CHAPTER 4

FAULT DETECTION IN ELECTRICAL SYSTEMS
USING ESTIMATORS

4.1 INTRODUCTION

This chapter discusses on the use of estimators like the Adaptive Peak Filter (APF), Enhanced Phase-Locked Loop (EPLL) estimator used to detect unknown signal parameters like amplitude, frequency, phase of a measurable time varying signal $x(t)$, which is an active area of research interest.

The DSP techniques like STFT is suitable for tracing signal parameters of a time varying signal, though the limitation is that it cannot identify both the low frequency and the high frequency nonstationarity in signals simultaneously. If the time varying signal has multiple frequency components say, at $kw_1$, $(k+1)w_1$, (where $k=1, 1.2, 1.8...$) as harmonic frequency components it is difficult to estimate the relative shift in these harmonics using a unified low pass filter in the spectrogram algorithms. This requires newer algorithms that estimate individual unrelated frequency components in the signal independent of each other.

The estimators based on APF, EPLL are specially applicable to trace variations in the signal parameters since they are constructed based on the combination of broad and narrow bandpass filters which are of lesser
complexity and the implementation of the estimator algorithms are easily programmable in advanced microprocessors. These extract parametrical variation in the signal by using these parameters itself as the tunable/adaptive quantities in the realization of the estimator blocks (Ghartemani & Iravani (2003), Mojiri & Bakhshai (2007), Mojiri et al (2007), Tarek et al (2007), Hayes & Treichler (2008)).

Electrical disturbances in voltage and current signals are common in malfunctioning equipments of the power grid either in generation or transmission or distribution side of the grid whenever the equipments herein are subject to abnormal supply or load variations while in operations. The electrical disturbances need to be early detected and mitigated before it propagates as a grid disturbance or as an equipment fault.

The earlier chapters discussed on fault diagnosis in the static electric machine like transformer and in the rotating machine like three phase induction motors. This chapter discusses on the comparative performance of the estimators for detecting parameterical variations in the voltage and current signals obtained from electrical equipments. The inquisitive approach on decision for optimal choice of the estimator for fault diagnosis is based on the fact that each of the above mentioned estimators are of unique characteristics because of which each one of them seems optimal selection for certain applications though not for another application.

APF can extract harmonic signals, estimate amplitude, frequency and phase of those signals. APF has been used for winding fault detection in electrical equipments like motors and transformers. APF has been used to analyze Single Phase Matrix Converter (SPMC) as a cycloconverter, which is used to control a variable speed drive.
4.2 COMB FILTERS

Comb filter is a notch filter in which the nulls occur at particular frequency and its multiples. It has multiple pass bands and multiple stop bands. The peaks and dips look like the teeth of a comb, with very narrow, deep notches where signals are attenuated (Gayakwad 1995, LabVIEW manual).

Comb filter is mainly classified into two types namely peak and notch. One type has notches or peaks at the frequencies $kf_s/N$, while the other type has notches or peaks at the frequencies $(k+1/2)f_s/N$, where $k$ is an integer in the range $[0, N-1]$, $N$ is the order of the filter and $f_s$ is the sampling frequency in Hz. The properties of the notch filter are infinite attenuation at notch frequencies, unity gain at all other frequencies and no effect on the phase of the signal. The peak type filter extracts (i.e., lets pass through) one specific sinusoidal component and rejects all other components including noise.

The control parameter involved in the design of comb filter are the center frequency of the notch or peak $f_o$, a level on the magnitude squared response of the filter $A_b$ in deciBels, the frequency bandwidth $\Delta f$ at $A_b$ and the sampling frequency $f_s$ is shown in Figure 4.1.

Figure 4.1 Control parameters of comb filter
4.2.1 Notch Filter

There are two types of notch filters namely notch type-1 and notch type-2. One type of notch filter has notches at the frequencies $k f_s/N$, while the other type of notch filter has notches at the frequencies $(k+1/2)f_s/N$, where $k$ is an integer in the range $[0, N-1]$. The magnitude response of the notch filter is shown in Figure 4.2(a).

The transfer function of the notch type-1 filter is given in Equation (4.1)

$$H(z) = \frac{b(1 - z^{-N})}{1 - a z^{-N}}$$

The transfer function of the notch type-2 filter is given in Equation (4.2)

$$H(z) = \frac{b(1 + z^{-N})}{1 + a z^{-N}}$$

$a$ and $b$ are the filter coefficients.

The filters magnitude response for notch type-1 and notch type-2 with the cut off frequency ($f_o$) of 50Hz and sampling frequency ($f_s$) of 1000Hz is shown in Figure 4.2(b) and Figure 4.2(c). The type-1 is used to suppress both even and odd harmonics and the type-2 filter is for odd harmonics only.
4.2.2 Peak Filter

Similarly there are two types of peak filters namely peak type-1 and peak type-2. One type of peak filter has peaks at the frequencies $k f_s / N$, while the other type of peak filter has peaks at the frequencies $(k+1/2)f_s / N$, where $k$ is an integer in the range $[0, N-1]$. The magnitude response of the peak filter is shown in Figure 4.3(a).

The transfer function of the peak type-1 filter is given Equation (4.3)

$$H(z) = \frac{b(1 + z^{-N})}{1 - a z^{-N}}$$

(4.3)
The transfer function of the peak type-2 filter is given in Equation (4.4)

\[ H(z) = \frac{b(1 - z^{-N})}{1 + az^{-N}} \]  

The filters magnitude response for peak type-1 and type-2 with the cut off frequency \((f_o)\) of 50Hz and sampling frequency \((f_s)\) of 1000Hz are shown in Figure 4.3(b) and Figure 4.3(c). The type-1 peak filter passes both even and odd harmonics whereas the type-2 peak filter passes only the odd harmonics.

Figure 4.3 (a) Representation of peak filters (b) Magnitude response of type-1 peak filter (c) Magnitude response of type-2 peak filter
4.3 ADAPTIVE PEAK FILTER

An ideal peak filter is a linear filter whose frequency response is characterized by a unit gain at a particular frequency and zeros at all other frequencies. A peak filter is able to extract the desired sinusoidal component of a given signal provided that the signal frequency remains constant. When the signal is nonstationary, i.e., its frequency varies with time, the peak filter cannot achieve this task. Using a high-order peak filter will not accomplish this goal because the fundamental frequency is not constant. A filter that actively tracks the changing amplitude, phase and frequency is needed to extract the fundamental from the transient. The dynamics of the algorithm presents a peak filter in the sense that it extracts (i.e., lets pass through) one specific sinusoidal component and rejects all other components including noise. It is adaptive in the sense that the peak filter accommodates variations of the characteristics of the desired output.

Regalia (1991) has developed a lattice-based adaptive IIR notch filter which features independent tuning of the notch frequency and bandwidth attenuation.

Joshi & Roy (1999) have presented a new method for the design of multiple notch filters using an all-pass filter of order 2N, N being the number of notch frequencies.

Ghartemani & Iravani (2002) present various applications of a non-linear adaptive notch filter which operate based on the concept of an Enhanced Phase-Locked Loop (EPLL). Applications of the filter for online signal analysis for power systems protection, control and PQ enhancement are presented. The proposed scheme can be applied for signal analysis both under stationary and non-stationary conditions.
Carvalho et al (2009) described a Phase-Locked Loop (PLL)-based power systems harmonic estimation algorithm, which uses an analysis filter bank and multirate processing. The filter bank is composed of bandpass filters. The initial center frequency of each filter is purposely chosen to be equal to harmonic frequencies. An adaptation strategy makes it possible to track time-varying frequencies as well as interharmonic components.


Chen & Chang (2010) has reviewed several commonly used methods namely Fast Fourier Transform (FFT), Estimation of Signal Parameters via Rotational Invariance Technique (ESPRIT), Multi Signal Classification (MUSIC), Prony, Adaptive Linear Element (ADALINE), Kalman filtering, PLL and Artificial Neural Network (ANN) for time varying harmonic and interharmonic detection of measured voltage and current waveforms and implemented in an integrated virtual instrumentation.

Yazdani et al (2010) introduced a new three-phase adaptive notch filtering approach for the extraction of harmonic and reactive current components for use in grid-connected converters. The main function of this method is to provide synchronized harmonic and reactive current components for control purposes. This method is capable of detecting a selective order of harmonics. This feature is useful where elimination of certain harmonics is of concern. The methodology is applicable for a wide range of equipment like regenerative converters, distributed generation system, as a basis for detection of the reference signals.
Sohn et al (2012) proposed a method to decompose the input power signal using a filter bank system that is a modular binary tree structure with the fundamental filter banks arranged successively in each stage. The fundamental filter bank is designed to separate the odd and even harmonics to reduce spectral leakage. An adaptive filter is used to improve the accuracy of parameter estimation for each decomposed harmonic.

Ghartemani et al (2012) introduced the concept of in-loop filters and window functions into PLL, which enables smoother estimation of the signal parameters such as phase angle, frequency, and amplitude in the presence of noise and harmonics with selective frequency removal.

4.3.1 Processing of Harmonics and Interharmonics using Adaptive Peak Filter

Let \( r(t) \) denotes a measured (or computed) signal as discussed in Mojiri et al (2010) is given in Equation (4.5).

\[
 r(t) = A_k \sin \phi_k(t) + \sum_{k=2}^{n} A_k \sin \phi_k(t) + n(t)
\]

where \( \phi_k(t) = \omega_k t + \phi_k \)

is the total phase angle of the \( k^{th} \) constituting component called \( r_k(t) \), and \( n(t) \) denotes the total noise imposed on the signal. The first component called the fundamental component is deliberately signaled out in Equation (4.5) due to its significance in power signals. The signals attribute \( A_k, \omega_k \) and \( \phi_k \) are magnitudes, frequencies and phase angles of the constituting components and can be time-varying. The noise \( n(t) \) is assumed as a zero-mean white Gaussian noise with a variance of \( \sigma^2 \). The \( k^{th} \) component is primarily specified by its amplitude \( A_k \) and its frequency \( \omega_k \) but its instantaneous value \( A_k \sin \phi_k(t) \) can also be of interest in some applications. The APF technique is a structure
composed of n-coupled parallel filters so that each one extracts a component of the power signal and estimates its frequency. The dynamic behavior of the algorithm is characterized by the following set of differential equations as in Equation (4.6).

\[
\begin{align*}
\ddot{x}_k(t) &= -\theta_k^2(t)x_k(t) + 2\zeta_k \dot{\theta}_k(t)e(t) \\
\dot{\theta}_k(t) &= -\gamma_k x_k(t)\theta_k(t)e(t), \quad k = 1, \ldots, n \\
e(t) &= r(t) - \sum_{k=1}^{n} \dot{x}_k(t)
\end{align*}
\]  

(4.6)

where the dot sign is used to denote the time derivative. In Equation (4.6), \(\theta_k\) is the estimated frequency (in radians per second) of the \(k\)th component, \(\zeta_k\) and \(\gamma_k\) are the real positive numbers which are the design parameters of the \(k\)th subfilter and it determines its behavior in terms of accuracy (steady state) and convergence speed (transient). The normalized \(\gamma_k\) is used to remove the dependency of the averaged system to amplitude as given below as in Equation (4.7).

\[
\gamma_k = \frac{\gamma_0}{\mu + N(x_k^2 \theta_k^2 + \dot{x}_k^2)}
\]

(4.7)

where \(\gamma_0 = \alpha N\), \(\zeta_k = \zeta_0 / k^2\), \(\mu\) and \(N\) are real positive constants. The constant \(\mu\) is a small positive constant to prevent the denominator of \(\gamma_k\) from becoming zero. The two design parameters \(\alpha\) and \(\zeta_0\) determines the dynamical behavior of the whole algorithm in terms of the transient and steady-state responses.

The dynamic system Equation (4.6) for the input signal Equation (4.5) with \(\phi_k(t) = \omega_k t + \phi_k\) and \(n(t) = 0\) has a unique quasi periodic orbit given by as in Equation (4.8).

\[
P(t) = (P_1^T(t) \ldots P_n^T(t))^T
\]

(4.8)
where \( P_k(t) \) is given by Equation (4.9)

\[
P_k(t) = \begin{pmatrix}
\bar{x}_k \\
\tilde{x}_k \\
\bar{\theta}_k
\end{pmatrix} = \begin{pmatrix}
-\frac{A_k \cos \phi_k(t)}{\omega_k} \\
A_k \sin \phi_k(t) \\
\omega_k
\end{pmatrix}
\]

\( k = 1, ..., n. \) \hspace{1cm} (4.9)

Equation (4.9) represents the \( k \)th constituting component of the input signal as well as its frequency are directly provided by the \( k \)th set of the differential equations, and hence a full set of decomposition is achieved, if a sufficient number of filters are employed. The structural block diagram of the algorithm is shown in Figure 4.4(a) in which the detailed implementation block diagram as discussed in Mojiri et al (2010) of the \( k \)th subfilter is shown in Figure 4.4(b). The error signal \( e(t) \) is applied to each subfilter, and the \( \theta_i \) update law of Equation (4.6) is employed to force the error signal to zero. This enables each subfilter to focus on a component and extract it. Each subfilter adapts for a band of frequencies around the \( f_{\text{cutoff}} \). The bandwidth is adjustable using frequency limiter in Figure 4.4(b). The initial value of the integrator of the frequency update law in Equation (4.6) is set to \( \theta_{i0} \), which is equal to the nominal value of the frequency of the \( k \)th component to be extracted.

The amplitude and the phase of \( k \)th component of the input signal is given by Equation (4.10).

\[
\left( \bar{\theta}_k \bar{x}_k + \tilde{x}_k \right)^{1/2} = A_k
\]

\[
\phi_k(t) = \begin{cases} 
\arccos \left( \frac{-\theta_k x_k}{A_k} \right) ; \dot{x}_k > 0 \\
2\pi - \arccos \left( \frac{-\theta_k x_k}{A_k} \right) ; \dot{x}_k < 0
\end{cases}
\]

\hspace{1cm} (4.10)
Thus the APF estimates harmonics and interharmonics.

Figure 4.4  (a) Block diagram representation of the frequency estimator  
(b) The detailed representation of the $k^{th}$ paralleled sub filter

Steps for realizing the APF

1. Real time or synthesized signal  
2. Applied into APF realized using Equation (4.6)-(4.10)  
   
   \[
   \dot{x}_k(t) = -\theta_k^2(t)x_k(t) + 2\zeta_k \dot{x}_k(t)e(t) \\
   \dot{\theta}_k(t) = -\gamma_k x_k(t)\theta_k(t)e(t), \quad k = 1, \ldots, n \\
   e(t) = r(t) - \sum_{k=1}^{n} \dot{x}_k(t) 
   \]
\[ \gamma_k = \frac{\gamma_0}{\mu + N(x_k^2 + \hat{x}_k^2)} \]

\[ \gamma_0 = \alpha N, \quad \zeta_k = \frac{\zeta_0}{k^2}, \mu \quad \text{and} \quad N \text{ are real positive constants}, \quad k = 1, 2, \ldots, 5 \text{ is the harmonic order or number of sub filters used.} \]

\[ \left( \frac{\theta_k x_k^2}{\mu + N(x_k^2 + \hat{x}_k^2)} \right)^{1/2} = A_k \]

\[ \phi_k(t) = \begin{cases} \arccos \left( -\frac{\theta_k x_k}{A_k} \right); & x_k > 0 \\ 2\pi - \arccos \left( -\frac{\theta_k x_k}{A_k} \right); & x_k < 0 \end{cases} \]

\[ \zeta_0 = 1.25, \alpha = 5500, \mu = 0.01 \text{ and } N = 2500 \]

3. APF estimates harmonic signal \( A_k \sin \phi_k(t) \), amplitude \( A_k \), frequency \( \omega_k \), phase \( \phi_k \) filter through low pass filter \( 1/(\tau s + 1) \) where \( \tau = 1/(2\pi 10) \) s.

The case analysis is here executed with MATLAB/Simulink wherein the filter blocks are modeled.

### 4.3.1.1 Performance analysis of APF

The input signal \( x(n) \) as in Equation (4.11) consists of five harmonics which is given to APF which has five sub-units. A set of values of \( \zeta_0 = 1.25, \alpha = 5500, \mu = 0.01 \text{ and } N = 2500 \) are chosen as the basis for all simulations. The frequency limiters are adjusted to operate within the range of \( \pm 25 \text{Hz} \) around the nominal frequency of each harmonic. The included posterior filters are first-order, low-pass filters with the transfer function of \( 1/\tau s + 1 \) where \( \tau = 1/(2\pi 10) \) s. The amplitude and frequency of the test input signal chosen here as in Equation (4.11) is \( A_1 = 1 \text{p.u}, \quad A_2 = 0.8 \text{p.u}, \quad A_3 = 0.6 \text{p.u}, \quad A_4 = 0.4 \text{p.u}, \quad A_5 = 0.2 \text{p.u} \) and \( F_1 = 50 \text{Hz}, \quad F_2 = 100 \text{Hz}, \quad F_3 = 150 \text{Hz}, \quad F_4 = 200 \text{Hz}, \quad F_5 = 250 \text{Hz} \). The Figure 4.5(a) shows the extracted individual harmonic components. The amplitude and frequency estimation of harmonic components through the APF estimator are shown in Figure 4.5(b) and Figure 4.5(c).
\[ x(n) = \sin(2\pi 50t) + 0.8\sin(2\pi 100t) + 0.6\sin(2\pi 150t) + \\
0.4\sin(2\pi 200t) + 0.2\sin(2\pi 250t) \]  

Figure 4.5  (a) Harmonic signal extraction (b) Amplitude estimation through APF
Figure 4.5 (c) Frequency estimation (d) Phase estimation through APF
The phase angle estimation through APF is shown in Figure 4.5(d). The order of the harmonic component can be identified from the phase characteristics. For example first harmonic has only one notch for a period and the third has three notches for the same specified period and so on.

4.3.1.2 Estimation of interfrequency components

The input signal $x(n)$ as in Equation (4.12) consists of frequencies that are not interharmonics are given to APF. The amplitude and frequency of the input signal is $A_1=1\text{p.u}$, $A_2=0.8\text{p.u}$, $A_3=0.6\text{pu}$, $A_4=0.4\text{p.u}$, $A_5=0.2\text{p.u}$ and $F_1=70\text{Hz}$, $F_2=100\text{Hz}$, $F_3=135\text{Hz}$, $F_4=210\text{Hz}$, $F_5=250\text{Hz}$.

$$x(n) = \sin(2\pi 70t) + 0.8\sin(2\pi 100t) + 0.6\sin(2\pi 135t) + 0.4\sin(2\pi 210t) + 0.2\sin(2\pi 250t)$$

(4.12)

![Figure 4.6 Frequency estimation through APF](image)

Each APF unit has a specific frequency band, so that each unit will extract the particular range of frequency present in the signal. Thus any interfrequency component present within the frequency band will be extracted by that particular unit as shown in Figure 4.6. The interfrequency component 60Hz, 135Hz and 210Hz are extracted by first, third and fourth unit of APF.
4.3.1.3 Frequency tracking feature

The frequency tracking of APF is tested through the following input signal. The input signal \( x(n) \) as in Equation (4.13) has frequency of \( F_1=50\text{Hz}, F_2=100\text{Hz}, F_3=150\text{Hz} \) till 1.2sec; thereafter the frequencies changes are \( F_1=65\text{Hz}, F_2=110\text{Hz}, F_3=160\text{Hz} \) which have been tracked through APF as shown in Figure 4.7.

\[
t = (0 : 0.001 : 1.2) \\
t_i = (1.201 : 0.001 : 5) \\
x(n) = \sin(2\pi 50t) + 0.8\sin(2\pi 100t) + 0.6\sin(2\pi 150t) + \sin(2\pi 65t_i) + 0.8\sin(2\pi 110t_i) + 0.6\sin(2\pi 160t_i) 
\]

(4.13)

![Figure 4.7 Frequency tracking through APF](image)

4.3.1.4 Noise elimination

The advantage of APF method is that it can be implemented in a DSP processor with lower computation complexity. The added advantage in using APF filter based feature extraction in signals is that APF has the capability to isolate noise out of signal due to efficient filtering.
The input signal \( y = \sin(20) + \text{WGN (Weight Gaussian Noise)} \), is given to APF with variance \( \sigma^2 = 0.5 \), and switching frequency \( F_s = 10\text{kHz} \) has been passed through APF which successfully extracts a single sinusoid as in Figure 4.8.

![Figure 4.8 Noise elimination through APF](image)

Another method for harmonic extraction with amplitude, frequency and phase estimation using Phase-Locked Loop (PLL) has been discussed for comparative study in the following section. The concept of the Phase-Locked Loop (PLL) is very much similar to that of an APF.

### 4.3.2 Phase-Locked Loop

The principal idea of phase locking PLL is to generate a signal whose phase angle is adaptively tracking variations of the phase angle of a given signal. The complexity of phase locking is however due to noise, distortions and frequency variations as discussed in Ghartemani et al (2004).

The conventional strategy for phase locking is to estimate the difference between phase angle of the input signal and that of a generated output signal and to regulate this value to zero by means of a control loop. The phase difference is estimated by a Phase Detector (PD) which is usually a
multiplier. The output signal is generated by a Voltage-Controlled Oscillator (VCO) whose signal is provided by a loop controller or a Loop Filter (LF).

An Enhanced PLL (EPLL) system is introduced in next section. Operation of the EPLL is based on estimating amplitude and phase angle of the fundamental component. The major improvement introduced by the EPLL is in the Phase Detection mechanism. The conventional PD mechanism is replaced by a new strategy which allows more flexibility and provides more information such as amplitude and phase angle. Upgrading the proportional to a proportional-integrating controller enables the EPLL to estimate the frequency as well.

4.3.2.1 Enhanced phase-locked loop

This section discusses a harmonic extraction and measurement unit as discussed in Ghartemani & Iravani (2005) which employs the Enhanced Phase-Locked Loop (EPLL) as the main building blocks as shown in Figure 4.9(c). The two main features of the harmonic estimator are (1) accurate performance in non-stationary frequency-varying environments and (2) immunity to noise.

This EPLL algorithm is capable of extracting the specified sinusoidal signal, estimating the corresponding amplitude and phase and accommodating variations in the amplitude and phase through Equation (4.14)-(4.18).

\[
\dot{A}(t) = \mu e(t)\sin \phi(t) \tag{4.14}
\]

\[
\dot{\phi}(t) = \mu \dot{e}(t)\cos \phi(t) \tag{4.15}
\]

\[
\phi(t) = \omega(t) + \mu, \dot{\omega}(t) \tag{4.16}
\]
State variables $A(t)$ and $\phi(t)$ directly provide instantaneous estimates of the amplitude and phase of the extracted sinusoid, respectively. Other components and noise imposed on the sinusoidal component of interest is $e(t)$. The parameters $\mu_1$, $\mu_2$ and $\mu_3$ are the controlling parameters of the algorithm in tracking variations in the amplitude and phase angles.

The block diagram of the EPLL is shown in Figure 4.9(b). The desirable objectives are adaptability to frequency varying situations, having a simple and robust structure suitable for practical applications and desirable noise immunity.

**Steps for realizing EPLL**

1. Real time / synthesized signal generation
2. EPLL realized in MATLAB Simulink using Equation (4.14)-(4.18)

\[
\dot{A}(t) = \mu_1 e(t) \sin \phi(t) \\
\dot{\phi}(t) = \omega(t) + \mu_3 \dot{\phi}(t) \\
\dot{\phi}(t) = \omega(t) + \mu_3 \dot{\phi}(t)
\]

where $\omega(t) = \omega_0 + \Delta \omega(t)$

\[
e(t) = u(t) - A(t) \sin \phi(t) \\
\mu_1 = 10, \mu_2 = 100 \text{ and } \mu_3 = 0.2
\]

3. Amplitude (A), harmonic signal $A(t) \sin \phi(t)$ extracted has been filtered through a low pass filter $1/\tau s + 1$ where $\tau = 1/(2\pi 10) s$

4. Example simulation through case analysis
Figure 4.9  (a) Block diagram of Phase-Locked Loop (b) Block diagram representation of frequency estimator  (c) Detailed representation of the $k^{th}$ paralleled EPLL
4.3.2.1 Performance analysis of EPLL

This section presents the results of digital simulation as case studies carried out to evaluate the performance of the harmonic measurement scheme. A set of five core units are employed to the block in Figure 4.9. The fundamental frequency is $f_o = 50\, \text{Hz}$. A set of values of $\mu_1 = 10, \mu_2 = 100$ and $\mu_3 = 0.2$ is chosen as the basis for the parameter settings. Each core unit has two identical first-orders, low-pass filters with cut-off frequency of $f_c = 10\, \text{Hz}$. The initial conditions of the integrators are set to zero. The amplitude and frequency of the recorded data from each unit are passed through low-pass filters of the same structure to produce smoother output data.

The input signal as in Equation (4.11) which consists of five harmonics is given to EPLL estimator which has five sub-units. The individual harmonic components and its amplitudes are shown in Figure 4.10 and Figure 4.11.

![Figure 4.10 Harmonic signal extraction using EPLL](image-url)
4.3.2.2 Amplitude tracking feature

The amplitude tracking feature of EPLL is tested through the following input signal. The input signal is simulated as in Equation (4.19) has amplitude of $A_1=1\text{p.u}$, $A_2=0.8\text{p.u}$, $A_3=0.6\text{p.u}$, $A_4=0.4\text{p.u}$, $A_5=0.2\text{p.u}$ till $t=1.9\text{sec}$ thereafter the amplitude changes are $A_1=2\text{p.u}$, $A_2=1\text{p.u}$, $A_3=0.8\text{p.u}$ which has been tracked through EPLL as shown in Figure 4.12.

$$t = (0:0.001:1.9);$$
$$t_1 = (1.901:0.001:10);$$
$$t_2 = [t_1; t_2];$$
$$x(n) = \sin(2\pi50t) + 0.8\sin(2\pi100t) + 0.6\sin(2\pi150t) + 0.4\sin(2\pi200t) + 0.2\sin(2\pi250t) + 1.4\sin(2\pi50t_1) +$$

$$\sin(2\pi100t_1) + 0.8\sin(2\pi150t_1)$$

(4.19)
4.4 COMPARISON BETWEEN APF, EPLL and WPT

Liao et al (2011) proposed a noise-suppression algorithm to enhance the WT technique in processing the noise-riding signals. The abilities of the WT in detecting and localizing the disturbances can hence be restored.

Barros et al (2012) present an extensive bibliographic review of the applications of wavelet transforms in the measurement and analysis of harmonic distortion in power systems.

This section presents various simulation results of the Wavelet Packet Transform as discussed in Section 1.7.3.1, Enhanced Phase Locked Loop and Adaptive Peak Filter as these methods possess signal decomposition technique. The performances of these three methods are compared in terms of accuracy, time of convergence and signal-tracking feature. The input signal taken for all the simulation is given as in Equation (4.20).
Signals’ sampling frequency applied to the Wavelet Packet Transform analysis is set as 1600Hz, the mother wavelet ‘Haar’ has been used with three levels of decomposition, so that each coefficient will have 100Hz as the uniform bandwidth. The designing parameter for EPLL method is set as $\mu_1=10$, $\mu_2=100$, $\mu_3=0.1$ and for Adaptive Peak Filter analysis are $\varphi_0=1.25$, $\alpha=5500$, $\mu=0.01$, and $N=2500$. Eight sub-units with odd harmonics have been taken for EPLL and APF analysis. The first order low pass filter with the transfer function of $1/((\tau s+1)$ where $\tau=1/(2\pi10)$s is included in posterior to amplitude and frequency estimation.

### 4.4.1 Harmonics Decomposition

The input signal as in Equation (4.20) is given to WPT, EPLL and APF to compare the performances of these methods. Extracted individual harmonics by WPT with its FFT has been plotted in Figure 4.13. Three levels of DWPT decomposition provides eight coefficients namely $C_1, C_2…C_8$.

\[
y(t) = 1 \times \sin(2\pi 50t) + \left( \frac{1}{3} \right) \times \sin(2\pi 50t \times 3) + \left( \frac{1}{5} \right) \times \sin(2\pi 50t \times 5) + \\
\left( \frac{1}{7} \right) \times \sin(2\pi 50t \times 7) + \left( \frac{1}{9} \right) \times \sin(2\pi 50t \times 9) + \left( \frac{1}{11} \right) \times \sin(2\pi 50t \times 11) + \\
\left( \frac{1}{13} \right) \times \sin(2\pi 50t \times 13) + \left( \frac{1}{15} \right) \times \sin(2\pi 50t \times 15)
\] (4.20)

The $C_1$ possess 50Hz fundamental, which lies in 0-100Hz band,

- $C_2$ has third harmonic of 150Hz, which lies in 100-200Hz band,
- $C_3$ has seventh harmonic of 350Hz, which does not lie in 200-300Hz band,
- $C_4$ has fifth harmonic of 250Hz, which does not lie in 300-400Hz band,
- $C_5$ has $15^{th}$ harmonic of 750Hz, which does not lie in 400-500Hz band,
- $C_6$ has $13^{th}$ harmonic of 650Hz, which does not lie in 500-600Hz band,
- $C_7$ has $9^{th}$ harmonic of 450Hz, which does not lie in 600-700Hz band,
- $C_8$ has $11^{th}$ harmonic of 550Hz, which does not lie in 700-800Hz band.
Coefficients $C_1$, $C_2$ alone give the correct location of the decomposed signal, but remaining coefficients $C_3\ldots C_8$ give their appearance at a different harmonic band. Each coefficient has only one harmonic for a specified band of frequency.

**Figure 4.13 Harmonic Decomposition using WPT**

EPLL with eight sub-units as shown in Figure 4.9 has been used to extract harmonic components and estimate their amplitude. The amplitude of the extracted individual component is shown in Figure 4.14. The range of frequency limiter in each sub unit has been chosen as ±25Hz, such that no overlapping occurs.

**Figure 4.14 Amplitude estimation through EPLL**
The same input signal is given to Adaptive Peak Filter discussed in Figure 4.4 and the estimated amplitude of individual harmonics is shown in Figure 4.15.

**Figure 4.15 Amplitude estimation through APF**

The eighth sub-unit which estimates fifteenth harmonic does not provide correct value up to 2sec, to avoid this effect an extra subfilter (i.e., ninth sub-unit) has to be added. Extracted harmonic component of eighth unit is shown in Figure 4.16.

**Figure 4.16 Decomposed $15^{th}$ order harmonic in eighth sub-unit of APF**
4.4.2 Speed of Convergence

The extracted fundamental component through APF and EPLL has been taken for analysis. The speed/time of convergence for EPLL method is 0.8sec. The actual and extracted fundamental component is shown in Figure 4.17.

![Figure 4.17 Actual and estimated fundamental component using EPLL](image)

By using APF, time of convergence for the fundamental component is 0.08sec as seen in Figure 4.18 which is ten times smaller than EPLL based analysis.

![Figure 4.18 Actual and estimated fundamental component using APF](image)
4.4.3 Harmonic Signal Tracking Capability

Tracking feature is the parametrical feature extraction wherein the
time-varying signal is identified and frequency switching in signal is
signatured at specific time instants. APF output for the frequency tracking is
compared with that of WPT as discussed in Section 4.4.1.

The tracking capability of APF and EPLL has been analyzed using
the signal as in Equation (4.16) with some changes incorporated in amplitude
and frequency. The amplitude of fundamental and third harmonic has been
decreased and increased by 0.1p.u; fifth and eleventh harmonic has been
eliminated from 1sec onwards. Using EPLL method the first, second, third
and sixth units’ extracted harmonic signals are shown in Figure 4.19. EPLL
takes approximately 0.3sec to track changes in the signal.

![Figure 4.19 Amplitude tracking capability of EPLL](image)

Extracted harmonic signals through APF’s first, second, third and
sixth units’ are shown in Figure 4.20. APF takes 0.05sec to track the variation
in the signal.
4.5 CASE STUDY: POWER QUALITY EVENT DETECTION THROUGH APF


Transformer winding fault has been used for discussion in this section to emulate different power quality disturbances, which have been organized into seven categories as transients, interruptions, sag or under voltage, swell or overvoltage, waveform distortion, voltage fluctuations, frequency variations. This section discusses on the above listed fault created
in the experimented prototype model of a transformer which has been analyzed through WT, APF and the corresponding PVRSD.

Experiment has been conducted using 230V, 50Hz, 30mA, single phase transformer with outputs at different tapings of 3V, 6V, 9V, 12V, 15V, 18V with 3A current rating. In general, during the winding short due to reduction in number of turns, the resistance of the phase coils decreases which lead to increase in current. A fault also leads to changes in the lumped parameter values of the coil due to change in inter-turn capacitance, change in coil inductance with change in coil length reduction due to shorted turns of coil.

4.5.1 Double Winding Short

The multipoint winding short in transformer occurs very seldom. The probability is that the event occurs with voltage variation. This case is analyzed based on test bench experimentation on the model test transformer of 3A rating discussed above. Figure 4.21(a) shows the signal for double winding short and its WT detail coefficient which depicts the time instant of short circuit at 1.2sec, 2.2sec and 2.6sec. Figure 4.21(b) shows the frequency and amplitude estimated using APF estimation, which shows explicitly the tracking of change in amplitude along with the time instant of amplitude switching for signal in Figure 4.21(a).
Figure 4.21  (a) Double winding short voltage signal and its WT detailed coefficients (b) Amplitude estimation through APF (c) PVRSD

4.5.2 Swell

A swell is the reverse form of sag, having an increase in AC voltage for time duration of 0.5cycles to 1minute. The sources of swells are
high-impedance neutral connections, sudden (especially large) load reductions and a single-phase fault on a three-phase system. The result can be data errors, flickering of lights, degradation of electrical contacts, semiconductor damage in electronics and insulation degradation in equipments.

The experiment is conducted with the 3A rating test bench transformer to study the event of swell using signal analysis technique. Here Figure 4.22(a) represents the signal, its WT detail represents the time instant of amplitude swell between 3.6sec and 5.6sec. Figure 4.22 (b) shows the role of APF to track the amplitude variation in signal.

4.5.3 Sag

Sag is a reduction of AC voltage at a given frequency for the duration of 0.5 cycles to 1 minute’s time. Sags are usually caused by system faults and switching on loads with heavy startup currents. Common causes of sags include starting large loads (starting of a large air conditioning unit) and remote fault clearing performed by utility equipment.

Figure 4.23(a) shows the event of sag obtained with experimenting in 3A test bench transformer. Figure 4.23(b) shows the WT detailed coefficients which signify the time instance of occurrence of sag as 0.1sec and 0.22sec. Figure 4.23(c) is obtained using APF estimators, which more explicitly traces the change in voltage level from 5V to 2V during 0.1sec to 0.22sec as sag which again rises after 0.22sec. Figure 4.23(d) depicts the PVRSD with frequency variation as 50Hz, 150Hz and 250Hz occurrence during the event of sag.
Figure 4.22 (a) Output voltage swell; its WT detailed coefficient; amplitude estimation through APF (b) PVRSD
Figure 4.23 (a) Output voltage sag and its one level WT detailed coefficient (b) amplitude estimation
Figure 4.23 (c) Frequency estimation through APF (d) PVRSD
4.5.4 Transients

Transients are potentially the most damaging type of power disturbance. Transients fall into two subcategories namely impulsive and oscillatory. Impulsive transients are sudden high peak events that raise the voltage and/or current levels in either a positive or a negative direction. Impulsive transients can be very fast events (5nsec rise time from steady state to the peak of the impulse) of short-term duration (less than 50nsec). Causes of impulsive transients include lightning strokes, poor grounding, switching of inductive loads, utility fault clearing, and ESD (Electrostatic Discharge). The results can range from the loss (or corruption) of data, to physical damage of equipment, though among these causes, the lightning impulse is probably the most damaging. An oscillatory transient is a sudden change in the steady-state condition of a signal’s voltage, current, or both, at both the positive and negative signal limits, oscillating at the natural system frequency. These transients occur when an inductive or capacitive load, such as a motor or capacitor bank is turned on or off. An oscillatory transient results because the load resists the change.

Figure 4.24(a) shows the event of transient obtained with introduction of a switch in the experiment with 3A test bench transformer. The WT detailed coefficients signifies the time instance during which transient has occurred is 0.1sec. The same has been detected through APF as in Figure 4.24(b).
Figure 4.24 (a) Shorting two tappings of the transformer for a moment and its one level WT detail coefficients (b) amplitude estimation through APF (c) PVRSD
4.5.5 Interrupt

An interruption is defined as the complete loss of supply voltage or load current. Depending on its duration, an interruption is categorized as instantaneous (0.5 to 30 cycles), momentary (30 cycles to 2 seconds), temporary (2 seconds to 2 minutes), or sustained (greater than 2 minutes).

Figure 4.25(a) shows the event of interrupt obtained with experimenting in the 3A test bench transformer. Figure 4.25(b) shows the WT detailed coefficients which signify the time instance of occurrence of transient as 0.1sec. Figure 4.25(c) shows that APF traces the amplitude change between 3.4sec as a dip which after 4.7sec rises towards 3V. Figure 4.25(a) interestingly traces the frequency tracking from 50Hz to 30Hz during the event, which is otherwise not observable through PVRSD.

4.5.6 Flicker

Voltage fluctuations are fundamentally different from the rest of the waveform anomalies. A voltage fluctuation is a systematic variation of the voltage waveform or a series of random voltage changes, of small dimensions, namely 95 to 105% of nominal at a low frequency, generally below 25Hz. Any load exhibiting significant current variations can cause voltage fluctuations. Arc furnaces are the most common cause of voltage fluctuation on the transmission and distribution system. One symptom of this problem is flickering of incandescent lamps.
Figure 4.25 (a) Interrupt signal and its amplitude and frequency estimation through two levels of APF for interrupt signal (b) WT detail coefficients (c) PVRSD
From PVR in supply signal Spectral Distribution (PVRSD) it is possible to trace the frequency components in the signal though however it is not possible to detect the nature of events as sag, swell or transient that occurs in the system.

Figure 4.26 (a) Input current of the flicker and its WT detailed coefficient (b) amplitude estimation through APF (c) PVRSD
The WT is good but it demands multiple level decomposition to identify fault signature in the signal. Also the difficulty is that it is not possible to apriori predict as to how many levels the WT based decomposition is to be evaluated, hence deeming WT to be based on trial and error method only. WT isolates the time instant of the event occurrence, however does not detect the nature of event as whether it is due to sag or swell or transient phenomenon. On the other hand, the APF identifies the event as sag or swells or transient based on its tracking of the signal amplitude and frequency through it requires a multistage filter structure. Also the time instant of signal switching in amplitude, frequency variation is also tracked in a simpler way. The analysis of signal shows that the adaptive comb filter gives best performances in all power system disturbances and the algorithm is also easier to implement.

4.6 ANALYSIS OF MACHINE FAULTS THROUGH APF

Figure 4.27(a) shows the extracted harmonic component for the signal shown in Figure 2.8 obtained for the healthy three phase SCIM with three phase balanced supply input under half-loaded condition. The signal exhibits peaks at both even and odd harmonic values as observed in PVR in spectrum of current signal. The signal obtained from experimental test bed is applied to Equation (4.6) and the APF response for amplitude and frequency estimation is shown in Figure 4.27(b) and Figure 4.27(c). The extracted harmonic signals through each stage of APF subfilter has frequency components of 50Hz, 100Hz, 150Hz, 200Hz which has been verified through PVRSD and shown in Figure 4.27(d)-4.27(g).
Figure 4.27 (a) APF based harmonic signal extractions (b) Frequency estimation with APF (c) Amplitude estimation through APF for input voltage supply given to SCIM
Figure 4.27 (d)-(g) PVRSD of first, second, third and fourth harmonic signal extracted through APF
Fault detection in machines has been done through APF. The stator winding fault signal discussed in Figure 2.13 is passed through WPT with ten levels of decomposition. The detail coefficient of the eighth level has been passed through APF to track the fault frequencies. This extracts the fault frequencies of 25Hz and its odd multiples namely 75Hz, 125Hz and 175Hz as shown in Figure 4.28(a). The detailed coefficient of the seventh level contains other harmonic components as shown in Figure 4.28(b).

Figure 4.28  (a) Stator winding fault frequencies (b) harmonics estimation using WPT and APF
The case examples discussed in Section 4.5 explicit the need and usefulness of APF for estimating signals variation in amplitude and frequencies in a simpler algorithm, rather than using the spectral estimation and wavelet transform methods that are computationally more intensive.

4.7 EFFECT OF SWITCHING FAULT IN POWER CONVERTERS ON DRIVES

Many reports discuss on electrical and mechanical faults due to open circuit or short circuit occurring at semiconductor power switch in a regulating converter module which would cause an unbalanced output voltage and current, leading to faults and eccentricities to loads like motors.

Simulation and experimental analysis of switching faults of a Matrix Converter (MC) (also termed as cycloconverter) connected to induction machine is considered as a special case for discussion here. The voltage/current output is analyzed through DSP techniques that indicate specific frequency patterns for faults occurring in converter-switching which are different from patterns for faults occurring in rotating machine.

Speed variation is observed when MC switching has the PWM duty cycle varied from 50% to 75% and 50% to 25%. The effect of converter subject to irregulated switching, would event into a fluctuation in motor speed that may lead to decrease in drive efficiency. It is required to trace such switching events in converters. Converter switching induces harmonics that need to be detected earlier. The following section discusses on APF estimator detecting matrix converter switching fault.

Kwak & Toliyat (2007) propose a PWM modulation strategy for fault-tolerant operation of the three phase matrix-converter-based drives against opened switch, opened phase faults, and shorted winding failures.
Sunter & Avdogmus (2008) discuss on implementation of a single-phase matrix converter for induction motor drive, using two modulation technique namely sinusoidal and square wave PWM through simulation and hardware. The author claims that through square wave PWM, single phase matrix converter drive provides good performance by reducing the order of harmonics and increasing the amplitude of the lower order harmonics.

As discussed by Filippetti et al (2000) if the induction machine is supplied by a frequency converter, the machine diagnostic system has to be combined with the fault detection possibilities in the converter.

For this purpose, an integrated diagnostic system associated with the monitoring of disturbances to electric drive has been considered in the next section. Typical faults of a converter cause changes in the supply waveforms and, consequently, in the spectrum of the input-output currents and voltages.

4.7.1 Single Phase Matrix Converter (SPMC) as Cycloconverter

Cycloconverters are used in high power applications such as variable frequency speed control for AC machines, constant frequency power supplies, controllable reactive power supply for an AC system and induction heating systems (Wheeler et al (2002), (Rashid 2005), Nguyen et al (2010), Nagalingam & Gurusamy (2010)). MC is a forced commutated converter that uses an array of controlled bi-directional switches as the main power elements to create a variable output voltage system with unrestricted frequency. It does not have any DC-link unit and does not need any large energy storage elements. In the conventional Single phase matrix converter the AC output voltage cannot increase the input voltage and both bi-directional switches of any phase leg can never be turned on at the same time. As the power electronic devices are prone for harmonics, matrix converter has been taken as
an example for the source of harmonics and notches/spikes, which has been analyzed through comparison on ease of detection with Adaptive Peak Filter, wavelet transform, spectrogram and cepstrum.

The single phase matrix converter that can convert the frequency from 50 Hz to 50 Hz and 50 Hz to 50/3Hz with resistive and inductive load as 100Ω and 20mH has been simulated using MATLAB/simulink and the same has been implemented in hardware using PIC16F877A microcontroller as in Figure 4.38. The output voltage signal from the matrix converter both from simulation and experiment have been given as input to Adaptive Peak Filter (APF) for harmonic extraction and estimation of amplitude, frequency and phase of these harmonics which have been shown in Figure 4.29-4.37 and tabulated in Table 4.1. Table 4.2 describes the quantitative comparison of simulated and experimental SPMC analysis through APF.

Table 4.1 List of figures for SPMC analysis through APF

<table>
<thead>
<tr>
<th>Figure No.</th>
<th>50Hz to 50/3 Hz SPMC</th>
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<tr>
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<td>Harmonic signal extraction</td>
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<tr>
<td>4.31</td>
<td>Frequency estimation of harmonics</td>
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<tr>
<td>4.32</td>
<td>Amplitude estimation of harmonics</td>
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<td><strong>50Hz to 50Hz SPMC</strong></td>
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<td>Amplitude estimation of harmonics</td>
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<td>4.36</td>
<td>Harmonic signal extraction</td>
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<td>4.37</td>
<td>Phase estimation 50Hz to 50Hz, 50Hz to 50/3Hz</td>
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<tr>
<td>4.38</td>
<td>Experimental setup</td>
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Figure 4.29 50Hz to 50/3Hz SPMC (a) simulated (b) Experimental
Figure 4.30 Harmonic signal extraction through APF for 50Hz to 50/3 Hz SPMC (a) Simulation (b) Experimental
Figure 4.31 Frequency estimation through APF for 50Hz to 50/3 Hz
SPMC (a) simulation (b) Experimental
Figure 4.32 Amplitude estimation through APF for 50Hz to 50/3Hz SPMC output voltage (a) simulation (b) Experimental
Figure 4.33 50Hz to 50Hz SPMC output voltage
(a) simulation
(b) Experimental
Figure 4.34 Frequency estimation through APF for 50Hz to 50Hz SPMC output voltage (a) simulation (b) Experimental
Figure 4.35 Amplitude estimation through APF for 50Hz to 50Hz SPMC output voltage (a) simulation (b) Experimental
Figure 4.36  Harmonic signal extraction through APF for 50Hz to 50Hz SPMC output voltage (a) simulation (b) Experimental
Figure 4.37  Phase estimation through APF for SPMC (Experimental) output voltage (a) 50Hz to 50/3Hz (b) 50Hz to 50Hz
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<td>50</td>
<td>83.33</td>
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<td></td>
<td></td>
<td>V</td>
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<td>3.76</td>
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<td>APF – Frequency Estimation</td>
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<td>APF – Amplitude Estimation</td>
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<td>4.94</td>
<td>2.6</td>
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Figure 4.38 Experimental setup of the matrix converter
Waveform distortion: There are five primary types of waveform distortion namely DC offset, harmonics, interharmonics, notching and noise. The periodic voltage disturbance caused by line-commutated thyristor circuits is called notching. The notching appears in the line voltage waveform during normal operation of power electronic devices when the current commutates from one phase to another. During this notching period, there exists a momentary short-circuit between the two commutating phases, reducing the line voltage; the voltage reduction is limited only by the system impedance. Notching is repetitive and can be characterized by its frequency spectrum. The frequency of this spectrum is quite high. Usually it is not possible to measure it with equipment normally used for harmonic analysis. Notches can impose extra stress on the insulation of transformers, generators, and sensitive measuring equipment. In matrix converter, the inductive load introduces a spike which is otherwise called notching which has been analyzed through spectrogram, cepstrum and one level Haar wavelet transform for 50Hz to 50/3Hz and 50Hz to 50Hz matrix converter with RL load as shown in Figure 4.39 and Figure 4.40.
Figure 4.39 (a) Input voltage signal (b) output voltage signal (c) Spectrogram (d) Cepstrum (e) wavelet approximation coefficient (f) Detail coefficient of 50Hz to 50/3Hz matrix converter output signal
Figure 4.40 (a) Input voltage signal (b) output voltage signal (c) Spectrogram (d) Cepstrum (e) wavelet approximation coefficient (f) Detail coefficient of 50Hz to 50Hz matrix converter output signal
The output voltage waveform due to faulty switching in the converter is shown in Figure 4.41(a) and its power spectrum in Figure 4.41(b), which is observed to induce slip changes in the rotating machine. This new analysis supports on-line detection of the faults even when the machine is running under loaded condition. Even harmonic frequency values are induced when the pulse period is changed.

![Waveform](image)

![Power Spectrum](image)

**Figure 4.41** (a) 50Hz-50/3Hz matrix converter with switching fault (b) PVRSD for signal shown in Figure 4.41(a)
SPMC with sinusoidal PWM and 0.7 modulation index has been used to convert 50Hz to 50/4Hz. The pulse generator parameters used for correct and fault switching pattern for the operation of SPMC are given in Table 4.3. Single phase induction motor has been connected to 50Hz to 50/4Hz SPMC, whose specifications are given in Table 4.4. Due to change in the pulse period the pattern of the output voltage waveform changes and thus there is change in speed in the machine. Speed rises to 400RPM and settles to 350RPM approximately for correct switching as shown in Figure 4.42(a). Speed rises to 400RPM and then decreases to nearly 325RPM due to impact of change in period of pulse width as seen in Figure 4.42(b).

Table 4.3 Parameters of pulse generator to 50Hz to 50/4Hz SPMC

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<tr>
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<th>Period</th>
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<th>Phase delay</th>
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<td>50</td>
<td>0.04</td>
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Table 4.4 Specifications of Single Phase Induction Motor (SPIM)

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<td>Main winding stator</td>
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<tr>
<td>Main winding rotor</td>
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</tr>
<tr>
<td>Main winding mutual inductance</td>
<td>0.18mH</td>
</tr>
<tr>
<td>Auxillary winding stator</td>
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<tr>
<td>Inertia</td>
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<tr>
<td>Friction factor</td>
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<td>Pole Pair</td>
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<tr>
<td>Turn ratio</td>
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<tr>
<td>Capacitor-start</td>
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Figure 4.42 (a) Speed of SPIM with SPMC fed with 50% duty cycle for 50Hz to 50/4Hz (correct switching) (b) Speed of SPIM with SPMC fed with 75% duty cycle for 50Hz to 50/4Hz (fault switching)
4.8 CONCLUSION

This chapter discusses on role of APF (Adaptive Peak Filter) to detect effect of power system event on equipment performances. The study is done with emulating disturbances through (i) transformer winding short circuit faults, (ii) the stator winding fault in SCIM machines, (iii) detecting effect of converter switching faults on rotating machine performance.

Switching fault in SPMC leads to changes in load like variation in machine speed demanding early detection. Switching in SPMC has been detected using different signal processing techniques. The APF filters estimate the signal frequencies, their respective amplitude and phase effectively compared to the other frequency transforms as FFT and WT.

This chapter illustrated the role of estimators with the APF observed to be most effective since it is successfully able to demarcate the variation in amplitude, frequency and phase of each one of the harmonic component in the signal. The APF estimator is observed to be very useful and can be implemented as a simple on-chip algorithm in an advanced embedded processor interfaced to any electrical apparatus. The need for advanced embedded processor is discussed in the next chapter.