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Novel SNR Estimation Teachnique In Wireless OFDM Systems

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Abstract

A novel front-end noise power and SNR estimation technique based on one OFDM preamble is proposed and compared with previously published SNR estimators – none of which are front-end estimators. This paper is extended and expanded version of our previous work. It provides complete mathematics and discussion of designed algorithms for front end SNR estimation. The proposed technique is divided into two parts. In the first part, SNR estimation technique for AWGN channel and wireless multipath channels is considered. In the second part, the proposed estimator takes into consideration the different noise power levels over the OFDM sub-carriers. The OFDM band is divided into several sub-bands using wavelet packet decomposition 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 noise power is estimated. The proposed estimator in first part is compared with Reddy estimator and Subspace based estimator for AWGN channel in terms of normalized mean square error and in the second part it is compared with Reddy's estimator for colored noise in terms of mean square error (MSE). It is observed that current estimator gives better SNR estimates than Reddy and subspace estimators and can estimate local statistics of the noise power when the noise is colored.

Keywords: *Noise power estimator, SNR estimation, Adaptive modulation, OFDM*

1. Introduction

Noise variance and hence SNR estimates of the received signal are very important parameters for the channel quality control in communication systems [1]. The search for a good SNR estimation technique is motivated by the fact that various algorithms require knowledge of the SNR for optimal performance. For instance, in OFDM systems, SNR estimation is used for power control, adaptive coding and modulation, turbo decoding etc. [1]-[4].

Many SNR estimation algorithms have been suggested in the last ten years [5], [6], [7] and successfully implemented in OFDM systems at the back-end of receiver using the system pilot symbols. Nidal et al. in [8] proposed linear prediction based SNR estimation for AWGN channel at front-end for single carrier systems.

Many SNR estimators in digital communication channels have been proposed over the last few decades [9]. 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 [10], [11]. There are several other SNR measurement methods which can be found in [12] and reference listed therein.

According to best knowledge of author there is no work done at front end of the receiver for multicarrier systems. 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 to extend 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 [20, 21]. But, in wireless communication systems, where noise is mainly caused by a strong interferer, noise is colored in nature.

This paper is expanded version of our previous work in which a front-end noise power and SNR estimator for the white noise as well as for colored noise in OFDM system is proposed [18, 19]. The algorithm is based on the two-identical-halves property of time synchronization preamble used in some OFDM systems. The proposed technique is divided into two parts. First part explains SNR estimation technique for AWGN channel. In the second part, the proposed estimator is takes into consideration the different noise power levels over the OFDM sub-carriers. The OFDM band is divided into several sub-bands using wavelet packet. The second-order statistics of the transmitted OFDM preamble are calculated in each sub-band and the noise power is estimated. Therefore, the proposed approach estimates both local (within smaller sets of subcarriers) and global (over all sub- carriers) SNR values.

The remainder of the paper is organized as follows. Section 2 describes the SNR estimation algorithms used to compare with the proposed SNR estimation technique. In Section 3 we discuss the formulation of proposed technique. In section 4 we discussed the results of the proposed techniques. Section 5 concludes the paper.

2. SNR Estimation Algorithms

The two back-end SNR estimators, the Reddy estimator [5], and subspace based estimator [1], are used for comparison in this work and discussed below.

2.1 Reddy SNR Estimation

In this method, channel estimation is performed in the first realization of the channel using pilot symbols, and this estimate is used to estimate the signal noise power. The suggested method can be used for an Additive White Gaussian Noise (AWGN) channel and for color dominated channel, in which the noise power varies across the frequency spectrum.

The system model is described in the frequency domain, where a signal is transmitted to obtain the estimated channel frequency response after which the instantaneous noise power mean square is determined. The transmitted signal includes white noise which is added by the channel. This is modeled in the frequency domain by the equation:

$$Y_m(k) = X_m(k)H_m(k)+W_m(k) \tag{1}$$

where

$X_m(k)$ = Transmitted signal

$Y_m(k)$ = Received signal

$W_m(k)$ = Channel white noise

The channel frequency response is estimated by transmitting preamble and performing division in the frequency domain of the received signal by the transmitted signal. When performing the division, the effect of noise is ignored. The pilot symbols are then used as the transmitted signal and the received signal in the pilot sub-carriers is used for the received signal and the estimated transfer function inserted in the equation to determine the noise power estimate. The noise power estimation is measured by finding the difference between the noisy received signal and the noiseless signal.

$$E_m(k) = \left| Y_m(k) \hat{X}_m(k) + \hat{H}_m(k) \right|^2 \tag{2}$$

The difference between the actual and estimated channel frequency response is the channel estimation error.

In this technique, noise power estimates for overall OFDM subcarriers as well as noise power estimates for noise variation within the transmission bandwidth (colored noise) are derived. In this work averaging across several OFDM symbols as well as averaging across OFDM subcarriers is proposed for reliable SNR estimates. For the noise power estimation of colored noise, the whole OFDM data is divided (i.e. the total number of sub-carriers) into sub-bands (i.e. to a set of subcarriers). If the number of sub-carriers in each sub-band is k, then the number of sub-bands will be N/k. Then, the absolute square of the instantaneous noise estimates in each sub-band are averaged,

$$\hat{\sigma}^2_{N_m}(j) = \frac{1}{k}\sum_{l=1}^{k} \hat{\sigma}^2_{N_m}(l) \qquad 1 \le j \le N/k \tag{3}$$

where $\hat{\sigma}^2_{N_m}(j)$ is the estimated noise power in the j^{th} sub-band.

Using the knowledge of channel estimates, signal power over each sub-band is estimated as

$$\hat{P}_s(j) = \frac{1}{k}\sum_{l=1}^{k} \left| Y_m(l) \right|^2 \tag{4}$$

where $\hat{P}_S(j)$ is the estimated signal power in the j^{th} sub-band.

Having knowledge of noise power estimates and signal power estimates in each sub-band, the SNR is computed as

$$S\hat{N}R(j) = \frac{\hat{P}_s(j)}{\hat{\sigma}^2_{N_m}(j)} \tag{5}$$

where $S\hat{N}R(j)$ is the estimated value of actual SNR in the j^{th} sub-band. In this work averaging across several OFDM symbols as well as averaging across OFDM subcarriers is proposed for reliable SNR estimates.

2.2. Subspace Based SNR Estimation

The second algorithm uses statistical analysis and represents the channel model in terms of a subspace defined by the number of propagation multipaths, which is the dimension of the observation vector that satisfies the Minimum Descriptive Length (MDL) criteria described in [13]. Each of the paths (L) is modeled as a Gaussian process with varying time delays (τ) and path gains, expressed by the equation for channel impulse response:

$$h(t,\tau) = \sum_{l=1}^{L} h_l(t)\delta(t-\tau_l) \tag{6}$$

The observation vector is defined as the signal received which is modeled as the maximum number of superimposed multipath signals (k). The MDL criteria is used to search for estimate of the number of multipath (L) by finding the value of k that minimizes the MDL, which is the partitioning of the observation vector into subspace [13]. The correlation matrix of observation vector is decomposed into its eigenvalues and eigenvectors, where the correlation matrix (R) is

$$R = \mathrm{W}_p E(\mathrm{h}_l \cdot \mathrm{h}_l^H)\mathrm{W}_p^H \tag{7}$$

where

W_p = FFT matrix of pilot symbols

$\mathbf{h}_l$ = channel impulse response for l-th multipath

$(\cdot)^H$ = Hermitian conjugate matrix transpose

P = pilot symbol index $P \in \{1, 2\ldots M\}$

M = number of pilot symbols

The eigenvalues of R when arranged in descending order of magnitude give an indication of the subspace. The smallest $M-L$ eigenvalues are equal to the noise variance. The estimation of the correlation matrix is done using the channel frequency response by assuming a noiseless channel and averaging it across K OFDM symbols. The search for L is done by performing iterations of the MDL function:

$$MDL(k) = -K(M-k)\log\left[\frac{\prod_{i=k+1}^{M}\hat{\lambda}_i^{\,1/(M-k)}}{{}^{1}\!/_{M}-k\sum_{i=k+1}^{M}\hat{\lambda}_i}\right] + {}^{1}\!/_{2}\,k(2M-k)\log(K) \tag{8}$$

The channel is characterized as a wide sense stationary uncorrelated process, which describes the random characteristics of the channel in which the first and second moments do not change with time. The probability density functions of the noise in each of the multipaths are uncorrelated because each of the paths is independent.

The assumptions made are that the channel has a guard interval which is greater than the channel delay spread and that the channel is quasi- stationary. This refers to the assumption made that the noise does not change within each OFDM symbol. This can also be extended to a few consecutive OFDM symbols, therefore assuming a constant noise in a block of symbols.

The signal to noise ratio (SNR) during the i^{th} OFDM symbol is:

$$\rho = \frac{\sum_{i=1}^{L}\left|h_l(iT_s)\right|^2}{\sigma_N^2} \tag{9}$$

where

σ_N^2 = noise variance

$\sum_{i=1}^{L} |h_l(iT_s)|^2$ = Channel power

The transmitted channel power is assumed to be unity and this is scaled down by the channel. This estimator makes use of minimum 20 OFDM symbols to provide better SNR estimates.

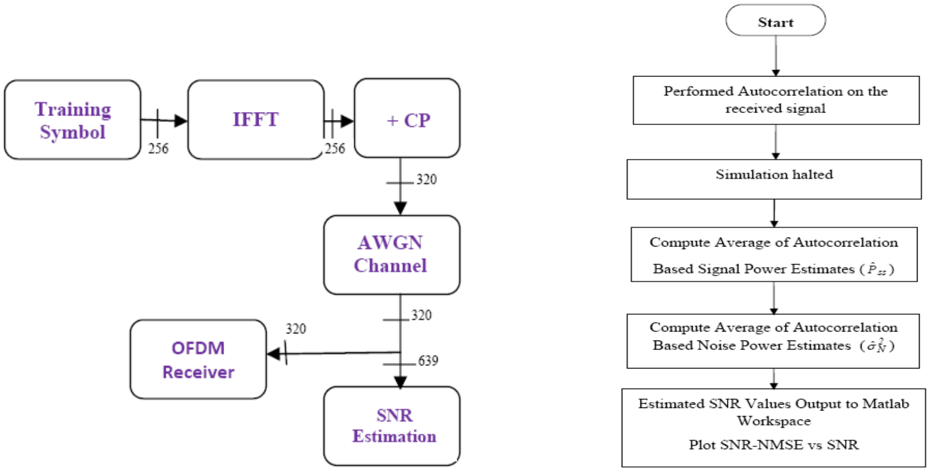
3. Proposed SNR Estimation Technique

The proposed front-end SNR estimator utilizes only one OFDM symbol to give better SNR estimates. The estimator makes use of two-identical-half property of time synchronization preamble used in OFDM systems and proposed by Schmidl et, al. [14]. The proposed technique is divided into two parts. The first part is the front-end SNR estimation technique for AWGN channel and multipath channels (Rayleigh & Rician) using white noise scenario. The second part is the extension of first part for noise power estimation of colored noise using wavelet-packet based filter bank analysis of the noise.

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

The block diagram of the OFDM system is depicted in the Fig 1. The synchronization OFDM preamble-the preamble which has two-identical-halves property, used for timing synchronization is derived from alternate loading of subcarriers with PN-sequence modulated constellation as follows:

$$P_{even}(k) = \begin{cases} \sqrt{2}P(m) & k = 2m \quad m = i,2 \ldots N/2 \\ 0 & otherwise \end{cases} \tag{10}$$



(a) (b)

Fig.1 (a) Block Diagram of the Proposed System (b) Flow Chart of Proposed System

Here, $P(m)$ is the PN sequence loaded onto even subcarriers taken from IEEE802.16d standard [15]. 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 wide, 28 carriers on either side are null carriers.

Therefore for our purpose (10) is rewritten as

$$P_{even}(k) = \begin{cases} \sqrt{2}P(m) & k = 2m\,;\ m = 15, 16,..(N/2 - 14) \\ 0 & m = 1,2.3\ldots14 \\ 0 & m = N/2 - 13, N/2 - 12\ldots N/2 \end{cases} \tag{11}$$

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

$$\begin{aligned} p(n) &= IDFT\left\{P_{even}(k)\right\} \\ &= \sum_{k=0}^{N-1} P_{even}(k).e^{j2\pi nk/N} \qquad 0 \le n \le N - 1 \end{aligned} \tag{12}$$

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 $k = 2m$,

$$e^{j2\pi n2m/N} = e^{j2\pi nm/N/2} \tag{13}$$

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

$$\begin{aligned} e^{j2\pi(n+N/2)m/N/2} &= e^{j2\pi nm/N/2}.e^{j2\pi m.N/2/N/2} \\ &= e^{j2\pi nm/N/2} \end{aligned} \tag{14}$$

In other words

$$p(n) = p(n + N/2) \tag{15}$$

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:

$$h(t,\tau) = \sum_{l=1}^{L} h_l(t)\delta(t - \tau_l) \tag{16}$$

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

$$y(n) = x(n) + w(n) \tag{17}$$

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.

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

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

3.1.1 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 (18), 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, $R_{nn}(m)$:

$$R_{yy}(m) = R_{xx}(m) + R_{nn}(m) \tag{18}$$

where

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

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

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

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

$$R_{ww}(m) = \sigma^2\delta(m) \tag{19}$$

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

3.1.1.1 Signal Power and Noise Power Estimation

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

$$R_{ss}(m) = P_o\delta(m) \tag{20}$$

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:

$$R_{ss}(m) = P_o\ \{\tfrac{1}{2}\delta(m - \tfrac{N}{2}) + \delta(m) + \tfrac{1}{2}\delta(m + \tfrac{N}{2})\} \tag{21}$$

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

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

$$= P_o \sum_i h(i)\, h^*(m+i) + \frac{P_o}{2} \sum_i h(i)\, h^*(m+i-N/2) + \frac{P_o}{2} \sum_i h(i)\, h^*(m+i+N/2) \qquad (22)$$

When $m = 0,$

$$R_{xx}(0) = P_o \sum_i |h(i)|^2 + \frac{P_o}{2} \sum_i |h(i)h(i-N/2)| + \frac{P_o}{2} \sum_i |h(i)h(i+N/2)| \qquad (23)$$

However, since

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

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

As explained in Fig.2 which shows the result of multiplication of $h(n)$ & $h(i-N/2)$ eq. (23) becomes

$$R_{xx}(0) = P_o \sum_i |h(i)|^2 \qquad (24)$$

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

When $m = N/2,$

Similarly as explained in Fig.2, it can be shown that

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

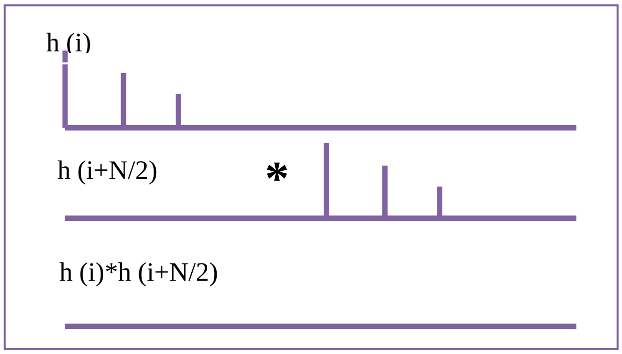


Fig.2 Multiplication of h (i) and h (i+N/2) (convolution example)

When $m=-N/2$,

$$R_{xx}(\frac{N}{2}) = P_o\sum_i \left|h(i)h(i+N/2)\right| + \frac{P_o}{2}\sum_i \left|h(i)\right|^2 + \frac{P_o}{2}\sum_i \left|h(i)h(i+N)\right| \tag{25}$$

$$R_{xx}(\frac{-N}{2}) = P_o\sum_i \left|h(i)h(i-N/2)\right| + \frac{P_o}{2}\sum_i \left|h(i)h(i-N)\right| + \frac{P_o}{2}\sum_i \left|h(i)\right|^2 \tag{26}$$

Similarly as explained in Fig.2, it can be shown that

$$R_{xx}(\frac{-N}{2}) = \frac{P_o}{2}\sum_i \left|h(i)\right|^2 \tag{27}$$

From the above equations, it is clear that the received signal power ($\frac{P_o}{2}\sum_i \left|h(i)\right|^2$) 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) the autocorrelation $R_{xx}(0)$, contains both the signal power estimate and noise power estimate indistinguishable from each other. 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. In Fig.3, $R_{xx}(m)$ has been sketched for $N=256$.

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.

a. Signal Power Estimation

In Fig.3, the autocorrelation 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 [15]. In such cases, the autocorrelation also has peaks where cyclic prefix matches with sections. This is shown in Fig.4 where the autocorrelation of the clean OFDM signal is shown in Fig.4(a) and that of the received signal with noise at SNR=7dB in Fig.4(b).

Taking into consideration the autocorrelation values for ’ $L-N/2$ ’,’ $L-N$ ’,’ $L+N/2$ ’,’ $L+N$ ’, signal power is given as:

$$\hat{P}_o = 2R_{yy}(L-N/2) - R_{yy}(L-N) \tag{28}$$

Or

$$\hat{P}_o = 2R_{yy}(L+N/2) - R_{yy}(L+N) \tag{29}$$

where $\hat{P}_0$ estimated signal power.

b. 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

Noise Power = $P_N = \hat{\sigma}_N^2 = R_{yy}(L) - \hat{P}_o$ (30)

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

c. SNR Estimation

Finally we can find the SNR estimates using (28) or (29) and (30).

$$S\hat{N}R = \frac{\hat{P}_o}{\hat{\sigma}_N^2} \tag{31}$$

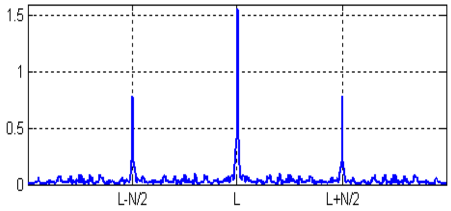
where $S\hat{N}R$ is the estimated value for SNR.

The technique presented above utilizes only one OFDM symbol to find SNR estimates unlike all previous techniques which required averaging of many OFDM symbols to get better SNR estimates. 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 1 show 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 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}

(a) Autocorrelation of Transmitted Preamble without Cyclic Prefix



(b) Autocorrelation of Received Preamble without Cyclic Prefix

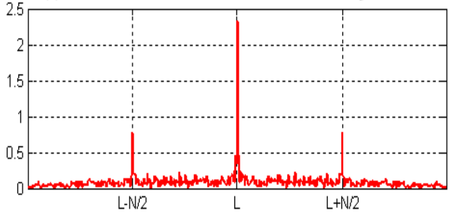
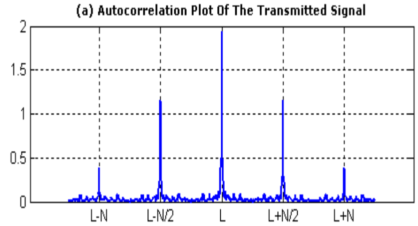


Fig.3 (a): Autocorrelation plot of transmitted signal, (b): Autocorrelation plot of received signal. (Plots show two identical halves with no cyclic prefix).



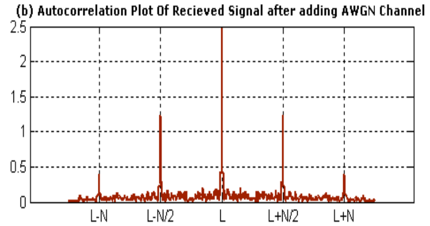


Fig 4 (a): Autocorrelation plot of transmitted signal, (b): Autocorrelation plot of received signal (Plots show two identical halves with cyclic prefix).

3.2 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.5. The colored noise in each sub-band is considered white as shown in Fig.6. 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 $l_{CPsub} = l_{CP}/2^n$.

3.2.1 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.7 (a) and Fig.7 (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 power from the zero-lag autocorrelation value.

a. Signal Power Estimation in each Sub-band

The explanation for the correlation peaks of Fig.7 is same as shown in Fig.4 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. The second peak rises at $L_{sub} - N_{sub}/2$ when one half of sub-band plus CP_{sub} matches with itself and main peak at zero-lag (L_{sub}) rises when complete sub-band matches with itself.

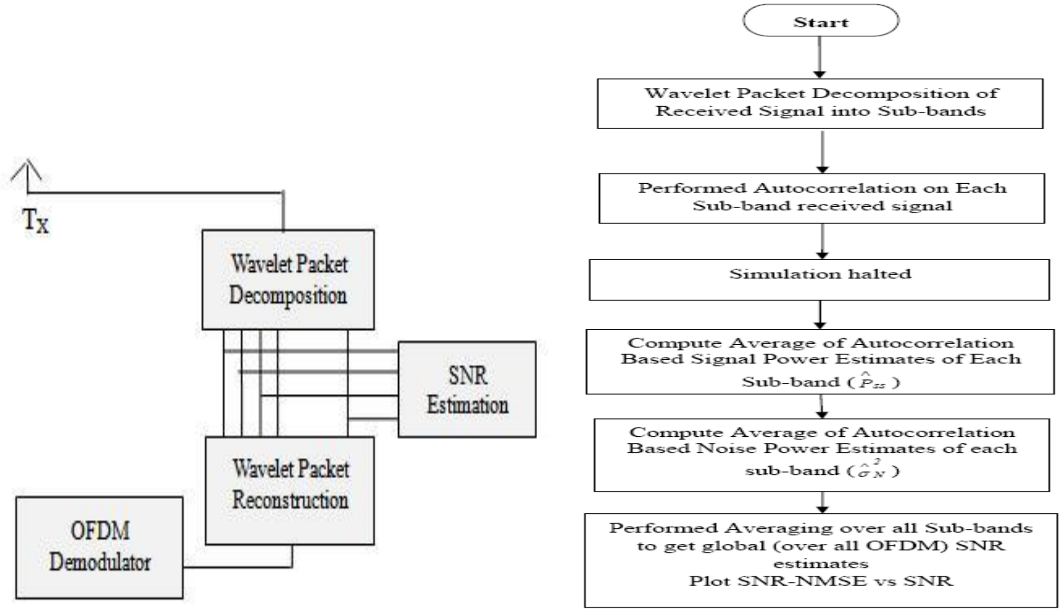
Taking into consideration the autocorrelation values for $L_{sub} - N_{sub}/2$ and $L_{sub} - N_{sub}$ lags or $L_{sub} + N_{sub}/2$ and $L_{sub} + N_{sub}$, signal power is given as

$$\hat{P}_{sub} = 2R_{yy}(L_{sub} - N_{sub}/2) - R_{yy}(L_{sub} - N_{sub}) \tag{32}$$

Or

$$\hat{P}_{sub} = 2R_{yy}(L_{sub} + N_{sub}/2) - R_{yy}(L_{sub} + N_{sub}) \tag{33}$$

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



(a) (b)

Fig. 5 Wavelet decomposition of transmitted OFDM data, (b) Flow chart of wavelet based technique

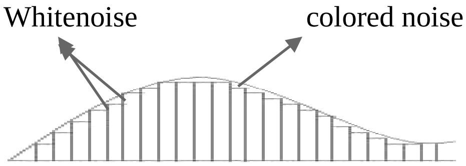
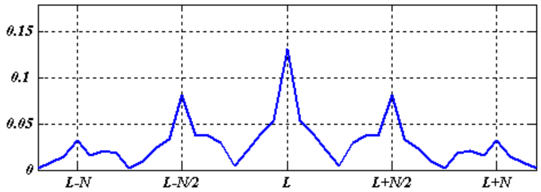


Fig. 6 Estimation of colored noise using white noise in small segments

(a) Autocorrelation Plot of Transmitted Sub-band signal



(b) Autocorrelation Plot of Recieved Sub-band signal

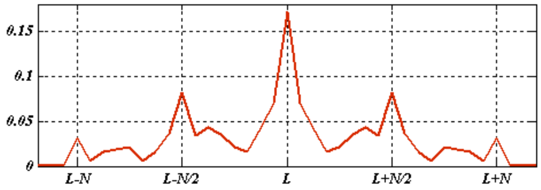


Fig.7 (a): Autocorrelation of transmitted signal. (b): Autocorrelation of received signal

b. Noise power Estimation in each Sub-band

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

$$\text{Noise Power} = \hat{\sigma}_N^2 = R_{yy}(L_{sub}) - \hat{P}_{sub} \tag{34}$$

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

c. SNR Estimation in each Sub-band

Finally we can estimate the SNR in each sub-band by using (32 or 33) and (34).

$$\hat{SNR} = \frac{\hat{P}_{sub}}{\hat{\sigma}_N^2} \tag{35}$$

where $\hat{SNR}$ is the estimated value for SNR in respective sub-band. After getting SNR estimates of all the sub-bands, averaging over all sub-bands is performed to get global SNR estimates of colored noise.

4. Results and Discussion

In the first part of proposed technique, the estimators are compared by plotting the normalized mean-squared error (NMSE) and estimated SNR for each of the estimators against the actual SNR values plotted from 2dB to 14dB. Parameters used for part first are shown in table 2. Each transmitted OFDM symbol results in a variation of the estimated SNR values and these instantaneous values are used to compute the average SNR estimate. The NMSE is computed as shown below.

$$NMSE = \frac{1}{2000}\sum_{m=1}^{2000}\left(\left(\hat{SNR}(m) - SNR\right)\Big/ SNR\right)^2 \tag{36}$$

For each SNR value 2000 iterations are performed. Mean NMSE and mean estimated SNR are computed for each SNR value. The NMSE results shows that proposed estimator with one

OFDM symbol performs better than both Reddy's and subspace estimator which gives accurate SNR estimates after averaging over many OFDM symbols. The performance of proposed estimator is compared with Reddy estimator and subspace estimator which use 30 OFDM symbols. The performance is evaluated via computer simulations using AWGN channel. The results are shown in the Figs. 8 and 9 for SNR-NMSE vs. SNRdB and Estimated SNR vs. Actual SNRdB. From Fig. 8, it is immediately apparent that the NMSE curve of proposed front-end SNR estimator is significantly lower than those of other estimators all the way over the entire considered SNR values. From Fig.9 it is very clear that the proposed estimator, with one OFDM symbol, estimates the SNR very accurately as compared to Reddy and Subspace based estimators. The Proposed estimator gives almost same estimates as actual SNR values for example at SNR= 8dB, the results for proposed, subspace based and Reddy estimator are 8.067dB, 7.44dB and 6.317dB respectively.

It is observed that the proposed estimator performs excellently with other multipath channels as well. To show this, simulations for Rayleigh and Rician multipath channels using 3-Tap, 5-Tap and 10-Tap channels are carried out. From Fig. 10 and Fig. 11 it is quite clear that the proposed estimator performs with Rayleigh and Rician multipath channels as good as with AWGN channel. It gives accurate SNR estimates (estimation error is less than 0.2 dB overall) while using different channel and more than one number of channel taps.

In the second part of proposed work, the assumption of the noise to be white is removed. Also, variation of the noise power across OFDM sub-carriers is allowed. The data is divided in to sub-bands using wavelet packets and SNR estimation is performed for each sub-band. Each sub-band data transmitted results in a variation of the estimated SNR value and these instantaneous values are used to compute the average SNR estimate. Table 3 shows the parameters used for second part of proposed technique. The mean-squared-error between the actual noise and estimated noise values in each sub-band are calculated and averaged. The MSE is computed as described in (37).

$$MSE = \frac{1}{2000}\sum_{i=1}^{2000}\left(\hat{SNR} - SNR\right)^2 \tag{37}$$

For each SNR value in each sub-band 2000 iterations are performed and MSE and mean estimated SNR values of each sub-band are computed for actual SNR value in that sub-band. The performance of proposed estimator is compared with Reddy estimator which uses 50 OFDM symbols. The results are shown in the Figs. 12 and 13 for SNR-MSE vs. SNRdB and Estimated global SNR vs. Actual SNRdB.

Fig.12 shows the mean-square-error performance of the estimator methods in colored noise. For global SNR values, mean-squared-error between the actual noise and estimated noise values in each sub-group are calculated and averaged. The proposed algorithm performs much better than Reddy and conventional noise power estimation in terms of finding the local noise power. It is clear from the figure 12 that the proposed estimator shows constant difference in performance than Reddy estimator over the entire considered range of SNR values. Fig. 13 shows the proposed estimator gives almost the same global estimates as actual SNR in case of colored noise.

Consistency Analysis: Fig. 14 shows the NMSE of SNR for different values of average SNR (SNR =5 dB, 10dB, 15dB). The estimates are seen to have good performance with one OFDM symbol and also very consistent when averaged all the way over 40 OFDM symbols unlike all previous techniques which take many OFDM symbols to give better estimates.

Table 2: Parameters for part one of proposed method

Nfft size	256
Sampling Frequency (*Fs*)	20*MHz*.
SubCarrier Spacing (Δf =Fs/Nfft)	1×10^{5}
Useful Symbol Time (Tb=1/ Δf)	1×10^{-5}
CP Time (*Tg*=*G***Tb*) where G=1/4	2.5×10^{-6}
OFDM Symbol Time (*Ts* = *Tb* + *Tg*)	1.25×10^{-5}
Ts=5/4**Ts* (Because ¼ CP makes the sampling faster by 5/4 times)	1.5625×10^{-5}

Table 3: Parameters of part second of proposed method

$T_{sub} = {}^{T_s}\!/_{16}$	9.8×10^{-7}
Wavelet Packet Object Structure Wavelet Decomposition Command : *wpdec* 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 : *Daubechies (db3)* , Entropy Name : *Shannon*	

Complexity and Accuracy: The results shows that the proposed estimator fulfill the criteria of good SNR estimator.The proposed estimator has relatively low computational complexity and easy to implement because it makes use of only one OFDM preamble signal to find the SNR estimates unlike all previous SNR estimators. It provides accurate SNR estimates as the estimation error is less than 0.2dB overall. It is computationally fast because there is no need of averaging over many OFDM symbols to get accurate SNR estimates as compared to previous SNR estimators.

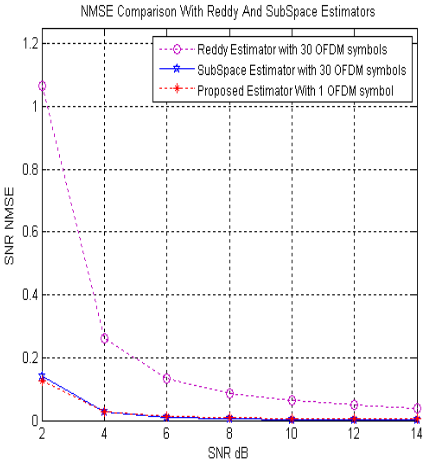


Fig.8 Comparison of part-one's proposed method

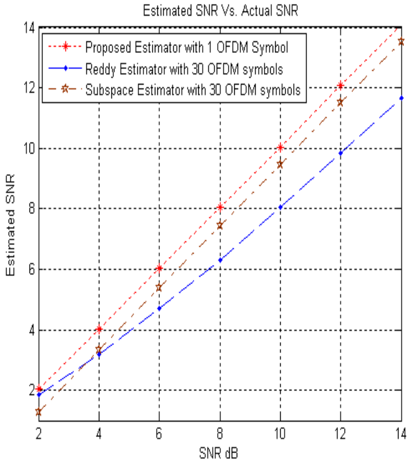


Fig.9 Comparison of part one's proposed method

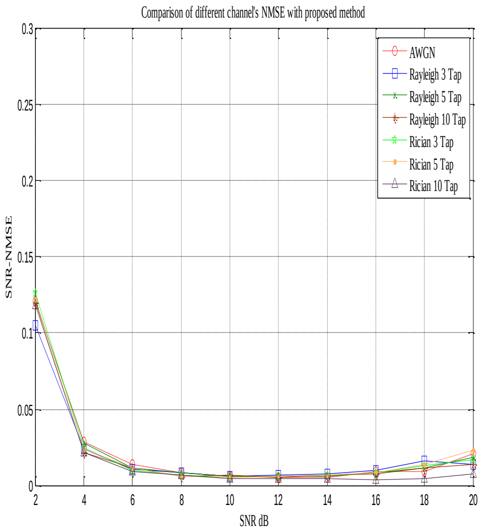


Fig. 10 Comparison of different channel's NMSE with part one's proposed method.

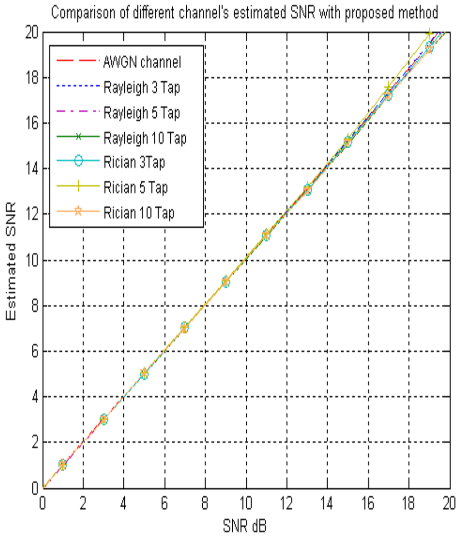


Fig.11 Comparison of different channel's est SNR with fist part of proposed method

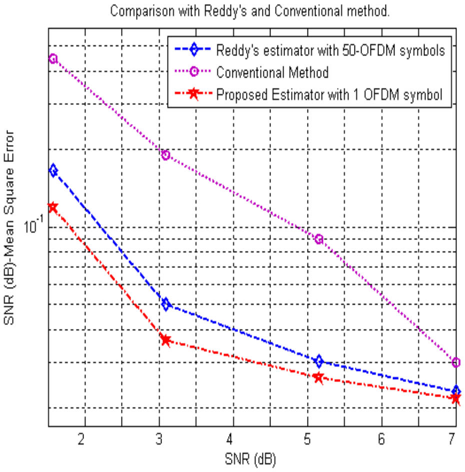


Fig.12 Mean-square-error performance of the proposed technique in 2nd part with other algorithms in colored noise.

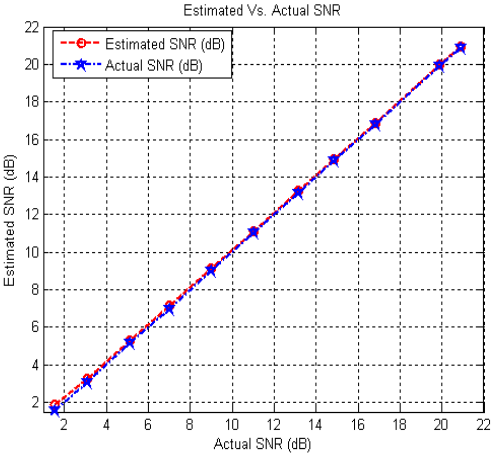


Fig.13 Actual SNR vs. global SNR estimates of colored noise with proposed technique in 2nd -part

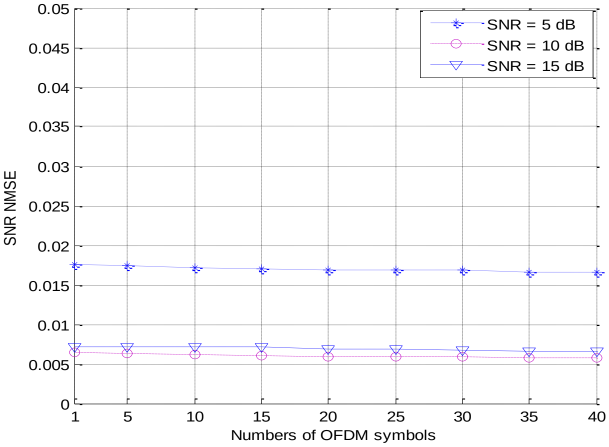


Fig.14 SNR NMSE for different values of average SNR (consistency check for estimates by averaging)

5. Conclusion

There is no work done at front-end of the receiver for multicarrier systems like OFDM. In contrast to other SNR estimators that derive SNR estimates at the back-end of the receiver, a novel SNR estimator is presented which can operate at the front-end of the receiver. Proposed estimator makes use of one OFDM symbol to estimates the SNR unlike all previous estimators which makes use of many OFDM symbols to get SNR estimates. 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. The performance of proposed technique has been evaluated via computer simulations and implemented in OFDM systems. The results show that the current estimator performs better than other conventional methods. Complexity to find SNR estimates is much lower because the current estimator makes use of only one OFDM preamble signal. To check the consistency of estimator, averaging SNR NMSE results for different values of average SNR are taken and results shows the developed SNR estimation technique is provides very reliable estimates of SNR.

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