currents have dc offsets.

Future work will be focused on the following areas:

1. Advanced optimization techniques for fast and easy tuning of adaptive gains when the input current’s model does not satisfies PE condition.
2. Fully electrical implementation and test of the proposed adaptive algorithm by applying it to an active filter application for harmonic selective cancellation.
3. Comprehensive experimental comparison of the proposed adaptive algorithm and d-q algorithm in active filter based power quality improvement applications.
APPENDIX A

PSO TUNING OF THE PROPOSED ADAPTIVE DETECTION ALGORITHM

Although the new adaptive detection algorithm proposed in previous section has the advantage of choosing arbitrary large gains for fast convergence speed in theory. In real applications, due to noise, dc components, higher order harmonics’ interference and errors from current modeling, the gains may not be chosen arbitrarily. Therefore, this section proposes the application of PSO (Particle Swarm Optimization) to adaptive algorithms’ gain selection, minimizing the time consuming problem of tuning adaptive gains. As an efficient heuristic algorithm, PSO can generate high quality solutions for nonlinear optimization problems. Therefore, by combining the adaptive algorithm and PSO, the time consuming gains selection process is shortened and the performance of the stability and dynamic speed are enhanced compared with completely hand tuned adaptive algorithm. Minor hand tuning may still be required in the real application but with much less effort.

A.1 Particle Swarm Optimization

Particle Swarm Optimization (PSO) is a population based stochastic search algorithm. It was first introduced by Kennedy and Eberhart in 1995 and it has been widely used to solve a broad range of optimization problems. The algorithm was presented as simulating animals’ social activities, e.g. insects, birds, etc. It attempts to mimic the natural process of group communication to share individual knowledge when such swarms flock, migrate, or hunt. If one member sees a desirable path to go, the rest of this swarm will follow quickly. In PSO, this behavior of animals is imitated by particles with certain positions and velocities in a searching space, wherein the population is called a swarm, and each member of the swarm is called a particle. The decision process of PSO is similar to human beings learning process. People use two important kinds of information in the decision process, where the first one is their own experience and the second one is other people’s experiences [69] [70] [71] [72] [73].

According to the background of the above section, the PSO implementation is as follows: Starting with a randomly initialized population, each particle in PSO ‘flies’ through the search space, while remembering the best position it has seen. Members of particles of a swarm
communicate good positions to each other and dynamically adjust their own position and velocity based on these good positions. The velocity adjustment is based upon the historical behaviors of the particles themselves as well as their companions’. In this way, the particles tend to fly towards better and better searching areas over the searching process. The searching procedure based on this concept can be described by the following two equations:

\[ v_i^{k+1} = w v_i^k + c_1 \cdot rand_1 \cdot (pbest_i - x_i^k) + c_2 \cdot rand_2 \cdot (gbest - x_i^k) \]  \hspace{1cm} (A-1)

\[ x_i^{k+1} = x_i^k + v_i^{k+1} \]  \hspace{1cm} (A-2)

Where the parameters in (A-1) and (A-2) are defined as:

- \( c_1, c_2 \): acceleration coefficients;
- \( w \): the inertia weight factor;
- \( rand \): random functions in the range of \([0,1]\);
- \( x_i \): the \( i \)th particle’s position;
- \( k \): \( k \)th iteration;
- \( pbest_i \): the best previous position of \( x_i \);
- \( gbest \): the best particle among the entire population;
- \( v_i \): the velocity for particle \( x_i \).

From equation (A-1), velocity updates in three parts, i.e. the momentum part, the cognitive part, and the social part. Every particle’s current position is then evolved in the solution space, which produces a better next position in the solution space. The concept of the search process can be further explained by Figure A-1. The position of each particle is represented by an XY axis position. The velocity is represented by the velocities of X axis and Y axis. Modification of the particle position is realized by the position and velocity information according to equation (A-1) and equation (A-2).

As introduced in previous section, the adaptive gains in (4-20) and (4-31) are chosen by experience using trial and error method. When more harmonics are required for estimation, more adaptive gains are required in the algorithm. The true relationship with gains and performance is not clear when one looks at the algorithm. Therefore, the next section proposes the PSO tuning of the proposed adaptive method. This PSO tuning is not a part in the proposed adaptive method.
and may be applied to most adaptive methods of this type to solve the existing problems associated with traditional adaptive identification method.

**A.2 Proposed PSO Tuned Adaptive Method**

To apply PSO method to adaptive gains tuning, all the adaptive gains are grouped to formulate the $x$ parameter. Therefore, for an estimate up to $n$th harmonic, $x$ has dimension $2n + 3$ (two for each frequency component, one for the frequency estimation and the last two for dc component) or $2n + 4$ (the last one for a forgetting factor if used). For fast convergence purpose, ITSE (integral of time squared error) is used as evaluation function as shown in equation (7-3) where $r(t)$ is the input and $y(t)$ is the output.

$$ITSE = \int_{0}^{\tau} \tau [r(\tau) - y(\tau)]^2 d\tau = \int_{0}^{\tau} \tau e^2(\tau) d\tau \quad (7-3)$$

The PSO searching method to identify the adaptive harmonic detection parameters is shown in the flowchart Figure A-2. There are basically three main tasks of the method. The first task is the initialization of the optimization with the lower and upper bounds of the adaptive gains, the particle number and problem dimension, the algorithm’s time step and time window. The position and velocity vectors associated with each particle is initialized by setting random values or the values from last run and saved data.

![Figure A-1 Concept of PSO searching process](image-url)
After initialization, the PSO algorithm is running to calculate and evaluate values via the objective function shown in equation (7-3). Based on the fitness of each particle, $p_{best}$ and $g_{best}$ are stored, and velocity and position of each particle are updated. The entire process is repeated until either the iteration limit is reached or preset criteria are satisfied. Finally, the PSO outputs objective satisfying adaptive gains for adaptive parameter identifier. The PSO tuned adaptive harmonic selection method may then applied for harmonic selection in harmonic cancellation.

![Figure A-2 Flowchart of proposed PSO tuning method](image-url)
A.3 Simulation of PSO Tuned Adaptive Detection Algorithm

As the number of harmonic currents increase, so do the number of gains. This leads to significant difficulty in tuning adaptive algorithms in general. For the experimental phase of the work, PSO was used to tune the adaptive algorithm to be implemented in hardware. To verify the proposed PSO tuned adaptive harmonic detection algorithm, a synthesized current signal with fundamental to 13th harmonic was used for tests. The test current signal parameters are shown in Table A-1. The simulation is running in Matlab/Simulink with 100μs time step using a discrete time solver. The test current was through both a hand tuned adaptive and PSO tuned adaptive identifiers in the same simulation and the results are shown in Figure A-3. Figure A-3 (a) shows the estimation errors of both algorithms. Although the original adaptive algorithm is fast, the PSO tuned adaptive algorithm is significantly faster. PSO tuned adaptive algorithm has slightly better performance at fundamental frequency and significantly better performance at the 13th harmonic as can be seen in Figure A-3 (b) and (c), respectively. PSO tuned adaptive algorithm can reach steady state well below one cycle. This is not to say optimized tuning will always perform better than hand tuned algorithm, but the tuning process quicker and less tedious.

The adaptive gains of both methods are shown in Table A-2, where 10 adaptive gains are tuned to have an improved performance based on the objective function. Figure A-4 shows the PSO optimization convergence with respect to iteration process. The ITSE is reduced during each iteration process and reaches its minimum value after 10 iterations. The hand tuned adaptive identifier took authors tens of hours to tune. The PSO tuned adaptive algorithm took the authors several seconds to find satisfactory performance.

<table>
<thead>
<tr>
<th>Table A-1 Test current parameters</th>
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<tbody>
<tr>
<td><strong>Magnitude (A)</strong></td>
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<tr>
<td>-------------------</td>
</tr>
<tr>
<td>5</td>
</tr>
<tr>
<td>7</td>
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<table>
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<tr>
<th>Table A-2 Adaptive gains parameters</th>
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<tr>
<td><strong>Gains</strong></td>
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<tr>
<td>-----------</td>
</tr>
<tr>
<td>Original</td>
</tr>
<tr>
<td>Optimized</td>
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Figure A-3 Simulation comparison of adaptive algorithm and PSO tuned adaptive algorithm
Figure A-4 Optimization process
APPENDIX B

POWER QUALITY IMPROVEMENT APPLICATION

This subsection introduces the hardware in the loop experiment for a power quality improvement application by using the proposed adaptive harmonic selective detection algorithm. Figure B-5 shows the controller HIL setup of this test. The proposed new detection algorithm, dc voltage control algorithm and current control algorithm are running in dSAPCE controller in real time. More details about those algorithms can be found in previous sections. The dSPACE controller controls a virtual electrical system running in real time in RTDS for the controller-HIL tests.

A desk computer running Matlab and RSCAD is working as a user interface for algorithm development, algorithm downloading, virtual system downloading and data acquisition. The currents $i_s$, $i_L$ and $i_{inv}$ are the source current, load current and active filter output current respectively. The load current is detected by this proposed PSO tuned adaptive method and the harmonic currents information is extracted by the proposed method. The current $i_{inv}$ is measured and feedback for current tracking purpose with a PI+Resonant controller [76] [77].

The switching frequency $f_s = 5000Hz$ and the time step used in the controller is $100 \mu s$. The reference voltage of the dc link $u_{dc}^*$ is 800V. Other parameters in this test are shown in Table B-3. The HIL test has three steps. The first step is no control of the converter and by the functionality of the diodes in the IGBT bridge, the dc link capacitor’s voltage is rectified to about 650V. The second step is to boost the dc link voltage to about 800V by applying the dc voltage control and current control module. After the dc link voltage settles at about 800V, harmonic cancellation takes place.

<table>
<thead>
<tr>
<th>Table B-3 System parameters</th>
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<tr>
<td>$V_s$ (L-L)</td>
</tr>
<tr>
<td>480V</td>
</tr>
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</table>
Figure B-5 Controller HIL experimental setup for power quality application
B.1 Single Harmonic Cancellation

As seen in Figure B-6, the first test is about single harmonic current selective cancellation. Only the 5\textsuperscript{th} harmonic cancellation is applied to verify the proposed PSO tuned adaptive detection algorithm. Figure B-6 (a) shows the source current before and after the 5\textsuperscript{th} harmonic current cancellation is applied. The source current is significantly corrupted because of the nonlinear load before harmonic cancellation. Spectrum analysis of the distorted source current in Figure B-7 (a) shows that significant 5\textsuperscript{th} and 7\textsuperscript{th} harmonics and some higher order harmonics existing in the source current. The total harmonic distortion (THD) factor is 33.92\%, which is very high.

The dc link voltage control first takes effect from the sampling points about 800 and some big fundamental current is generating for charging purpose. The source current from about sampling points 1500 shows that the waveform is restored to match sinusoidal waveform after 5\textsuperscript{th} harmonic cancellation takes place. Spectrum analysis of the source current with 5\textsuperscript{th} harmonic cancellation in Figure B-7 (b) clearly demonstrates that 5\textsuperscript{th} harmonic is greatly attenuated and fundamental and other harmonics are almost unchanged. The THD is greatly reduced to 19.75 \%. The output current of the converter is shown in Figure B-6 (c) and significant current with frequency of 300 Hz flow into the power grid for harmonic cancellation. Some fundamental current is used for keeping the dc bus voltage constant at reference voltage. The nonlinear load current is shown in Figure B-6 (b), which is not changed during the whole application process. The DC link voltage response is shown in Figure B-6 (d), where the charging process is clearly demonstrated and the dc link voltage is well maintained during harmonic cancellation process.
Figure B-6 HIL results of single harmonic cancellation
As seen in Figure B-8, the second simulation presents multiple harmonic current selective control with 5\textsuperscript{th} and 7\textsuperscript{th} harmonic cancellation. The 5\textsuperscript{th} and 7\textsuperscript{th} harmonic current references are added as the control reference. The source current has the heavily distorted waveform as shown Figure B-8 (a) before multiple harmonic cancellation. Spectrum analysis of the source current in Figure B-9 (a) shows that significant 5\textsuperscript{th}, 7\textsuperscript{th} harmonics together with some 11\textsuperscript{th} and 13\textsuperscript{th} harmonic exist in the source current. Figure B-8 (a) also shows the source current of the whole test, where a charging process with big fundamental current is used for dc capacitor charge. The harmonic cancellation takes place from about point 1500 and the source current’s shape is greatly altered to match a sinusoidal waveform. Spectrum analysis of the source current in Figure B-9 (b) clearly shows that 5\textsuperscript{th}, 7\textsuperscript{th} harmonics are greatly reduced although they are not zeros. The THD of the source current is further reduced to 13.65% compared with 19.75% with only 5\textsuperscript{th} harmonic cancellation.

Figure B-8 (b) shows the output current of the inverter, which means compensating 5\textsuperscript{th} and 7\textsuperscript{th} harmonics are sent to the power grid for harmonic current cancellation. The fundamental current in these output currents is for the dc link capacitor voltage maintenance. The dc link voltage is also well controlled near 800V as shown in Figure B-8 (d). Both experimental tests verify that this power quality improvement application with the proposed harmonic detection algorithm works well. The reason that 7\textsuperscript{th} harmonic is not reduced to zero is due the limitation of the current controller and bandwidth of the controller. The controller’s sampling time also limits the test with higher order harmonic cancellation.
Figure B-8 HIL results of multiple harmonic cancellation
(a) Before 5\textsuperscript{th} and 7\textsuperscript{th} harmonic cancellation

(b) After 5\textsuperscript{th} and 7\textsuperscript{th} harmonic cancellation

Figure B-9 Spectrums of source currents
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BIOGRAPHICAL SKETCH

Lewei Qian received his B.S. and M.S. degree in Electrical Engineering from Hefei University of Technology, China, in 2000 and 2003 respectively. He is currently enrolled in the doctoral program in Mechanical Engineering department of Florida State University and working as a research assistant at the Center for Advanced Power Systems, Tallahassee, Florida, USA. His research interests are active filter control, reconfigurable power conversion and real-time digital simulations. He worked as a research engineer at Caterpillar Inc. from May 2006 to Aug. 2006 and worked as a product engineer at Siemens Energy and Automation, Inc. from May 2007 to Aug.2007.
PUBLICATION LIST


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