CHAPTER

3

FRONT-END SNR ESTIMATION TECHNIQUE FOR OFDM SYSTEMS

3.1 Introduction

In the last chapter we discussed the background of OFDM systems and SNR estimation. The completion of the background study now sets the stage for the development of an improved SNR estimation technique. Formulation of proposed SNR estimation technique will be discussed in this chapter. In order to validate the proposed technique, schemes of Reddy [Reddy, S et al, 2003] and subspace [Xiaodong et al, 2005] are used. The formulation of these techniques is also described before the formulation of our technique. Methodologies of these estimation techniques will be discussed in the next chapter.

SNR estimation indicates the reliability of the link between the transmitter and receiver. In adaptive system, SNR estimation is commonly used for measuring the quality of the channel and accordingly changing the system parameters. For example, if the measured channel quality is low, the transmitter may add some redundancy or complexity to the information bits (more powerful coding), or reduce the modulation level (better Euclidean distance), or increase the spreading rate (longer spreading code) for lower data rate transmission. Therefore, instead of implementing fixed information rate for all levels of channel quality, variable rates of information transfer can be used to maximize system resource utilization with high quality of user experience.

Many SNR estimation algorithms have been suggested and implemented in the last ten years in OFDM systems at the back-end of receiver using the system pilot symbols. The essential requirement for an SNR estimator in OFDM system is of low computational load. This is in order to minimize hardware complexity as well as the computational time. Most of these techniques derive the symbol SNR estimates solely from the received signal at the output of the matched filter (MF). The estimators assume perfect carrier and symbol synchronization while at the same time implicitly assuming intersymbol interference (ISI)-free output of the MF (the decision variable). However, in practice, multipath wireless communication gives rise to much intersymbol interference, especially in indoor and urban areas. In these ISI dominated scenarios, SNR estimators that do not presume ISI-free reception are highly desirable.

In contrast to other SNR estimators, the proposed technique operates on data collected at the front-end of the receiver, imposing no restriction on ISI. This will improve the SNR estimates in severe ISI channels and also help extending the implementation of SNR estimators towards systems that require SNR estimates at the input of the receiver. One such application is antenna diversity combining, where at least two antenna signal paths are communicably connected to a receiver. The combiner can use the SNR estimates obtained for each antenna signal to respectively weight each signal and thereby generate a combined output signal.

In many SNR estimation techniques, noise is assumed to be uncorrelated or white. But, in wireless communication systems, where noise is mainly caused by a strong interferer, noise is colored in nature.

In this chapter, formulation and methodology of Proposed SNR estimation technique are presented. A front-end noise power and SNR estimator for the white noise as well as for colored noise in OFDM system is proposed. The algorithm is based on the two identical halves property of time synchronization preamble used in some OFDM systems. The proposed technique is divided in to two parts. In the first part, SNR estimation technique for AWGN channel and multipath channels is considered. In the second part, the proposed estimator is taking into consideration the different noise power levels over the OFDM sub-carriers. The OFDM band is divided into several sub-bands using wavelet packet and noise in each sub-band is considered white. The second-order statistics of the transmitted OFDM preamble are calculated in each sub-band and the power noise is estimated. Therefore, the proposed approach estimates both local (within smaller sets of subcarriers) and global (over all sub-carriers) SNR values. The short term local estimates calculate the noise power variation across OFDM sub-carriers. When the noise is white, the proposed algorithm works as well as the conventional noise power estimation schemes, showing the generality of the proposed method.

The remainder of the chapter is organized as follows. Section3.2 describes the Formulation of our proposed SNR estimation technique. In Section3.3 we discuss the formulation of our proposed technique.

3.2 Formulation of Proposed Front-End SNR estimation Technique

In this section, we presented our proposed front-end SNR estimation technique. This is a data-aided technique based on the preamble appended to data frames. According to the best knowledge of the author, there is no estimation technique that estimates SNR at front-end of the receiver for OFDM systems. The motivation in using a preamble for SNR estimation is that a preamble is always present in most OFDM systems, and, therefore, using it will not cause any additional burden. Preamble is used both for channel estimation and timing and frequency synchronization. A good preamble must sound the channel very efficiently. It will also have low peak-to-average power ration (PAPR). Also it will have features that are best utilized for synchronization. For example popular Schmidl & Cox [Schmidl et al, 1997] technique for synchronization loads a PN-sequence on alternate subcarriers. This generates a preamble that has two identical halves, whose correlation gives an identification of the start of the frame. This also implies that channel will have to be interpolated at subcarriers where PN-sequence was not loaded.

Next subsection discusses the work that has been done in the literature on arriving at a suitable preamble.

3.2.1 Selection of a Preamble

For the proposed technique we aim to select a preamble which is already employed in OFDM systems for timing & frequency synchronization and channel estimation. Given below is a brief historical development of the work done in developing synchronization schemes and a suitable preamble for it.

> There have been several papers on the subject of synchronization for OFDM in recent years. Moose gives the maximum likelihood estimator for the carrier frequency offset which is calculated in the frequency domain after taking the FFT [Moose et al, 1994]. He assumes that the symbol timing is known, so he just has to find the carrier frequency offset. The limit of the acquisition range for the carrier frequency offset is the subcarrier spacing. He also describes how to increase this range by using shorter training symbols to find the carrier frequency offset. For example shortening the training symbols by a factor of two would double the range of carrier frequency acquisition. This approach will work to a point, but the estimates get worse as the symbols get shorter because there are fewer samples over which to average, and the training symbols need to be kept longer than the guard interval so that the channel impulse response does not cause distortion when estimating the frequency offset.
>
> Nogami present algorithms to find the carrier frequency offset and sampling rate offset [Nogami et al, 1995]. They use a null symbol where nothing is transmitted for one symbol period so that the drop in received power can be detected to find the beginning of the frame. The carrier frequency offset is found in the frequency domain after applying a Hanning window and taking the FFT. The null symbol is also used by Bot [Bot et al,1994]. This extra overhead of using a null symbol is avoided by using the technique described in this work. If instead of a continuous transmission mode, a burst mode is

used, it would be difficult to use a null symbol since there would be no difference between the null symbol and the idle period between bursts.

Beek describes a method of using a correlation with the cyclic prefix to find the symbol timing [Beek et al, 1995]. If this method were used to find the symbol timing, while using one of the previous methods to find the carrier frequency offset, there would still be a problem of finding the start of the frame to know where the training symbols are located.

Classen introduces a method which jointly finds both the symbol timing and carrier frequency offset [Classen et al, 1995]. However, it is very computationally complex because it uses a trial and error method where the carrier frequency is incremented in small steps over the entire acquisition range until the correct carrier frequency is found. It is impractical to do the exhaustive search and go through a large amount of computation at each possible carrier frequency offset.

Schmidl introduces some modifications of Classen's method which both greatly simplify the computation necessary for synchronization and extend the range for the acquisition of carrier frequency offset [Schmidl et al, 1997]. The method in this paper avoids the extra overhead of using a null symbol, while allowing a large acquisition range for the carrier frequency offset. By using one unique symbol which has a repetition within half a symbol period, this method can be used for bursts of data to find whether a burst is present and to find the start of the burst. Acquisition is achieved in two separate steps through the use of a two-symbol training sequence, which will usually be placed at the start of the frame. First the symbol/frame timing is found by searching for a symbol in which the first half is identical to the second half in the time domain. Then the carrier frequency offset is partially corrected, and a correlation with a second symbol is performed to find the carrier frequency offset.

After studying we select an OFDM synchronization preamble proposed by Schmidl [Schmidl et al, 1997]. As will be shown later, the preamble has two identical halves property. In this preamble, the two halves of the training symbol in time obtained by transmitting a pseudonoise (PN) sequence on the even frequencies, while zeros are used on the odd frequencies. This means that at each even frequency one of the points of a QPSK constellation is transmitted. In order to maintain approximately constant signal energy for each symbol the frequency components of this training symbol are multiplied by $\sqrt{2}$ at the transmitter. Transmitted data will not be mistaken as the start of the frame since any actual data must contain odd frequencies. Note that an equivalent method of generating this training symbol is to use an IFFT of half the normal size to generate the time domain samples. The repetition is not generated using the IFFT, so instead of just using the even frequencies, a PN sequence would be transmitted on all of the subcarriers to generate the time domain samples which are half a symbol in duration. These time-domain samples are repeated (and properly scaled) to form the first training symbol. The second training symbol contains a PN sequence on the odd frequencies to measure these sub-channels, and another PN sequence on the even frequencies to help determine frequency offset.

The selection of a particular PN sequence should not have much effect on the performance of the synchronization algorithms. Instead the PN sequence can be chosen on the basis of being easy to implement or having a low peak-to-average power ratio so that there is little distortion in the transmitter amplifier.

3.2.2 Proposed Front-End SNR Estimator

The second stage after selecting synchronization preamble - the preamble which has two identical halves property, is to develop a preamble based front-end SNR estimation technique. The selected preamble has low PAPR and can sound the channels well and can be power boosted more with the same system power requirements. This translates into higher signal to noise ratio (SNR). The proposed technique is divided in to two parts. The first part is the front-end SNR estimation technique for AWGN channel and multipath channels (Rayleigh, Rician, SUI channels and indoor channel models) using white noise scenario. The second part is the extension of first part to noise power estimation of colored noise using wavelet-packet based analysis of the noise.

3.2.2.1 First Part: Front-End Noise Power and SNR Estimation Technique for AWGN Channel and Wireless Multipath Channels

As discussed earlier, the preamble used for timing synchronization is derived from alternate loading of subcarriers with PN-sequence modulated constellation as follows:

$$P_{even}(k)=\begin{cases}\sqrt{2}.P(m) & k=2m \qquad m=1,2,3........N/2\\ 0 & otherwise\end{cases} \tag{3.1}$$

Here, $P(m)$ is the PN sequence loaded onto even subcarriers taken from IEEE802.16d. The factor $\sqrt{2}$ is related to the 3 dB boost and k shows the sub-carriers index.

In actual practice, an OFDM signal is provided with a guard band on either side of its spectrum. Accordingly the data are not loaded on the sides. For example, for a typical IEEE802.16d signal of length 256 subcarriers vide, 28 carriers on either side are null carriers as shown in Fig.3.1.

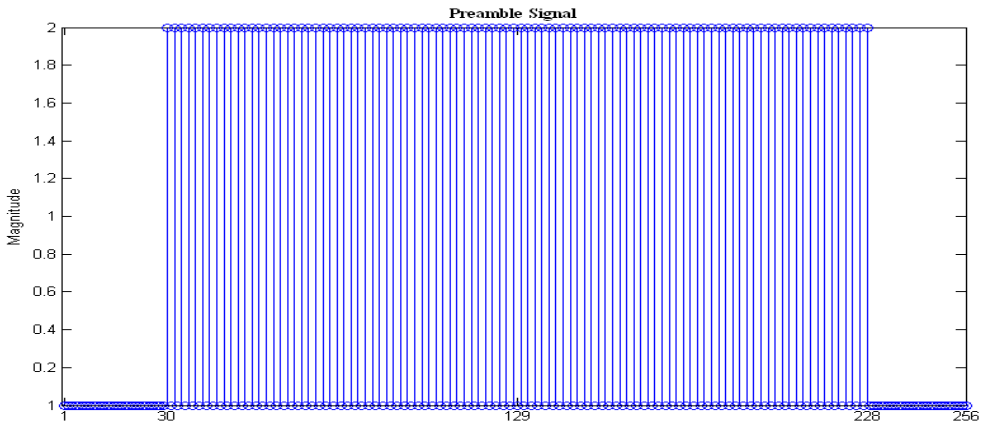

Fig.3.1 Preamble signal loaded on even subcarriers using PN sequence

Therefore for our purposes, eq.3.1 is rewritten as

$$P_{even}(k)=\begin{cases}\sqrt{2}.P(m) & k=2m \quad m=15,16,17.........(N/2-14)\\ 0 & m=1,2,3......14\\ 0 & m=N/2-13, N/2-12,.....N/2\end{cases} \tag{3.2}$$

The corresponding time-domain preamble $P(n)$, is obtained by Inverse discrete Fourier transform (IDFT) of $P_{even}(k)$ as follows.

$$\begin{aligned} p(n) &= IDFT\ \{P_{even}(k)\} \\ &= \sum_{k=0}^{N-1} P_{even}(k).e^{j2\pi nk/N} \qquad 0 \le n \le N-1 \end{aligned} \tag{3.3}$$

Since $P_{even}(k)$ has values only at even subcarriers, this can be seen from the properties of $e^{j2\pi nm/N/2}$ (also written as $W_{N/2}^{-nm}$, where W_N is the *N-th* root of unity).

For $k = 2m,$

$$e^{j2\pi n2m/N} = e^{j2\pi nm/N/2} \tag{3.4}$$

So, for $n = n + N/2$,

$$
\begin{aligned}
e^{j2\pi(n+N/2)m/N/2} &= e^{j2\pi nm/N/2} . e^{j2\pi m.N/2/N/2} \\
&= e^{j2\pi nm/N/2}
\end{aligned}
\tag{3.5}
$$

In other words

$$p(n) = p(n + N/2) \tag{3.6}$$

To avoid intersymbol interference (ISI) caused by multipath fading channels, cyclic prefix (CP) of length l_{CP} is added so that the total length of OFDM data becomes $N_{total} = N + l_{CP}$. It is assumed that the signal is transmitted over Rayleigh multipath fading channel characterized by

$$h(t, \tau) = \sum_{l=1}^{L} h_l(t)\delta(t - \tau_l) \tag{3.7}$$

where $h_l(t)$ are the different path complex gains, τ_l are different path time delays, and L is the number of paths. $h_l(t)$ are wide-sense stationary (WSS) narrow-band complex Gaussian processes. At the receiver side, with the assumption that the guard interval duration is longer than the channel maximum excess delay, the received OFDM data can be represented by

$$y(n) = x(n) + n(n) \tag{3.8}$$

where

$x(n) = s(n) * h(n)$

$*$ = Linear convolution

$s(n) = IDFT\ \{S(K)\}$, $S(K)$ are the constellation symbols, and $S(n)$ is the transmitted signal in time-domain.

$n(n) =$ white Gaussian noise with variance σ^2.

$h(n) =$ discretized version of impulse response of the system.

a. Autocorrelation based Front-End SNR Estimator

The proposed estimator is deployed right at the front-end of the receiver. It makes use of two identical halves property of time synchronization preamble padded with cyclic prefix and relies on the autocorrelation of the same. From eq.3.9, it can be shown that the autocorrelation function of the received signal, $R_{yy}(m)$, has the following relationship to the autocorrelation of the data signal, $R_{xx}(m)$ and the noise, $R_{nn}(m)$:

$$R_{yy}(m) = R_{xx}(m) + R_{nn}(m) \tag{3.9}$$

where

$$R_{yy}(m) = \sum_n y(n)\ y^*(n+m)$$

$$R_{xx}(m) = \sum_n x(n)\ x^*(n+m)$$

$$R_{nn}(m) = \sum_n n(n)\ n^*(n+m)$$

The noise in the channel is modeled as additive white Gaussian noise and its autocorrelation function only has a value at a delay of $m = 0$, with magnitude given by the noise variance (σ^2), expressed as

$$R_{nn}(m) = \sigma^2 \delta(m) \tag{3.10}$$

where $\delta(m)$ is the discreet delta sequence.

b. Signal Power and Noise Power Estimation

We undertake the study of OFDM signal statistics, and observe, as shown in Fig.3.2, that its power spectrum is nearly white. Hence its autocorrelation is generally given by:

$$R_{ss}(m) = P_o \delta(m) \tag{3.11}$$

where P_o is signal power.

However, two identical halves of the preamble OFDM symbol correlate separately and also together at *–N/2* lag, 0 lag and at *N/2* lag, giving rise to $R_{ss}(m)$ as:

$$R_{ss}(m) = P_o \{ {}^{1}\!/_{2}\, \delta(m - {}^{N}\!/_{2}) + \delta(m) + {}^{1}\!/_{2}\, \delta(m + {}^{N}\!/_{2}) \} \tag{3.12}$$

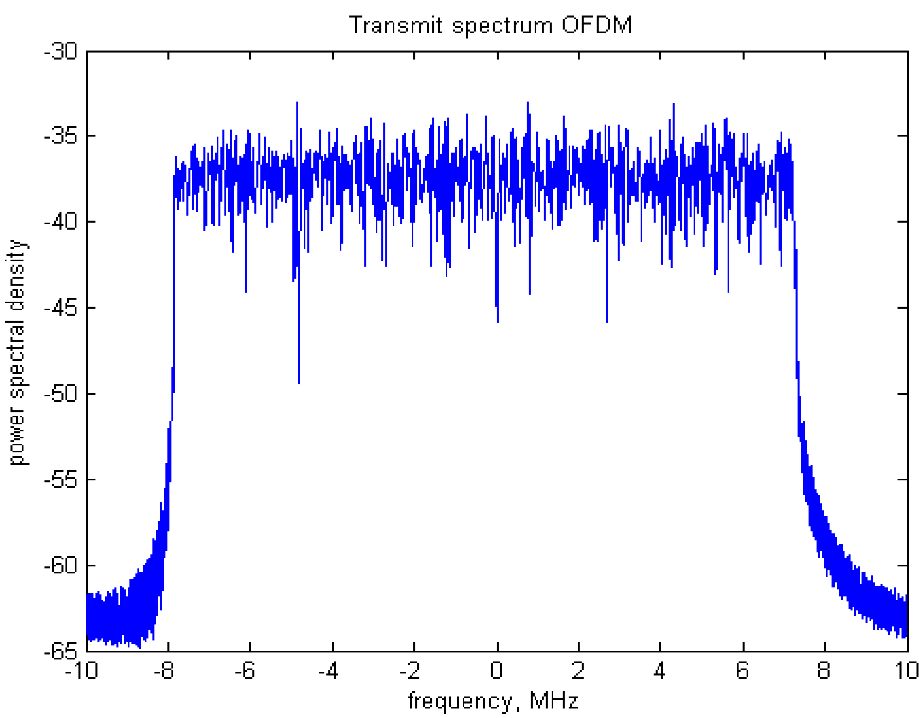

Fig.3.2 Power spectrum of an OFDM signal

As the transmitted signal passes through a channel, $h(n)$ the autocorrelation $R_{xx}(n)$ can be derived as follows:

$$
\begin{aligned}
R_{xx}(m) &= \sum_n s(n)\, s^*(n+m) \\
&= \sum_n \left[\sum_i h(i)\, p(n-i) \right] \left[\sum_j h^*(j)\, p^*(n+m-j) \right] \\
&= \sum_i \sum_j h(i)\, h^*(j) \left[\sum_n p(n-i)\, p^*(n+m-j) \right] \\
&= \sum_i \sum_j h(i)\, h^*(j) \left[\sum_n p(n-i)\, p^*(n-i+m-j+i) \right] \\
&= \sum_i \sum_j h(i)\, h^*(j).P_o.\delta(m-j+i) + \sum_i \sum_j h(i)\, h^*(j).\frac{P_o}{2}\delta(m-j+i-N/2) \\
&\quad + \sum_i \sum_j h(i)\, h^*(j).\frac{P_o}{2}\delta(m-j+i+N/2) \\
&= P_o \sum_i h(i) \left[\sum_j h^*(j)\,\delta(m-j+i) \right] + \frac{P_o}{2} \sum_i h(i) \left[\sum_j h^*(j)\,\delta(m-j+i-N/2) \right] \\
&\quad + \frac{P_o}{2} \sum_i h(i) \left[\sum_j h^*(j)\,\delta(m-j+i+N/2) \right] \\
&= P_o \sum_i h(i)\, h^*(m+i) + \frac{P_o}{2} \sum_i h(i)\, h^*(m+i-N/2) + \frac{P_o}{2} \sum_i h(i)\, h^*(m+i+N/2)
\end{aligned}
\tag{3.13}
$$

When $m = 0$,

$$
R_{xx}(0) = P_o \sum_i |h(i)|^2 + \frac{P_o}{2} \sum_i |h(i)h(i-N/2)| + \frac{P_o}{2} \sum_i |h(i)h(i+N/2)| \tag{3.14}
$$

However

$$\sum_i \left| h(i)\, h(i - N/2) \right| = 0$$

$$\sum_i \left| h(i)\, h(i + N/2) \right| = 0$$

As described in Fig.3.3.

So,

$$R_{xx}(0) = P_o \sum_i \left| h(i) \right|^2 \tag{3.15}$$

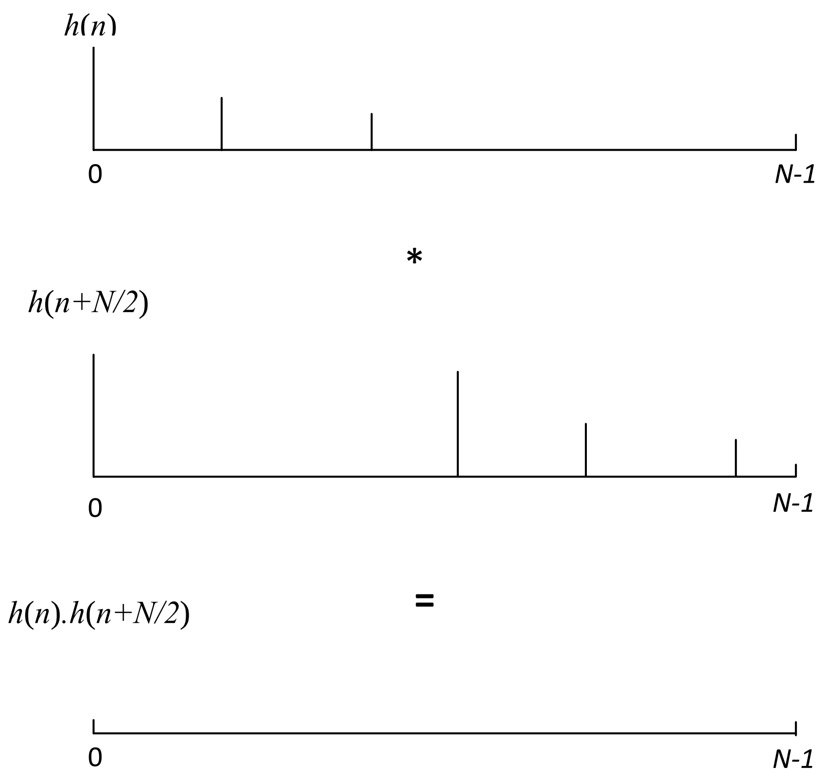

Fig.3.3. multiplication of *h*(*n*) and *h*(*n*+*N/2*)

Here $P_o \sum_i |h(i)|^2$ is the received power attenuated by $\sum_i |h(i)|^2$ factor.

When $m = N/2$,

$$R_{xx}(\frac{N}{2}) = P_o \sum_i \left|h(i)h(i+N/2)\right| + \frac{P_o}{2}\sum_i \left|h(i)\right|^2 + \frac{P_o}{2}\sum_i \left|h(i)h(i+N)\right| \tag{3.16}$$

However, similarly from Fig.3.3.

$$\sum_i \left|h(i)h(i+N/2)\right| = 0$$

$$\sum_i \left|h(i)h(i+N)\right| = 0$$

So

$$R_{xx}(\frac{N}{2}) = \frac{P_o}{2}\sum_i \left|h(i)\right|^2 \tag{3.17}$$

When $m = -N/2$,

$$R_{xx}(\frac{-N}{2}) = P_o \sum_i \left|\sum_i \left|h(i)h(i-N/2)\right|\right| + \frac{P_o}{2}\sum_i \left|h(i)h(i-N)\right| + \frac{P_o}{2}\sum_i \left|h(i)\right|^2 \tag{3.18}$$

However, similarly from Fig.3.3.

$$\sum_i \left|h(i)h(i-N/2)\right| = 0$$

$$\sum_i \left|h(i)h(i-N)\right| = 0$$

So,

$$R_{xx}(\frac{-N}{2}) = \frac{P_o}{2}\sum_i \left|h(i)\right|^2 \tag{3.19}$$

From the above equations, it is clear that the received signal power ($\frac{P_o}{2}\sum_i |h(i)|^2$) can be estimated from $R_{xx}(\frac{N}{2})$ and $R_{xx}(\frac{-N}{2})$ peak.

Hence, at zero lag (shown at 'L' in Fig.3.4) the autocorrelation $R_{xx}(0)$, contains both the signal power estimate and noise power estimate indistinguishable from each other as shown in eq.3.9 and eq.3.10 before. However, because of the identical halves nature of the preamble, the received signal power can be estimated from auto correlation peak at N/2 or at –N/2 as shown in Fig.3.3a. In Fig.3.3, $R_{xx}(m)$ has been sketched for $N=256$.

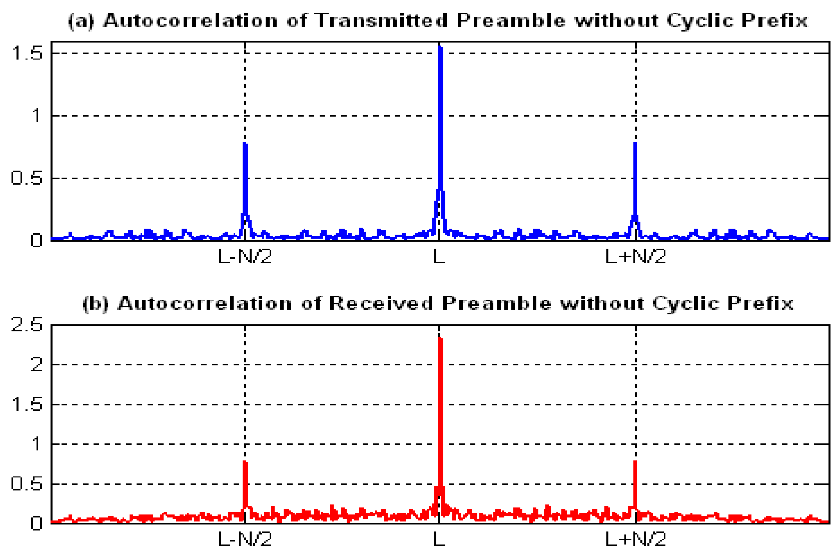

Fig.3.4 (a): Transmitted Preamble (b): Received Preamble after coming through a channel (Plots show two identical halves with no cyclic prefix).

It is clear that the autocorrelation values apart from the zero-offset are unaffected by the channel effects, so one can find the signal power from the *N/2* or *–N/2* lag autocorrelation value.

c. Signal Power Estimation

In Fig.3.4, the auto correlation for preamble has been shown without cyclic prefix. However, all OFDM symbols, including preamble, have cyclic prefix. For example, in WiMAX systems, a cyclic prefix 64 sample long is introduced where the data is 256 points long. In such cases, the autocorrelation also has peaks where cyclic prefix matches with sections. This is shown in Fig.3.5 where the autocorrelation of the clean OFDM signal is shown in Fig.3.5 (a) and that of the received signal with noise at SNR=7dB in Fig.3.5 (b).

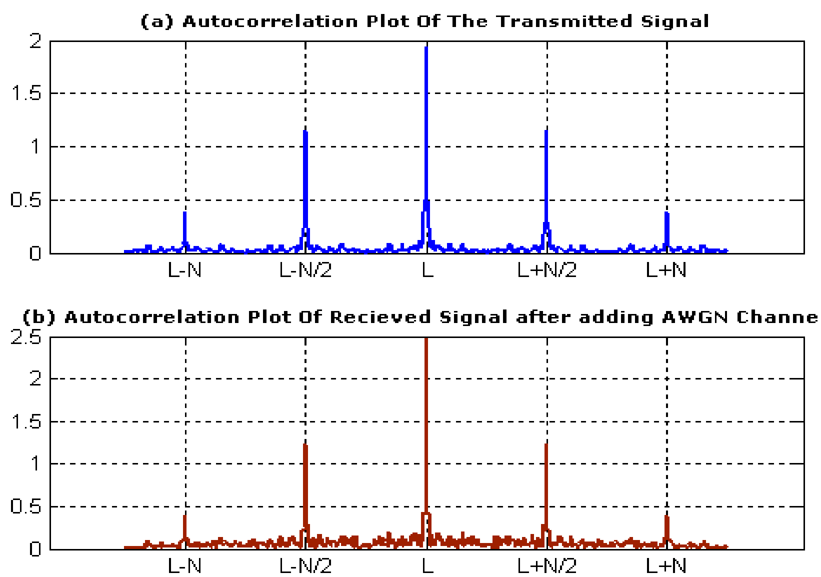

Fig 3.5 (a): Autocorrelation plot of transmitted signal (b): Autocorrelation plot of received signal.

The explanation for the correlation peaks of Fig.3.5 is pictorially given in Fig.3.6.

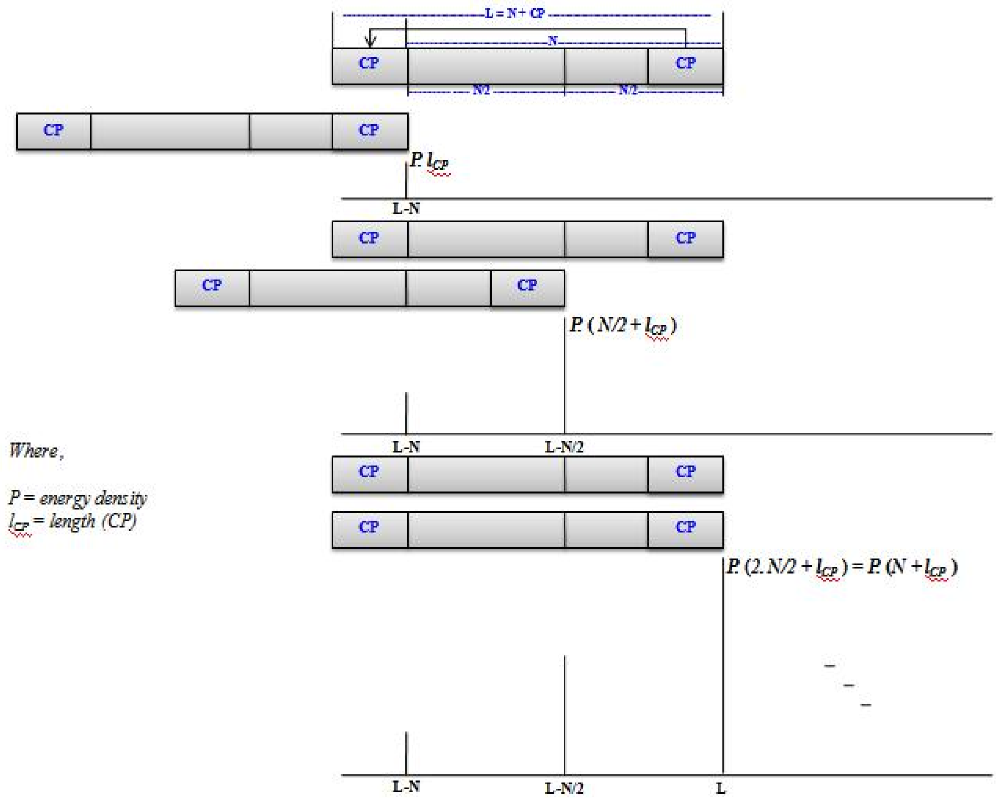

Fig.3.6: Autocorrelation of received preamble signal in time domain

If the energy density of the preamble is 'ρ' per unit length, the peak heights are equal to the energy of the matched portion of the signal. As it can be seen from the correlation shown in Fig.3.6 that, the first peak rises at *L-N* when CP matches with itself with energy of $\rho.\ l_{CP}$. The second peak rises at *L-N/2* when one half of preamble plus CP matches with itself with energy of $\rho.$ $(N/2+l_{CP})$ and main peak at zero-lag (shown at '*L*' in Fig.3.6) rises when full preamble matcheing with itself with energy of $\rho.(N+l_{CP})$. Taking into consideration the autocorrelation values for $'L-N/2','L-N','L+N/2','L+N'$, signal power is given as:

$$\hat{P}_o = 2R_{yy}(L-N/2) - R_{yy}(L-N) \tag{3.20}$$

Or

$$\hat{P}_o = 2R_{yy}(L+N/2) - R_{yy}(L+N) \tag{3.21}$$

where $\hat{P}_o$ estimated signal power.

d. Noise Power Estimation

Having obtained the power of signal, noise power, P_N given by noise variance $\hat{\sigma}_N^2$, can be calculated as

$$\text{Noise Power} = \hat{\sigma}_N^2 = R_{yy}(L) - \hat{P}_o \tag{3.22}$$

where $R_{yy}(L)$ is value at zero-lag.

e. SNR Estimation

Finally we can find the SNR estimates using eq.3.20 or eq.3.21 and eq.3.22.

$$S\hat{N}R = \frac{\hat{P}_o}{\hat{\sigma}_N^2} \tag{3.23}$$

where $S\hat{N}R$ is the estimated value for SNR.

Ideally, signal power and noise power are calculated without CP for the original data of length *N*. For example in WiMAX systems the data length is *N=256* and after adding cyclic prefix of *¼N* it becomes 320. Table 3.1 shows that signal power and noise power calculated for the proposed method is same as ideal case because the energy contained in CP is subtracted from the energy contained by total signal which is data plus CP.

Table 3.1: Ideal vs. calculated SNR for first-part of proposed technique

Signal power= P_{ss}	(Ideal) $\rho.N$	(Calculated) $P_{ss} = \rho\{2(N + l_{CP})\} - \rho.l_{CP} = \rho.N$
Noise Power= P_{NN}	$\hat{\sigma}_N^2.N$	$P_{NN} = \hat{\sigma}_N^2(N + l_{CP}) - \hat{\sigma}_N^2.l_{CP} = \hat{\sigma}_N^2.N$
SNR	P_{ss}/P_{NN}	P_{ss}/P_{NN}

3.2.2.2 Second Part: Front-End Noise Power and SNR Estimation of Colored Noise Using Wavelet-Packet

For the second part of our proposed technique; we develop a technique that takes into account the color and variation of noise statistics over OFDM sub-carriers. Unlike first part, the OFDM band is divided into several sub-bands using wavelet packet as shown in Fig.3.7 and appendix A shows the details of Wavelet Packet decomposition. The colored noise in each sub-band is considered white as shown in Fig.3.8. The proposed solution provides many local estimates, allowing tracking of the variation of the noise statistics across OFDM sub-carriers, which are particularly of use in sub-band adaptive modulation OFDM systems. The proposed technique estimates both local (within smaller sets of subcarriers) and global (over all sub-carriers) SNR values using noise power estimates knowledge.

After adding cyclic prefix as described in first part, OFDM data is divided into 2^n sub-bands using wavelet packets where '*n*' shows the number of levels. The length of each sub-band is $L_{sub}=N_{sub} + l_{CPsub}$, *where* $N_{sub}= N/2^n$ and $l_{CPsub} = l_{CP}/2^n$.

a. Signal Power and Noise Power Estimation in Sub-Bands

Sub-bands inherit the two identical halves property of synchronization preamble as discussed in first part of proposed technique. So, one can find the signal power and noise power in each sub-band using the same procedure as described in first part of proposed work. Due to wavelet packet decomposition, length of data is changed but location of zero lag and side peaks are unchanged. The autocorrelation of the transmitted and received 5th sub-band signal at SNR = 7 dB are shown in Fig. 3.9 (a) and Fig. 3.9(b), respectively. It is clear that the autocorrelation values apart from the zero-offset are unaffected by the AWGN, so one can find the signal and noise powers from the zero-lag autocorrelation value.

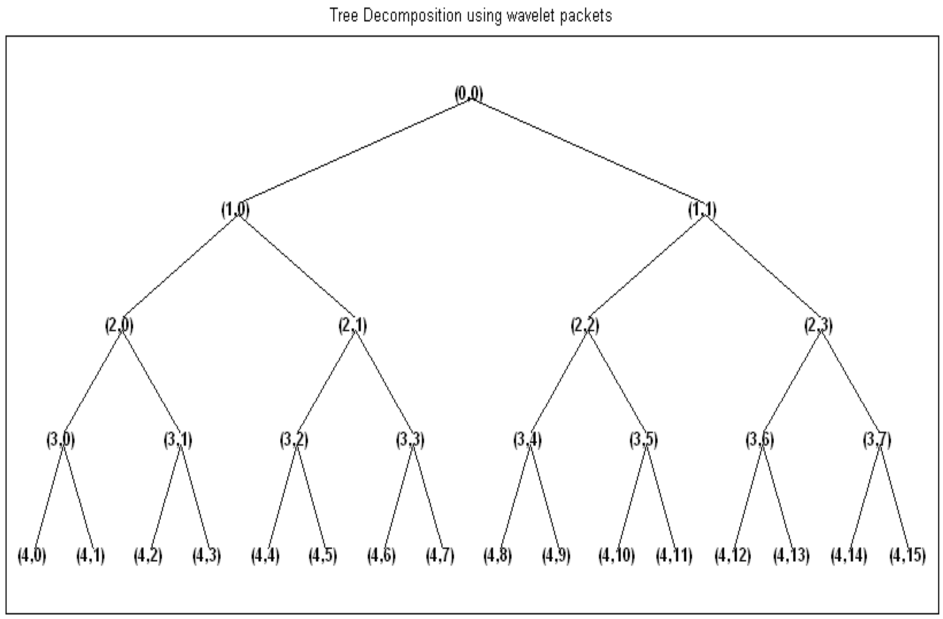

Fig. 3.7: Wavelet decomposition of transmitted OFDM data

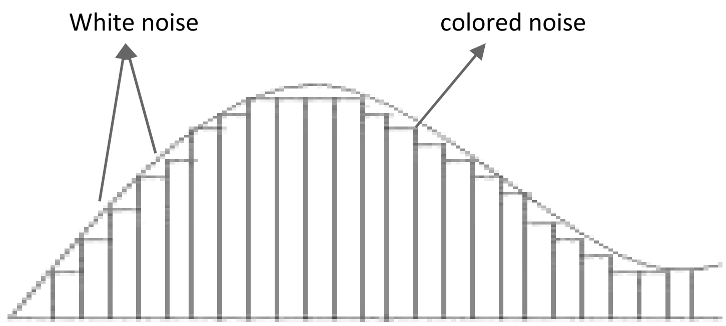

Fig. 3.8: Estimation of colored noise using white noise in small segments

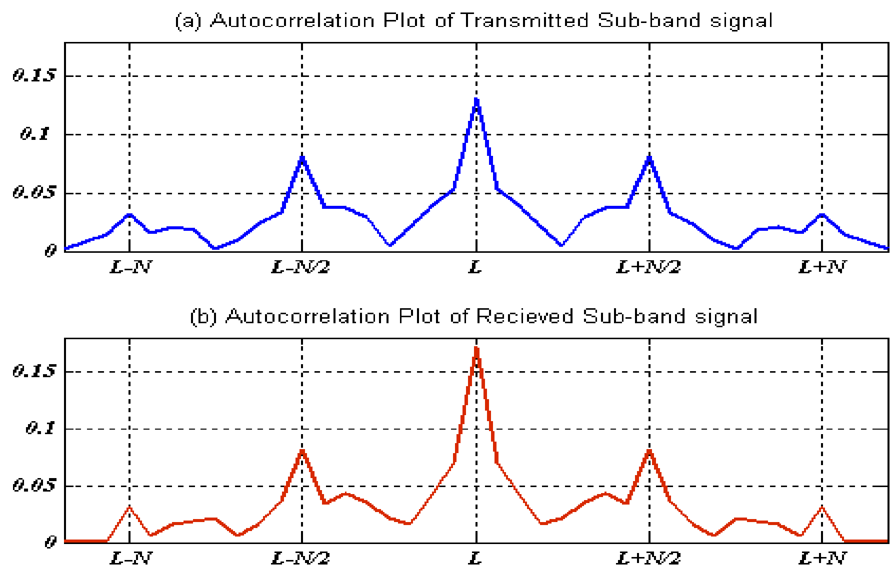

Fig.3.9 (a): Autocorrelation of transmitted signal. (b): Autocorrelation of received signal

b. Signal Power Estimation in each Sub-band

The Explanation for the correlation peaks of Fig.3.9 is same as shown in Fig.3.6 because sub-bands inherit the two identical halves property of synchronization preamble. After wavelet packet decomposition, the length of data and the length of CP are changed. So after correlation of each sub-band, first peak rises at L_{sub}-N_{sub} when CP matches with itself with energy of $\rho.\ l_{CPsub}$. The second peak rises at L_{sub}-$N_{sub}/2$ when one half of sub-band plus CP_{sub} matches with itself with energy of $\rho.\ (N_{sub}/2+\ l_{CPsub})$ and main peak at zero-lag (L_{sub}) rises when full sub-band matches with itself with energy of $\rho.(N_{sub}+\ l_{CPsub})$.

Taking into consideration the autocorrelation values for L_{sub}-$N_{sub}/2$ and L_{sub}-N_{sub} lags or L_{sub}+$N_{sub}/2$ and L_{sub}+N_{sub}, signal power is given as

$$\hat{P}_{sub} = 2R_{yy}(L_{sub} - N_{sub}/2) - R_{yy}(L_{sub} - N_{sub}) \tag{3.24}$$

Or

$$\hat{P}_{sub} = 2R_{yy}(L_{sub} + N_{sub}/2) - R_{yy}(L_{sub} + N_{sub}) \tag{3.25}$$

where $\hat{P}_{sub}$ is the estimated signal power of each sub-band.

c. Noise power Estimation in each Sub-band

Having obtained the power of signal in certain sub-band, noise power can be calculated as

$$\text{Noise Power} = \hat{\sigma}_N^2 = R_{yy}(L_{sub}) - \hat{P}_{sub} \tag{3.26}$$

where $R_{yy}(L_{sub})$ is value at zero-lag.

d. SNR Estimation in each Sub-band

Finally we can find the SNR estimates in the sub-band by using equation (3.24 or 3.25) and equation (3.26).

$$\hat{SNR} = \frac{\hat{P}_{sub}}{\hat{\sigma}_N^2} \tag{3.27}$$

where $\hat{SNR}$ is the estimated value for SNR.

Ideally, signal power and noise power are calculated without CP for the original data of length N. Table 3.2 shows that signal power and noise power calculated for the proposed method is same as ideal case because the energy contained in CP is subtracted from the energy contained by total signal which is data plus CP.

Table 3.2: Ideal vs. calculated SNR for second-part of proposed technique

	(Ideal)	(Calculated)
Signal power= P_{ss}	$\rho.N_{sub}$	$P_{ss} = \rho\{2(N_{sub} + l_{subCP})\} - \rho.l_{subCP} = \rho.N_{sub}$
Noise Power= P_{NN}	$\hat{\sigma}_N^2.N_{sub}$	$P_{NN} = \hat{\sigma}_N^2(N_{sub} + l_{subCP}) - \hat{\sigma}_N^2.l_{subCP} = \hat{\sigma}_N^2.N_{sub}$
SNR_{sub}	P_{ss}/P_{NN}	P_{ss}/P_{NN}

The methodology and parameters to perform simulation of proposed front-end based SNR estimation technique is discussed in the chapter 4. The proposed SNR estimation technique is of In-service type SNR estimator. There is no throughput penalty as the proposed technique makes use of synchronization preamble which is already employed in OFDM systems and another advantage of the proposed technique is that it generally works with both white noise as well as colored noise scenario.

According to the best knowledge of author this is the first SNR estimation technique for multicarrier systems like OFDM systems which performs SNR estimation at the front-end of the receiver. Previously, there is only one SNR estimation technique which performs SNR estimation at the front-end of the receiver proposed by Nidal [Nidal, 2007] for single carriers systems using AWGN channel.

3.3 Methodology for Analyzing Various SNR Estimators

In the last section we discussed the formulation of proposed SNR estimation in OFDM systems. The proposed SNR estimator performed SNR estimation at the front-end of the receiver unlike Reddy SNR estimator and subspace SNR estimator discusses in chapter 2 and used later for comparison. The completion of estimator's formulation now set the stage to define the methodology and parameters of developed SNR estimators to perform the simulations. The performance results when using the proposed SNR estimation technique will be obtained and discussed in the next chapter.

The methodology and the parameters needed for the simulations of the SNR estimators (Reddy estimator, Subspace estimator & Proposed front-end estimator) used in this thesis will be discussed in the next section.

3.3.1 Methodology for Analyzing Reddy's SNR Estimator

The methodology of the Reddy's estimator developed in Matlab® is depicted in Fig.3.10. An OFDM system with 256 sub-carriers is considered as shown in Fig.3.11. After the addition the cyclic prefix of length 64, IFFT is performed and the signal is then passed through the channel. At the receiver side cyclic prefix is removed and FFT is performed to convert the signal back into its original transform. SNR estimation is performed after the FFT process. Other parameters for this technique are shown in Table 3.3. The flow chart in Fig.3.12 shows how this technique works to get the SNR estimates. 8 Pilot (x-pilot) symbols are inserted in each OFDM block transmission. Channel frequency response (Hl_s) is computed from the pilot symbols for the transmitted signal (x-pilot) and the received signal (y-pilot) as discussed in the last chapter. Each of these pilot symbols are vectors of size 8x1, which is the number of pilots inserted in the OFDM symbol.

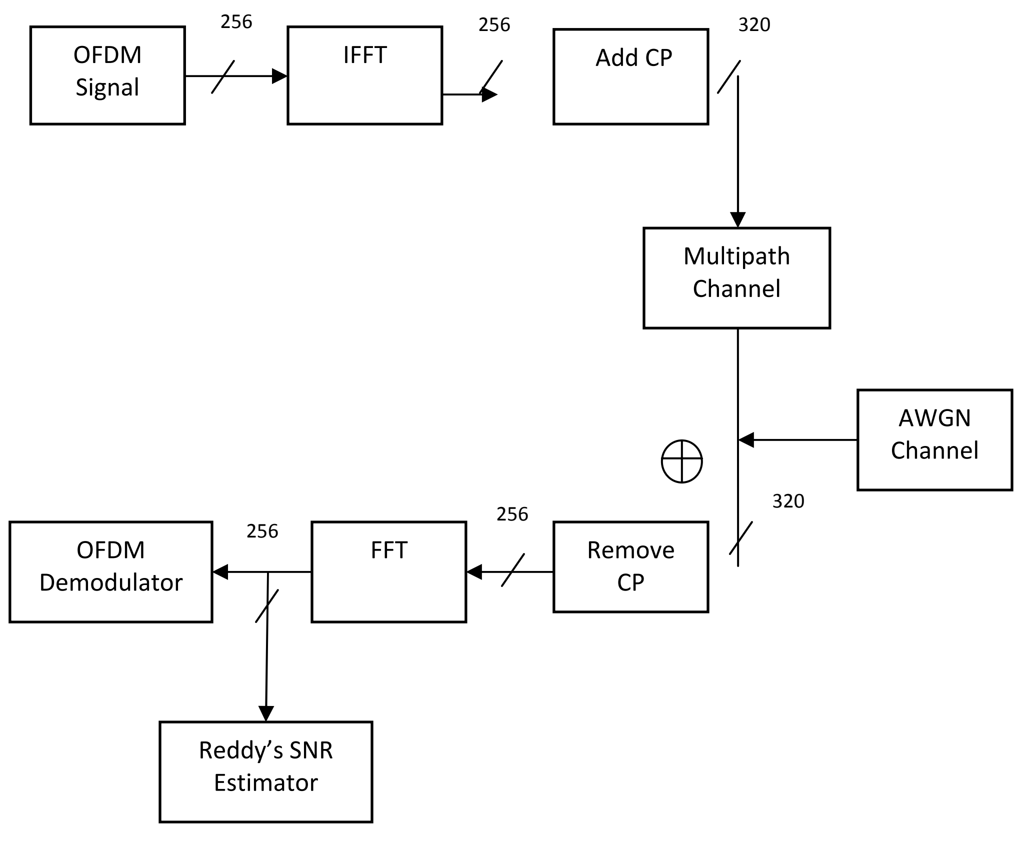

Fig. 3.10 Methodology of Reddy's SNR Estimator

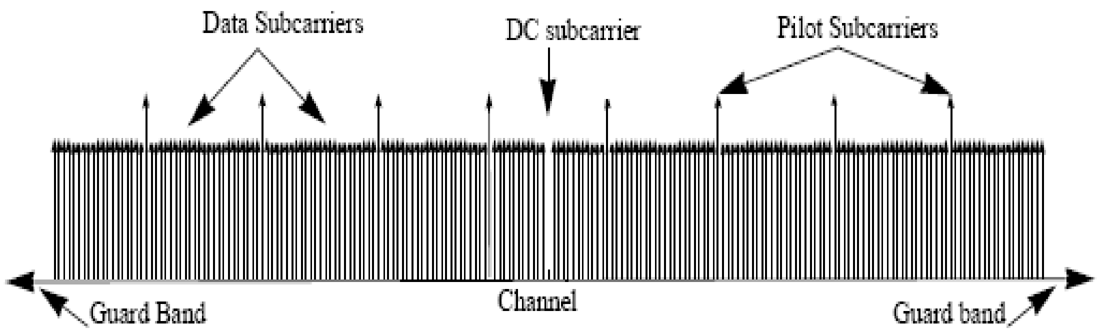

Fig. 3.11 OFDM Signal in frequency domain (IEEE802.16d standard)

Table3.3. Parameters for OFDM Systems Simulation

N_{fft} size (N)	*256*
N_{used} (*data carrier = 192 & Pilot Carrier = 8*)	*200*
Sampling Frequency (Fs)	20*MHz*.
Number of Lower frequency guard subcarriers	*28*
Number of Higher frequency guard subcarriers	*27*
Subcarrier Spacing $(\Delta f = Fs/N_{fft})$	1×10^{5}
Useful Symbol Time ($Tb=1/\Delta f$)	1×10^{-5}
Guard Interval (G)	$\frac{1}{4}(N\ )$
CP Time ($Tg=G*Tb$)	2.5×10^{-6}
OFDM Symbol Time ($Ts = Tb + Tg$)	1.25×10^{-5}

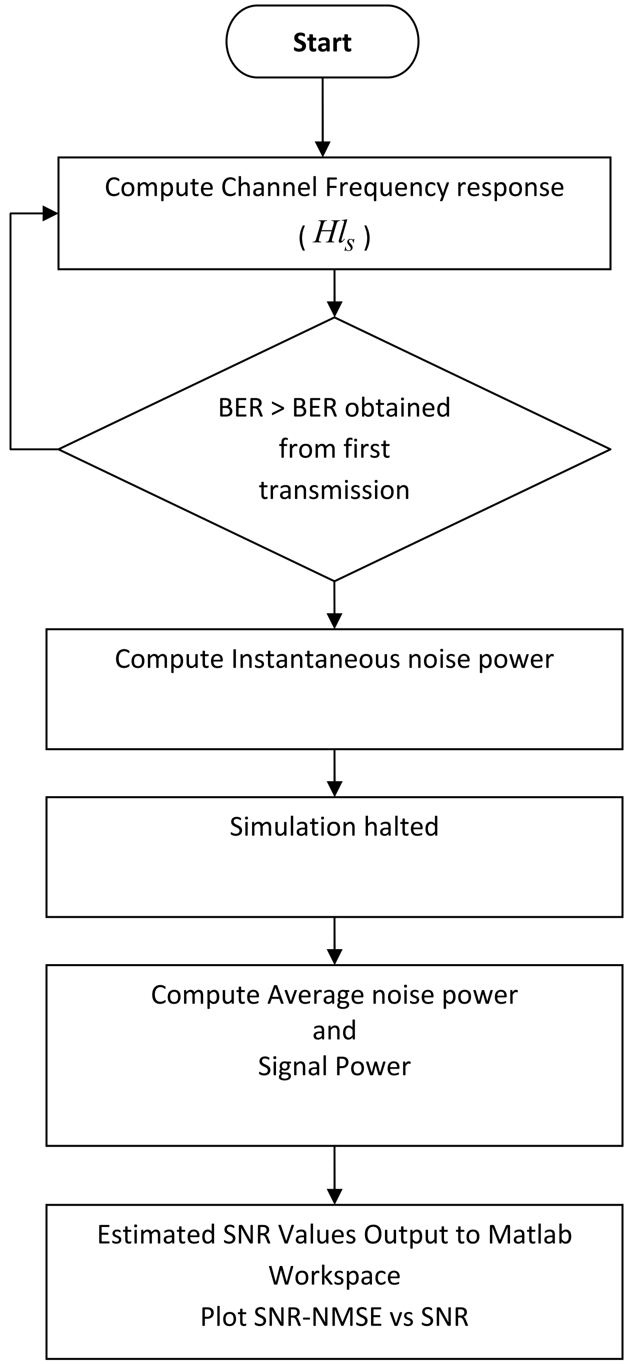

Fig. 3.12 Reddy’s estimator flow chart

Simulation results of Reddy’s SNR estimator for (30 OFDM symbols) are used for comparison with proposed SNR estimator and will be shown in the next results & discussion chapter. The comparison is performed in terms of normalized mean squared error (NMSE) and estimated SNR.

3.3.2 Methodology for Analyzing Subspace Based SNR Estimator

The methodology of the subspace base estimator developed in Matlab® is depicted in Fig.3.13. An OFDM system with the same parameters as discussed in Reddy's estimator is considered. Subspace based estimator also performs the SNR estimation at the back end of the receiver. The flow chart in Fig.3.14 shows how this technique works to get the SNR estimates. 8 Pilot (x-pilot) symbols are inserted in each OFDM block transmission. Subspace based estimator accepts the same two inputs (x-pilot & y-pilot) as the Reddy's estimator, which is used to compute the moving average correlation matrix estimate. The window size of the moving average, which is specified by K in eq.2.26 for minimum descriptive length (MDL) criteria is set to be equal to 10 in the results obtained as this is the suggested size by [Xiaodong et al, 2005]. The subspace estimator then continues to

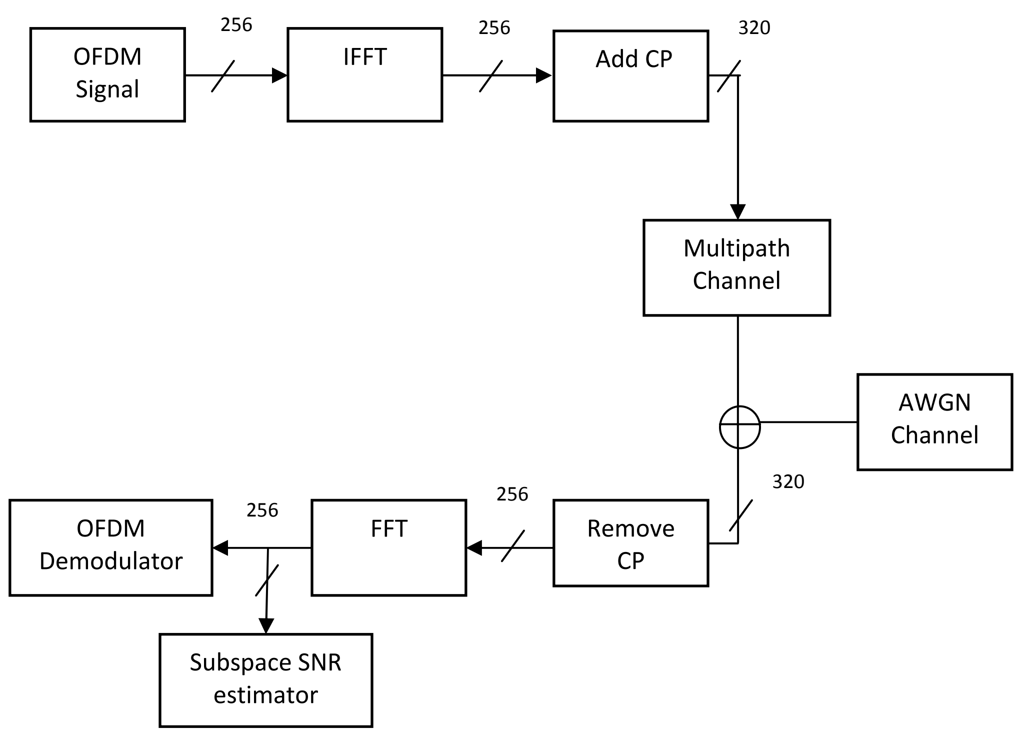

Fig.3.13 Methodology of subspace based SNR estimator

compute the eigenvalues which are used to compute total estimated number of multipath (L). This value of signal path which is calculated to be one for AWGN channel is used to compute the channel power using the eigenvalues. The remaining M-L eigenvalues are used to compute the noise power.

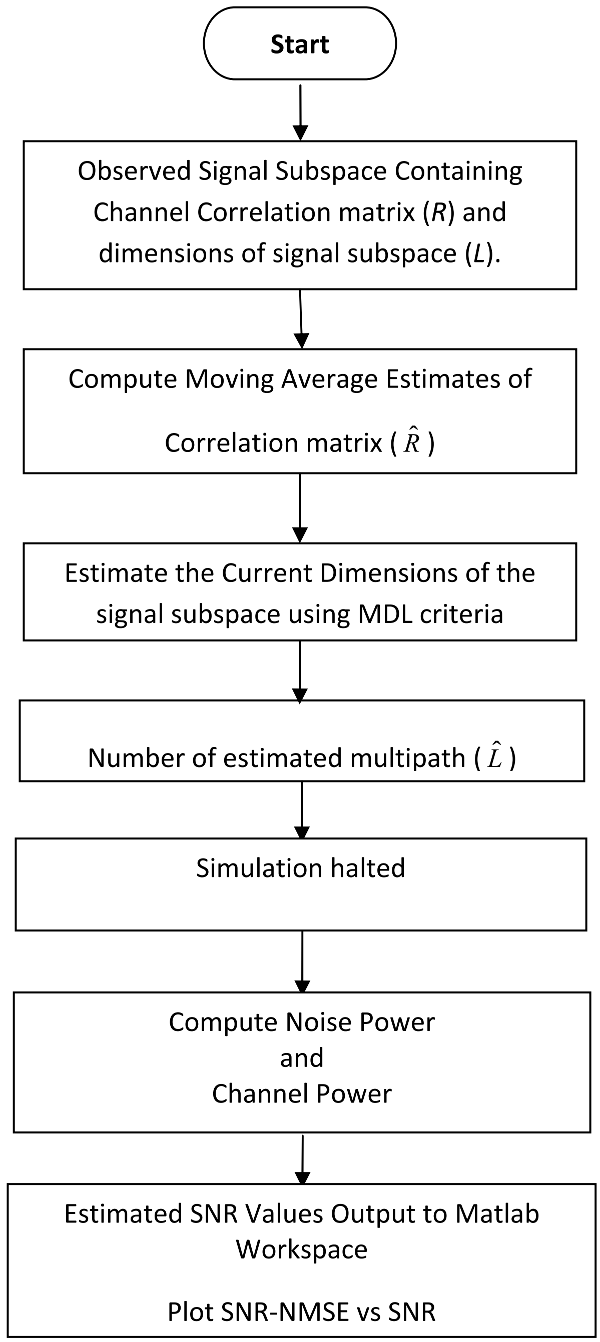

Fig. 3.14 Flow chart of subspace estimator

Simulation results of subspace based SNR estimator for (30 OFDM symbols) are used for comparison with proposed SNR estimator and will be shown in the next results & discussion chapter. The comparison is performed in terms of normalized mean squared error (NMSE) and estimated SNR.

3.3.3 Methodology for Analyzing Proposed Front-End SNR Estimator

The methodology of proposed front-end SNR estimator is divided in to two parts. In the first part methodology of proposed SNR estimation technique for white noise is discussed. The proposed technique is extended for SNR estimation of colored noise using wavelet packet and methodology of SNR estimator for colored noise is discussed in second part.

3.3.3.1: First Part: For Multipath Channels With AWGN

The methodology of proposed SNR estimator is shown in Fig.3.15. An OFDM based system is considered with parameters as shown in Table 3.4 and Table 3.5. The proposed estimator is based on one OFDM preamble signal and performed SNR estimation at the front-end of the receiver unlike Reddy estimator and subspace estimator. The synchronization OFDM preamble-the preamble which has two identical halves property as shown in Fig.3.16, is obtained by loading constellation (QPSK) points with a PN sequence (P_{seq}) at even sub-carriers. For simulations, the parameters are calculated using WiMAX standard (IEEE802.16, 2004) for 256 bit ling data and Wi-Fi (IEEE802.11a) for 64 bit long data. Cyclic prefix is chosen ¼ of the original data. OFDM training data sent from the transmitter. After the addition the cyclic prefix, IFFT is performed and the signal is then passed through the channel. Autocorrelation is performed on the received signal at the front-end of the receiver. Estimates of the signal power and the noise power are estimated from the autocorrelation results. The flow chart in Fig.3.17 shows how this technique works to get the SNR estimates.

Simulation results of proposed front-end SNR estimator for AWGN and multipath channels (Rayleigh, Rician, SUI channel and indoor channel models), using only one OFDM preamble, are used for comparison with Reddy's estimator and subspace based SNR estimator and will be shown in the next results & discussion chapter. The comparison is performed in terms of normalized mean squared error (NMSE) and estimated SNR.

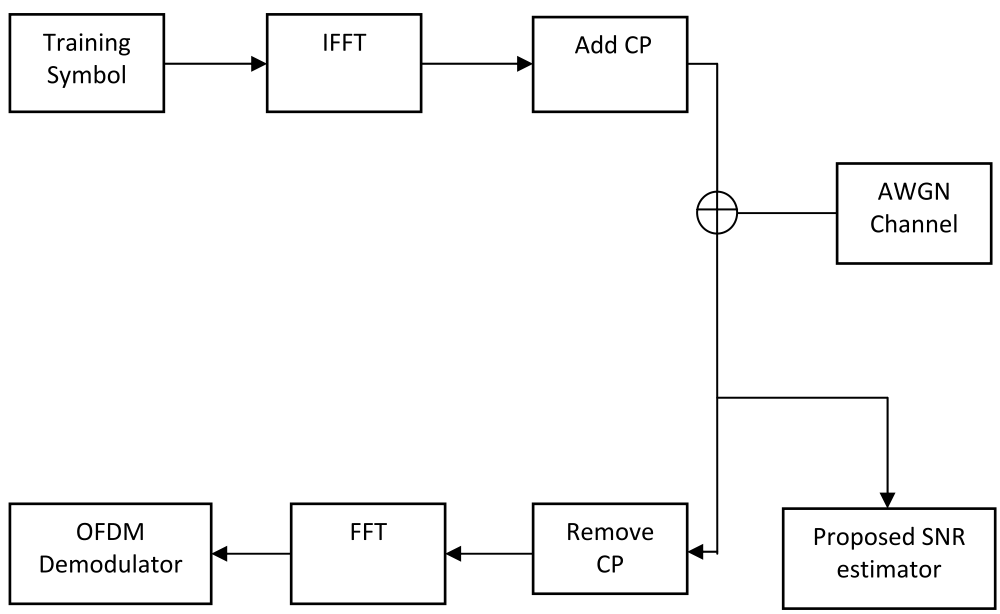

Fig 3.15: Methodology of first part of proposed technique.

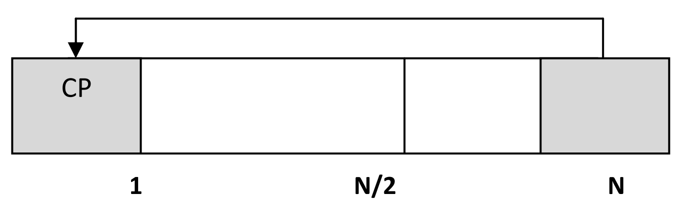

Fig. 3.16: OFDM Preamble symbol with cyclic prefix.

Table 3.4. Parameters of proposed Technique for first part (IEEE802.16, 2004)

N_{fft} size (N)	*256*
N_{used}	*200*
Sampling Frequency (Fs)	20MHz.
Number of Lower frequency guard subcarriers	*28*
Number of Higher frequency guard subcarriers	*27*
Subcarrier Spacing (Δf $=Fs/N_{fft}$)	1×10^{5}
Useful Symbol Time ($Tb=1/\Delta f$)	1×10^{-5}
Guard Interval (G)	$\frac{1}{4}(N\ )$
CP Time ($Tg=G*Tb$)	2.5×10^{-6}
OFDM Symbol Time ($Ts = Tb + Tg$)	1.25×10^{-5}

Table 3.5. Parameters of proposed Technique for first part (IEEE802.11a)

N_{fft} size (N)	*64*
N_{used}	*52*
Sampling Frequency (Fs)	20MHz.
Subcarrier Spacing (Δf $=Fs/N_{fft}$)	0.3125×10^{6}
IFFT period ($Tb=1/\Delta f$)	3.2×10^{-6}
Guard Interval (G)	$\frac{1}{4}(N\ )$
CP Time ($Tg=G*Tb$)	0.8×10^{-6}
OFDM Symbol Time ($Ts = Tb + Tg$)	4×10^{-6}

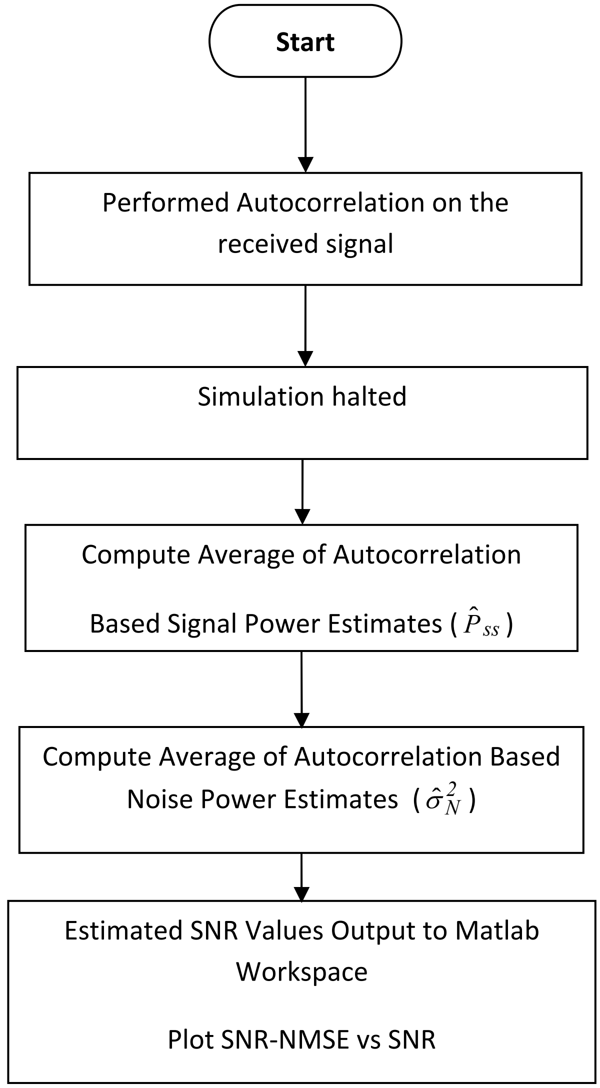

Fig.3.17 Flow chart of Proposed SNR estimation technique for white noise

3.3.3.2: Second part: For Multipath Channel With Colored Noise using Wavelet Packet Filter Banks.

The methodology of proposed SNR estimator is shown in Fig.3.18. For the second part of our proposed technique; An OFDM system, which takes into account the color and variation of noise statistics over OFDM sub-carriers, is considered. Unlike first part, the received signal at front end of the receiver, is divided into several sub-bands using wavelet packet and noise in each sub-band is considered white. Other parameters used are shown in Table 3.6. The Proposed estimator provides many local estimates, allowing tracking of the variation of the noise statistics across OFDM sub-carriers, which are particularly of use in sub-band adaptive modulation OFDM systems. Autocorrelation is performed on each received sub-band signal at the front-end of the receiver. Estimates of the signal power and the noise power in each sub-band are estimated from the autocorrelation results of each sub-band. Then SNR is estimated within each of the sub-band (local estimates of SNR values) and these local SNR values are averaged over all OFDM data to get global estimates of the SNR. The flow chart in Fig.3.19 shows how this technique works to get the SNR estimates.

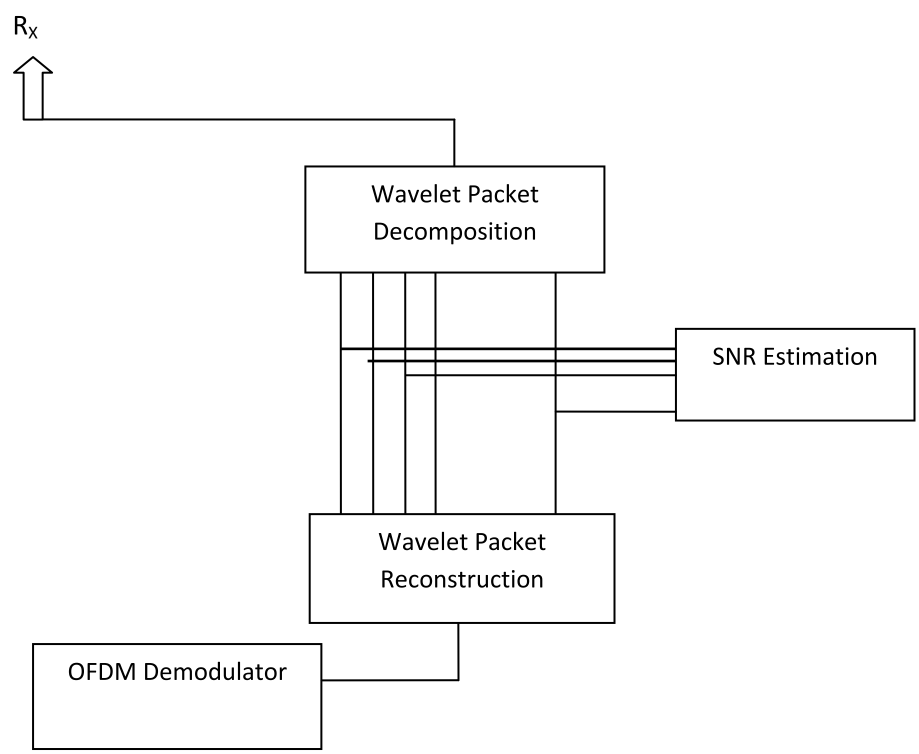

Fig.4.18. Methodology of Proposed SNR estimation technique for colored noise

Table 3.6: Parameters for Second-Part of proposed technique

I_{fft} size	*256*
N_{used} (*preamble on even sub-carriers*)	*200*
Sampling Frequency (Fs)	20MHz.
Number of Lower frequency guard subcarriers	*28*
Number of Higher frequency guard subcarriers	*27*
CP Time $= T_g = G * T_b$ where $G = 1/4$	2.5×10^{-6}
OFDM Symbol Time $= T_s = T_b + T_g$	1.25×10^{-5}
$T_s = 5/4 * T_s$ (Because ¼ CP makes the sampling faster by 5/4 times)	1.5625×10^{-5}
$T_{sub} = T_s / 16$	9.7656×10^{-7}
Wavelet Packet Object Structure *Wavelet Decomposition Command : wpt = wpdec(data,4,'db3') ,* *Size of initial data : [1 320]* *Order : 2* *Depth : 4* *Terminal nodes : [15 16 17 18 19 20 21 22 23 24 25 26 27 28 29 30]* --- *Wavelet Name : db3* *Entropy Name : Shannon*	

Simulation results of proposed front-end SNR estimator for colored noise (using only one OFDM preamble) is used for comparison with Reddy's estimator and will be shown in the next results & discussion chapter. The comparison is performed in terms of normalized mean squared error (NMSE) and estimated SNR.

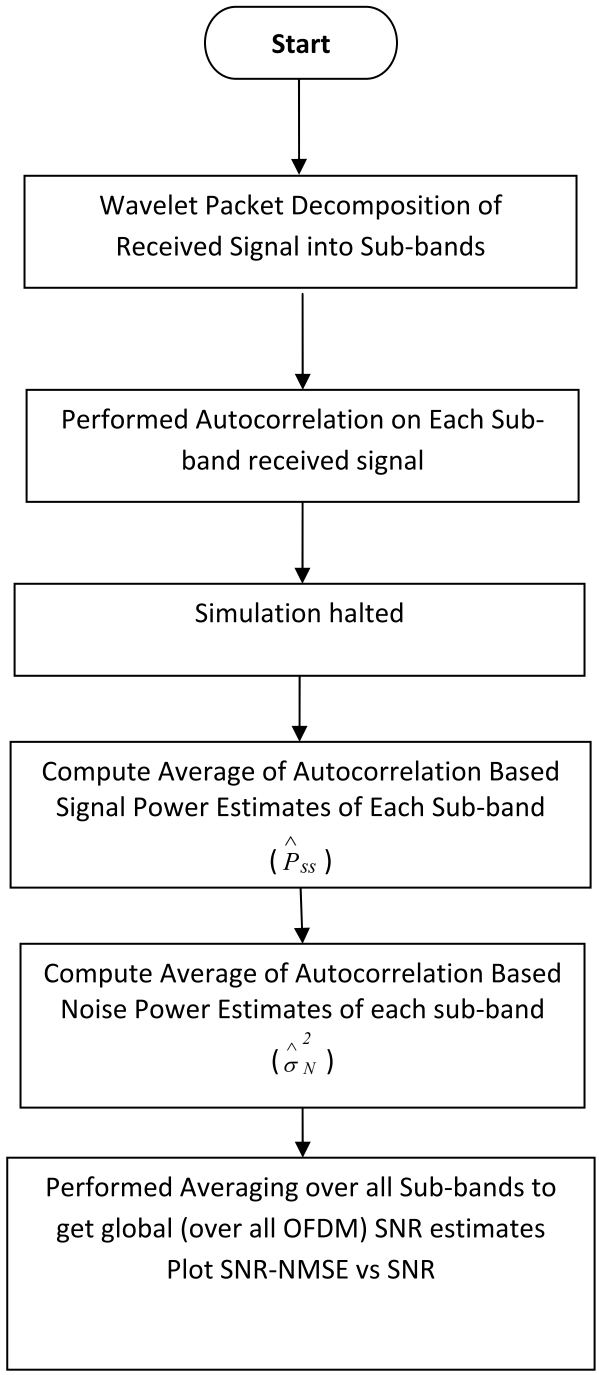

Fig.3.19 Flow chart of Proposed Front-End SNR estimator for Colored noise

3.4 Summary

In this chapter, formulation of Reddy estimator, sub-space estimator and our proposed front-end noise power and SNR estimator technique for OFDM wireless systems is presented. Reddy SNR estimator and subspace based SNR estimator are back end estimators unlike proposed SNR estimator in which SNR estimation is performed at front-end of the receiver. In the first part of proposed technique noise is assumed to be white and SNR estimation is done over all OFDM symbol. In the second part the assumption of the noise to be white is removed. Also, variation of the noise power across OFDM sub-carriers is allowed. Therefore, the proposed approach estimates both local (within smaller sets of subcarriers) and global (over all sub- carriers) SNR values. The short term local estimates calculate the noise power variation across OFDM sub-carriers. These estimates are specifically very useful for adaptive modulation, and optimal soft value calculation for improving channel decoder performance. Methodology and the parameters of Reddy estimator, Sub-space estimator and proposed front-end SNR estimator technique for OFDM wireless systems are presented. Reddy SNR estimator and subspace based SNR estimator are back end estimators unlike proposed SNR estimator in which SNR estimation is performed at front-end of the receiver. Reddy and subspace makes use of pilot symbol for their SNR estimation technique unlike proposed SNR estimator. All estimators discussed in this chapter are using the same parameters to provide a fare comparison of proposed SNR technique. To show the generality of proposed technique, designed methodology is also checked with parameters of Wi-Fi (IEEE802.11a) and WiMAX (IEEE802.16, 2004).

In Appendix B (FFT based on Wavelet Packet and its application to SNR estimation), it can be shown that, if FFT is built using Wavelet Packet algorithm, then for no extra cost, SNR estimates can be obtained inside FFT block after CP has been removed.