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Performance Comparisons of Equalizers for MC-CDMA UWB Communication Systems

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***Abstract*—Nowadays, wireless communication technology usually takes the advantage of orthogonal frequency-division multiplexing (OFDM). It becomes Multi-carrier code division multiple access (MC-CDMA) after utilizing spreading code in frequency domain. The chip-level equalizers of MC-CDMA system always equalize data before despreading that enlarge difficulty of hardware compared with that of equalizers in OFDM system. In this paper, we also design an bit-level equalizer which equalizes data after despreading. As a result, the hardware implementation difficulty of the bit-level equalizer may be reduced. Here, we study the performances of the proposed equalizers in the ultra-wideband (UWB) MC-CDMA system for wireless personal area network (WPAN). Simulation results show the performance comparisons of the proposed MC-CDMA equalizers.**

I. Introduction

ULTRA-WIDEBAND (UWB) TECHNIQUE IS GOOD FOR INDOOR SHORT-RANGE HIGH-SPEED WIRELESS COMMUNICATION. UWB IS DEFINED BY THE U.S. FEDERAL COMMUNICATIONS COMMISSION (FCC) AS ANY WIRELESS TRANSMISSION SCHEME THAT OCCUPIES A FRACTIONAL BANDWIDTH 20% OF CENTER FREQUENCY IN THE FREQUENCY RANGE FROM 3.1 TO 10.6 GHZ. THE POWER SPECTRAL DENSITY EMISSION LIMIT FOR UWB TRANSMITTERS IS −41.3 DBM/MHZ. BLUETOOTH IS A POPULAR TECHNIQUE FOR WIRELESS PERSONAL AREA NETWORK (WPAN) BUT ITS TRANSMISSION SPEED IS LESS THAN UWB [1]. THEREFORE, UWB IS QUALIFIED FOR ENORMOUS FILE TRANSMISION IN WPAN E.G., VIDEO FILE TRANSFER BETWEEN CELL-PHONE AND OTHER HANDHELD DEVICES.

UWB is considerably suffered by frequency selective fading that is described in Saleh-Valenzuela model (S-V model) due to the clustering phenomena observed at the measured UWB indoor channel data [2]. In realistic UWB channels, IEEE 802.15.SG3a task group proposed a modified S-V model as the UWB multipath channel model [3]. It is log-normal distribution rather than Rayleigh distribution for the multipath gain magnitude.

Many different pulse generation techniques may be used to satisfy the requirements of an UWB signal e.g., time-modulated ultra wideband (TM-UWB), pulse amplitude modulation (PAM), and pulse position modulation (PPM). The UWB channel is very similar to wideband channel as be experienced in spread spectrum or code division multiple access (CDMA) systems. The corresponding train of impulses can also be generated using a conventional direct sequence spread spectrum (DSSS) based measurement system [4]. In this paper, we discuss the equalization techniques for MC-CDMA UWB systems. MC-CDMA system is a mixture of DS-CDMA and OFDM, which is a promising broadband communication technique because of combing the advantages of OFDM and CDMA. The conventional equalizers in MC-CDMA system usually equalizes data before despreading [5, 6], which is named as chip-level equalizers. The difficulty of equalizers implementation is higher than that of equalizers in OFDM system because of utilizing spreading codes. Therefore, we proposed a new equalizer for MC-CDMA system in order to maintain similar hardware difficulty as in OFDM system, which is named as bit-level equalizers in this paper.

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We will describe MC-CDMA UWB system in Section II. In Section III, UWB channel is presented. Performance comparisons of bit-level and chip-level equalizers are provided in Section IV. Section V concludes this paper.

Throughout this paper, lower and uppercase letters denote time-domain and frequency-domain entities, respectively. Boldface letters denote column vectors and matrices. $\mathbf{I}$ denotes an identity matrix. The superscripts $(\cdot)^T$, $(\cdot)^H$, and $E\{\cdot\}$ symbolize the transpose, Hermitian operations, and mean, respectively.

II. System Model

For MC-CDMA systems shown as in Fig. 1 and Fig. 2, the input data stream is mapped to the symbols taken from binary phase shift keying (BPSK) modulation constellation. After serial to parallel conversion, spreader spreads the data stream over different subcarriers using a given spreading code in the frequency domain. Then, the inverse fast Fourier transform (IFFT) is taken to obtain the time domain samples. To avoid inter symbol interference (ISI), the cyclic prefix (CP) is inserted between each symbols, which length is no less than the delaying time of the multipath effect.

A. Common Structure

For MC-CDMA system, we denote the vector of the data transmitted by all users during the i^{th} symbol as

$$\mathbf{D}_i = \left[\mathbf{d}_i^1, \mathbf{d}_i^2 \ldots, \mathbf{d}_i^m, \ldots, \mathbf{d}_i^U\right]^T, \qquad (1)$$

where $\mathbf{d}_i^m$ represents the data symbol assigned to user m

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(m=1,…,U). When the m^{th} user is inactive $\mathbf{d}_i^m = 0$ .We denote $\mathbf{C}$ as the spreading code matrix, which column m represents the spreading code of user m, and L means the length of spreading code.

$$\mathbf{C} = \begin{bmatrix} c_1^1 & c_1^2 & \cdots & c_1^m & \cdots & c_1^U \\ c_2^1 & c_2^2 & \cdots & c_2^m & \cdots & c_2^U \\ \vdots & \vdots & & \vdots & & \vdots \\ c_L^1 & c_L^2 & \cdots & c_L^m & \cdots & c_L^U \end{bmatrix}. \quad (2)$$

The domain before IFFT can be regarded as frequency domain. On the other hand, the domain after IFFT can be regarded as time domain. Thus, the convolution process of channel impulse response can be equalized as the multiplication process of channel frequency response before IFFT. We denote a diagonal matrix $\mathbf{H} = \mathbf{diag}\{[h_1,...,h_L]\}$ to describe the channel frequency response. Let $\mathbf{F}$ denote the discrete Fourier transform (DFT) matrix with $\mathbf{F}_{p,k}$ = $1/\sqrt{L}\exp(-j2\pi(p-1)(k-1)/L)$, where $j = \sqrt{-1}$. Since $\mathbf{F}$ is a unitary matrix, we have $\mathbf{F}^{-1} = \mathbf{F}^H$.

The signal after IFFT can be expressed as

$$\mathbf{t} = \mathbf{F}^H \mathbf{HCD} \quad (3)$$

To simplify our study, CP will be ignore, so the received signal can be written as

$$\mathbf{r} = \mathbf{t} + \mathbf{n}, \quad (4)$$

where $\mathbf{n} = \left[n_1, n_2, \ldots, n_y, \ldots, n_L\right]^T$ is the additive white Gaussian noise (AWGN) term, with a variance equal to $\sigma_n{}^2 = E\left\{n_y{}^2\right\}$, y= 1,…, L, where n_y representing the noise term at the y^{th} subcarrier.

The fast Fourier transform (FFT) is used to transform the received signal $\mathbf{r}$ to the frequency domain. From (3) and (4), we have

$$\begin{aligned} \mathbf{Fr} &= \mathbf{Ft} + \mathbf{Fn} \\ \mathbf{R} &= \mathbf{HCD} + \mathbf{N}, \end{aligned} \quad (5)$$

where $\mathbf{R} = \mathbf{Fr}$ and $\mathbf{N} = \mathbf{Fn}$.

Based on Fig. 1 and Fig. 2, the structures are the same before FFT block, so we discuss the details of different equalizers below, respectively.

B. Bit- Level Zero Forcing Equalizer

To deduce bit-level zero forcing (ZF) equalizer we have to ignore $\mathbf{N}$ in (5). The received signal of the m^{th} user after despreader is expressed as $\mathbf{S}$. From (5), we have

$$\mathrm{S} = \mathbf{C}^{mH} \mathbf{HCD}, \quad (6)$$

where $\mathbf{C}^m$ represents the m^{th} column of $\mathbf{C}$.

The main purpose of the ZF equalizer is to estimate the transmitted signal $\mathbf{D}$. We denote $\mathbf{E}$ as the matrix of the proposed equalizer as follows,

$$\mathbf{E} = \mathbf{D}\mathrm{S}^{-1} = (\mathbf{C}^{mH}\mathbf{HC})^{-1} \quad (7)$$

According to assumption that only the m^{th} user is using the system, it can be expressed as

$$E_b = \frac{1}{\sum_{i=1}^{L} \frac{h_i}{L}}, \quad (8)$$

where b represents the b^{th} bit in the symbol.

C. Chip-Level Equalizers

Here, we will discuss ZF and minimum mean square error (MMSE) equalizers. The ZF criterion is used to eliminate the channel effect. The transfer function of the ZF equalizer can be expressed as

$$\mathrm{W}_{\mathrm{ZF}_n} = \frac{1}{h_n}, \quad (9)$$

where $\mathrm{W}_{\mathrm{ZF}_n}$ represents the weight of the equalizer for the n^{th} subcarrier.

To discuss MMSE equalizer, we assume

$$\mathbf{G}^m = \mathbf{C}^{mH} \mathbf{W}_{\mathrm{MMSE}}, \quad (10)$$

where $\mathbf{G}^m = \left[G_1^m, G_2^m, ..., G_L^m\right]$ is the optimal weight vector of the m^{th} user, $\mathbf{W}_{\mathrm{MMSE}}$ represents the equalization coefficients which is a L*L matrix from channel estimation. From (5), the estimated symbol of the m^{th} user can be expressed as

$$\hat{\mathbf{d}}^m = \mathbf{G}^m \mathbf{R} = \mathbf{C}^{mH} \mathbf{W}_{\mathrm{MMSE}} \mathbf{R} \quad (11)$$

The main purpose of the MMSE equalizer is to minimize the mean square error between the estimated symbol $\hat{\mathbf{d}}^m$ and the transmitted symbol, which can be expressed as

$$\min_{\mathbf{G}^m} E\left\{\| \mathbf{d}^m - \mathbf{G}^{mH} \mathbf{R} \|\right\} \quad (12)$$

The optimal weighting vector can be obtained by the Wiener filtering [7]. We have

$$\mathbf{G}^{\mathrm{m}} = \mathbf{\Gamma}_{d^m,R} \mathbf{\Gamma}_{R,R}^{-1}, \quad (13)$$

where $\mathbf{\Gamma}_{R,R}$ is the autocorrelation matrix of the received vector $\mathbf{R}$, $\mathbf{\Gamma}_{d^m,R}$ represents the crosscorrelation vector between the received vector $\mathbf{R}$ and the desired symbol of the m^{th} user $\mathbf{d}^m$. From (5), we have

$$\begin{aligned} \mathbf{\Gamma}_{R,R} &= E\left\{\mathbf{R}\,\mathbf{R}^H\right\} \\ &= \mathbf{HC}E\left\{\mathbf{D}\,\mathbf{D}^H\right\}\mathbf{C}^H\mathbf{H}^H + E\left\{\mathbf{N}\,\mathbf{N}^H\right\} \end{aligned} \quad (14a)$$

$$\mathbf{\Gamma}_{d^m,R} = E\left\{\mathbf{d}^m\,\mathbf{R}^H\right\} = E\left\{\mathbf{d}^m\,\mathbf{D}^H\right\}\mathbf{C}^H\mathbf{H}^H \quad (14b)$$

Therefore, from (13), (14a) and (14b), we can expressed the optimal weighting vector as

$$\begin{aligned} \mathbf{G}^m &= E\left\{\mathbf{d}^m\,\mathbf{D}^H\right\}\mathbf{C}^H\mathbf{H}^H \\ &\bullet\left(\mathbf{HC}E\left\{\mathbf{D}\,\mathbf{D}^H\right\}\mathbf{C}^H\mathbf{H}^H + E\left\{\mathbf{N}\,\mathbf{N}^H\right\}\right)^{-1} \end{aligned} \quad (15)$$

We assume that signals are mutual independent and have the same power ($E\left\{(\mathbf{d}^m)^2\right\} = \mathrm{E_p}$). Besides, the subcarrier noises are independent and have the same variance. Thus, we have $E\left\{\mathbf{N}\,\mathbf{N}^H\right\} = \sigma_N^2\mathbf{I}$. As a result, (15) becomes

$$\mathbf{G}^m = \mathrm{E_p}\mathbf{C}^H\mathbf{H}^H\left(\mathbf{HC}\mathrm{E_p}\mathbf{A}\mathbf{C}^H\mathbf{H}^H + \sigma_N^2\mathbf{I}\right)^{-1}, \quad (16)$$

where $\mathbf{A} = \{a_{kk}\}$ is a diagonal matrix with the term $a_{mm} = 1$ if user m is active and $a_{mm} = 0$ if user m is inactive [8].

Combine (10) and (11), we have

$$\begin{aligned}\mathbf{G}^m &= \mathbf{C}^{mH}\mathbf{W}_{\mathrm{MMSE}} \\ &= \mathbf{C}^H\mathbf{H}^H\left(\mathbf{HCAC}^H\boldsymbol{H}^H + \frac{\sigma_N^2}{\mathrm{E_p}}\mathbf{I}\right)^{-1} \qquad (17)\end{aligned}$$

Due to the orthogonal attribute of spreading code, from (17) we have

$$\mathbf{W}_{\mathrm{MMSE}} = \mathbf{H}^H\left(\mathbf{HCAC}^H\mathbf{H}^H + \frac{\sigma_N^2}{\mathrm{E_p}}\mathbf{I}\right)^{-1} \qquad (18)$$

In the full load case, the quantity $\mathbf{CAC}^H$ is equal to the identity matrix. So $\mathbf{W}_{\mathrm{MMSE}}$ becomes a diagonal matrix which can expressed as

$$\mathbf{W}_{\mathrm{MMSE}_n} = \frac{\mathrm{h}_n^*}{|\mathrm{h}_n|^2 + \frac{1}{\mathrm{r}_c}}, \qquad (19)$$

where $\mathbf{W}_{\mathrm{MMSE}_n}$ represents the n^{th} equalization coefficient for the n^{th} subcarrier and r_c represents the signal to noise ratio (SNR).

Besides full load condition, $\mathbf{W}_{\mathrm{MMSE}}$ is not a diagonal matrix. We assume each subcarrier is mutual independent. The secondary optimal equalization coefficient for the n^{th} subcarrier can be written as

$$\mathbf{W}_{\mathrm{MMSE}_n} = \frac{\mathrm{h}_n^*}{|\mathrm{h}_n|^2 + \frac{L}{\mathrm{N}_m\mathrm{r}_x}}, \qquad (20)$$

where r_x is the SNR of the received data symbol $\mathbf{d}^{\mathrm{m}}$, N_m is the number of user, L is the length of the spreading code.

III. UWB CHANNEL MODEL

The most widely studied UWB channel models is expressed as

$$h(x) = X\sum_{l=0}^{M}\sum_{k=0}^{K}\alpha_{k,l}\delta(t - T_l - \tau_{k,l}),$$

where M is the number of multipath clusters, K is the number of multipath components within a cluster, $\alpha_{k,l}$ is the multipath gain coefficients corresponding to the k-th multipath component of the l-th cluster, T_l is the group delay of the l-th cluster, $\tau_{k,l}$ is the delay of the k-th path within the l-th cluster relative to the first path arrival time T_l, and X represents the log-normal shadowing. The definition assumes that $\tau_{0,l}$=0. The distribution of cluster arrival time and the ray arrival time are given by

$$p(T_l \mid T_{l-1}) = \Lambda\exp[-\Lambda(T_l \mid T_{l-1})], l > 0$$

$$p(\tau_{k,l} \mid \tau_{(k-1),l}) = \lambda\exp[-\lambda(\tau_{k,l} - \tau_{(k-1),l})], k > 0$$

where Λ and λ are the cluster arrival rate and arrival rate of path within each cluster, respectively. Two different channel characteristics of the S-V model are shown in table 1 from measurement data by Intel [3]. A brief description of the selected channels is as follows.

- CM1: line-of-sight (LOS) model for 0–4-m.
- CM4: Non-line-of-sight (NLOS) for 4–10-m model

IV. SIMULATION RESULTS

We assume the following parameters for simulations.

- One User
- Channel model: S-V channel
- Channel sampling time: 0.167 ns
- CM1 channel delay time: 600 ns
- CM4 channel delay time: 1600 ns
- Bandwidth: 3GHz
- Modulation: BPSK
- OFDM symbol size: 512 bits
- Spreading code: Hadamard code

The number of weights for the chip-level equalizers is equal to that of the spreading code length multiplying OFDM symbol size. The number of weights for the bit-level equalizer is equal to the OFDM symbol size. Figs. 3–6 show the simulation result of bit error rate (BER) versus SNR of the MC-CDMA UWB system under two different S-V model channels (CM1 and CM4) and two different spreading code lengths (4 and 8).

In CM1 (line-of-sight model) channel, it is noticed from Fig. 3 and Fig. 5 that the performances of the chip-level MMSE equalizer and the bit-level ZF equalizer have similar BER and they both perform better than chip-level ZF equalizer. But both chip-level equalizers outperform bit-level ZF equalizer while in high SNR condition (higher than 22dB).

As illustrated in Fig. 4 and Fig. 6, in CM4 (non-line-of-sight model) channel, it can be seen that the chip-level ZF equalizer perform better than the bit-level ZF equalizer in the low SNR region, and the chip-level MMSE equalizer has superior performance than that of the both ZF equalizers. In CM4, we can realize that both the chip-level equalizers outperform the bit-level equalizer more significantly while SNR increased.

By comparing the performances by using different spreading code length in the same channel, we can conclude that all equalizers perform better when longer spreading code is used, also longer spreading code length enlarge the performance gap between the MMSE and the ZF equalizers in low SNR condition when in same channel environment.

V. CONCLUSION

In this paper, our simulation results show that the chip-level ZF equalizer and the bit-level ZF equalizer have similar performance in line-of-sight (LOS) model, but the number of weights in the bit-level ZF equalizer is much less than that of the chip-level ZF equalizer. Therefore, we advise to use the bit-level ZF equalizer while transmitting signals in unhostile channels in order to make hardware implementation more simple. On the other hand, if we don't consider about the difficulty of hardware implementation then the chip-level MMSE equalizer is recommended because of its outstanding performance in any conditions.

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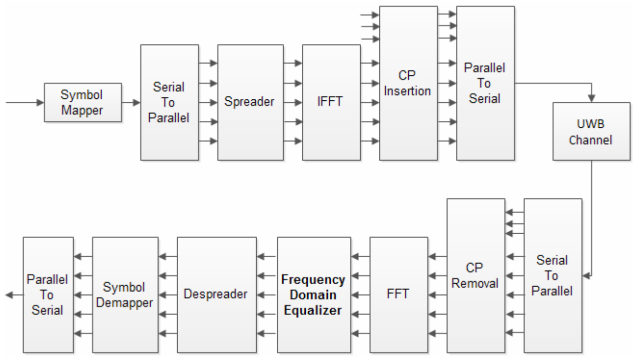

Fig. 1. Bit-level equalizer in MC-CDMA UWB Systems

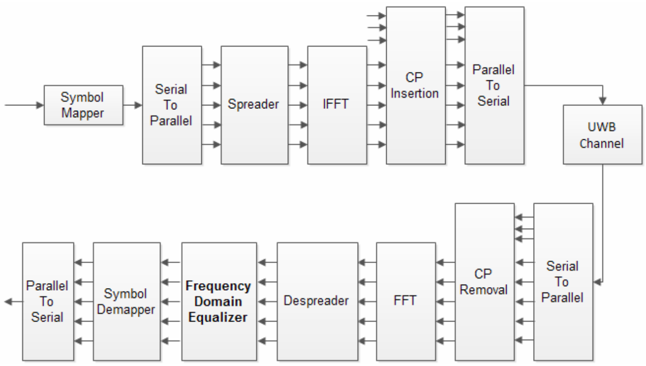

Fig. 2.Chip-level equalizer in MC-CDMA UWB Systems

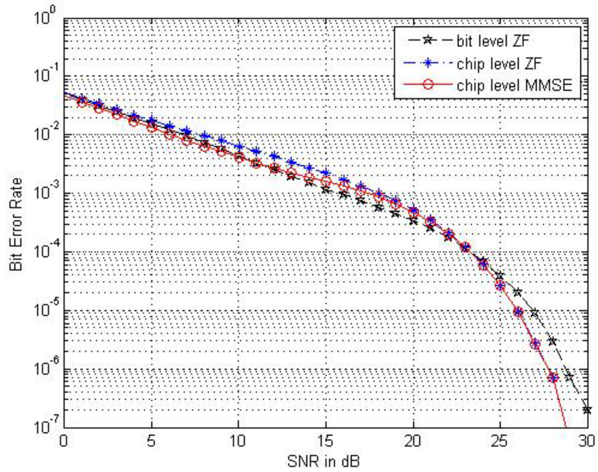

Fig. 3. BER performances for spreading code length 4 and in CM1.

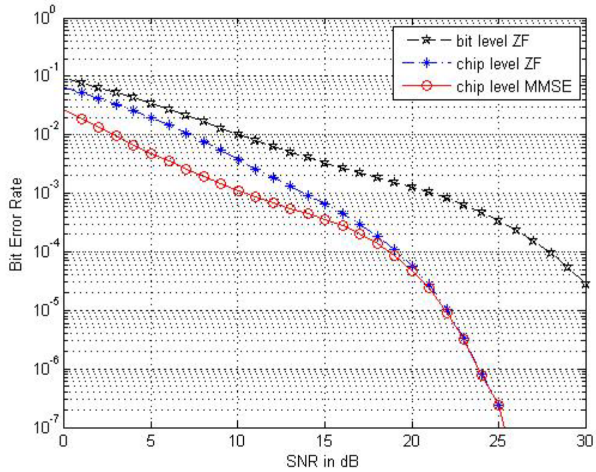

Fig. 4. BER performances for spreading code length 4 and in CM4.

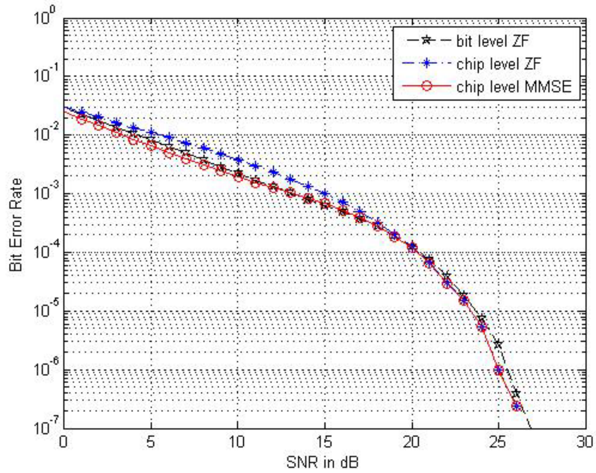

Fig. 5. BER performances for spreading code length 8 and in CM1.

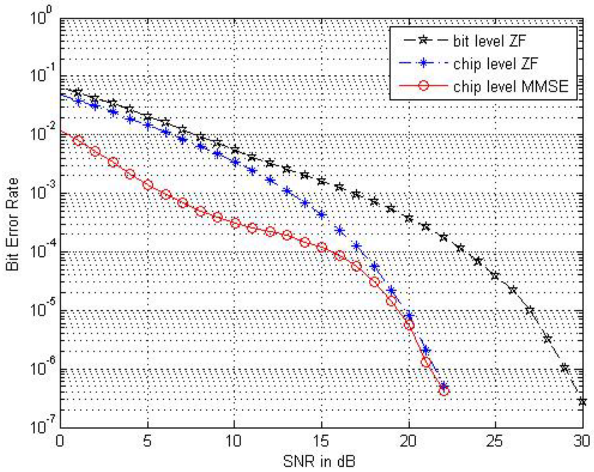

Fig. 6. BER performances for spreading code length 8 and in CM4.

TABLE 1 . S-V CHANNEL CHARACTERISTICS PROPOSED BY INTEL

Target ChannelCharacteristics	CM1	CM4
Mean excess delay (nsec) (τ m)	5.05	
RMS delay(nsec) (τ m)	5.28	25
NP10dB		
NP (85%)	24	
Model parameters		
Λ (1/nsec)	0.0233	0.0667
λ (1/nsec)	2.5	2.1
Γ	7.1	24.00
γ	4.3	12
σ 1 (dB)	3.3941	3.3941
σ 2 (dB)	3.3941	3.3941
σ x (dB)	3	3
Model Characteristics	CM1	CM4
Mean excess delay (nsec)	5.0	30.1
RMS delay (nsec)	5	25
NP10dB	12.5	41.2
NP (85%)	20.8	123.3
Channel energy mean (dB)	-0.4	0.3
Channel energy std (dB)	2.9	2.7

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