SELECTION DIVERSITY AND LINEAR EQUALIZATION OVER FREQUENCY SELECTIVE CHANNELS FOR SINGLE CARRIER FILTER BANK-BASED TRANSMISSIONS

fredharris, Xiaofei Chen*+, ElettraVenosa*, BhaskarRao+*

San Diego State University*, University of California, San Diego+

ABSTRACT

This paper investigates the filter bank (FB) based selection diversity combining as well as linear equalization for single carrier (SC)transmissions over frequency selective channels. In contrast to the multicarrier carrier (MC) transmissions, e.g., OFDM, the FB based approach avoids the use of cyclic prefix (CP) or guard band and offers a number of superior properties such as synchronization, and peak to average power ratio (PAPR), etc. However, due to the lack of practical diversity techniques and cost effective equalizers, the broadband SC signal can hardly be deployed under highly dispersive channel environment. We propose a practical FB based selection diversity based on non-maximally decimated filter banks with perfect reconstruction support (PR-NMDFB) with its companion linear equalizer in the FB transformed domain. We shall give detailed minimum mean square error (MMSE) and bit error rate (BER) analysis and compare it to the optimal maximum ratio combining (MRC) solution.

Index Terms— filter bank, selection diversity, maximum ratio combining, polyphase, single carrier

1. INTRODUCTION

Many modern communication systems require broadband communication over highly dispersive frequency selective channels. The MC transmission, e.g., OFDM [1], has captured most of the attention due to its simple equalization scheme as well as its low processing speed nature, i.e., serial to parallel conversion. These important characteristics allow one to build very broadband systems in a cost effective manner. However, as pointed out in [2], the OFDM based MC systems suffer from a number of drawbacks, such as CP overhead, high PAPR, sensitivity to carrier frequency offset; and these facts limit the system performance from throughput, power consumption, and BER perspectives. On the other hand, the legacy SC signal, i.e., square root raised cosine (SRRC) shaped QAM waveform, has these problems well controlled. Yet, legacy waveforms require very long equalizers, i.e., the ATSC-8VSB [3] has 6 MHz wide signal bandwidth; and requires total equalizer length exceeding 400 taps, the longest equalizer ever deployed. Moreover, the diversity techniques are not well supported for SC transmissions over frequency selective channels. Back in 1992, the author of [4] developed the optimum diversity receiver structure over frequency selective channel, as shown in Fig 1, which is later recognized as an MRC by [5]. The received signals from the *N*-branch independent antennas are first matched filtered (matched to the shaping pulse and the channel) and then summed to form the combined signal. The combined signal is then sent to a tapped delay line equalizer to produc desired symbol outputs subject to either zero forcing (ZF) or MMSE criteria.

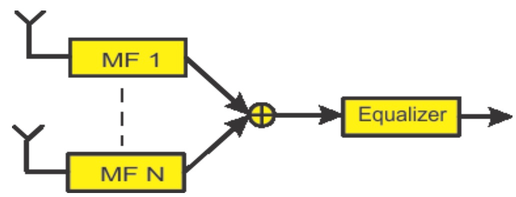

Fig. 1 Optimum Combiner / Linear Equalizer Structure

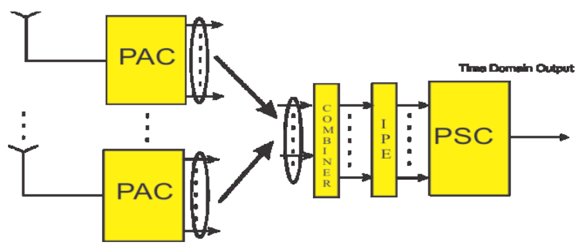

Fig.2 Filter Bank Based Diversity Combining Model

The direct implementation of Fig .1 not only requires significant hardware resources on filtering, but also needs precise channel knowledge for each branch. Comparing to MC techniques where filtering is FFT based and channel gains can be extracted from preamble, the SC transmission at first glance seems very incapable in broadband wireless communication. The author of [5] took the first step in introducing FB based combining technique to non-CP SC system; whose generalized block diagram is presented in Fig 2. The idea is to use analysis filter bank (AFB), whose polyphase form is named as polyphase analysis channelizer (PAC), to interface with each receiver's antenna; and then perform combining in the PAC transformed domain. After combining, the equalizer intermediate processing element (IPE) is applied. Finally, the combined and equalized signal is transformed back to time domain via synthesis filter bank (SFB), whose polyphase form is called polyphase synthesis channelizer (PSC). Although [5] introduced the structure in Fig. 2, and showed the combining mechanism matched to optimum combiner in Fig 1,the theoretical performance evaluation is not provided and the equalizer's performance is not discussed. In this paper, we attempt to answer these questions; and we shall propose FB based selection diversity, which does not require any channel knowledge. We will compare the performance between the proposed and the optimal MRC approaches.

The organization of this paper is: section 2 reviews the FB; section 3 presents signal model; section 4 derives the combining technique; section 5 gives simulation results; section 6 draws conclusion.

2. BACKGROUND ON PR-NMDFB

We adopt the modified discrete Fourier transform (MDFB) developed in [6]. This type of oversampled FB does not require adjacent channels to participate in the PR process. Therefore one is allowed to alter gain and phase to AFB outputs while preserving the PR property. The polyphase implementations are reported in [7], and we use the band pass filter (BPF) model, shown in Fig.3 as our analysis tool in this paper.

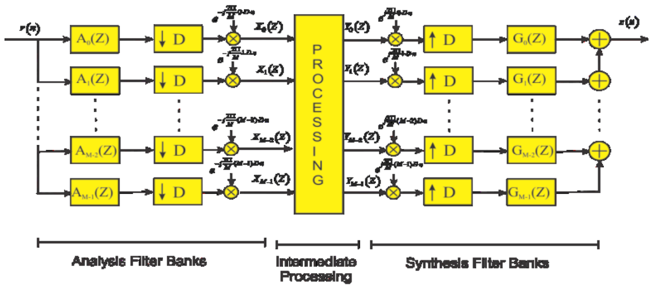

Fig.3 Generalized M-path PR-NMDFB Model

The AFB contain MBPFs, whose Z-transforms are $A_m(Z)$, $form = 0,1,\dots,M-1$. The M BPFs have equal bandwidth, and are centered on digital frequencies $\theta_m = \frac{2\pi}{M}m$, for $m = 0,1,\dots,M-1$. Let $a(n)$ be the impulse response of theLPPF. The m^{th} BPF is $a_m(n) = a(n)e^{j\frac{2\pi}{M}mn}$, whose Z-transform is $A_m(Z) = A\left(e^{-j\frac{2\pi}{M}m}Z\right) = A(W_M^m Z)$, and $W_M \overset{\text{def}}{=} e^{-j\frac{2\pi}{M}}$. A down sampling by a factor D (*D<M*), which is an integer that divides M, follows each BPF. As a standard practice, we require the AFB transformed signal to be centered on zero frequency and this is done via a set of complex rotators whose values are $e^{-j\frac{2\pi}{M}mnD}$. The AFB transformed signal then processed by the IPE at deeply decimated sampling rate. The SFBs perform the exact inverse process of AFBs: heterodyne, up sample and filtering. It can be shown that the Z-transform of the output signal $z(n)$ is:

$$
\begin{aligned}
Z(z) &= \frac{1}{D}\boldsymbol{G}_{1\times M}^T(\boldsymbol{Z})\mathbb{K}_{M\times M}\mathbb{A}_{M\times D}(\boldsymbol{Z})\boldsymbol{R}_{D\times 1}(\boldsymbol{Z}) \\
&= \frac{1}{D}\boldsymbol{T}^{\mathbb{K}}{}_{1\times D}(\boldsymbol{Z})\boldsymbol{R}_{D\times 1}(\boldsymbol{Z}) \\
&= \frac{1}{D}T_s^{\mathbb{K}}(Z)R(Z) + \frac{1}{D}\boldsymbol{T}_A^{\mathbb{K}}(Z)\bar{\boldsymbol{R}}(Z)
\end{aligned}
$$

where$G(Z) = [G(ZW_M^0)\dots G(ZW_M^{M-1})]^T$ is the BPFs for SFB; $\mathbb{K}_{M\times M} = diag\{k_0 \dots k_{M-1}\}$ is the IPE complex scalar gain applied in between AFB and SFB; column vector $R(Z) = [R(ZW_D^0)\dots R(ZW_D^{D-1})]^T = [R(ZW_D^0)\bar{R}(Z)]^T$ is the modulated versions of the input signal; matrix $\mathbb{A}_{\mathrm{M\times D}}(\mathrm{Z})$ is defined as:

$$
\begin{aligned}
\mathbb{A}(\boldsymbol{Z}) &= \begin{bmatrix} A(ZW_M^0W_D^0) & \dots & A(ZW_M^0W_D^{D-1}) \\ \vdots & \ddots & \vdots \\ A(ZW_M^{M-1}W_D^0) & \dots & A(ZW_M^{M-1}W_D^{D-1}) \end{bmatrix}_{M\times D} \\
&= \left[\boldsymbol{A}_{M\times 1} \mid \bar{\mathbb{A}}_{M\times(D-1)}\right]_{M\times D}
\end{aligned}
$$

And $\boldsymbol{T}^{\mathbb{K}}{}_{1\times D}(\boldsymbol{Z}) \triangleq \boldsymbol{G}_{1\times M}^T(\boldsymbol{Z})\mathbb{K}_{M\times M}\mathbb{A}_{M\times D}(\boldsymbol{Z}) = \left[T_s^{\mathbb{K}}(Z)\boldsymbol{T}_A^{\mathbb{K}}(Z)\right]$ is thetotal transfer function (TF) for the M-path, decimate by D, AFB and SFB;$T_s^{\mathbb{K}}(Z) \triangleq \boldsymbol{G}_{1\times M}^T(\boldsymbol{Z})\mathbb{K}_{M\times M}\boldsymbol{A}_{M\times 1}(\boldsymbol{Z})$is the desired signal TF, whereas$\boldsymbol{T}_A^{\mathbb{K}}(Z) \triangleq \boldsymbol{G}_{1\times M}^T(\boldsymbol{Z})\mathbb{K}_{M\times M}\bar{\mathbb{A}}_{M\times(D-1)}(\boldsymbol{Z})$, is the undesired aliasing TF. We require the NMDFB to have PR property [6,7] This condition translates to: 1) Zero aliasing TF $A(ZW_D^d)G(Z) = 0, \forall\, d = 1,\dots,D-1$; 2) Distortionless signal TF, $A(Z)\,G(Z) = H^{NYQ}(Z)$, where $H^{NYQ}(Z)$ is any Nyquist pulse. Additional details on this topic can be found in [12].

3. SIGNAL MODEL

Let us examine a baseband equivalent system for a QAM signal with complex notation for the in-phase (I) and quadrature (Q) branches. Denoting the expectation as [▪] and the k^{th} complex QAM data symbol as S_k, with symbol period T seconds. We assume the data symbols are stationary and uncorrelated, $[S_k S_{k'}] = \sigma_s^2\delta_{kk'}$, where $\delta_{kk'}$ is the Kronecker delta function. At the transmitter (Tx), the symbol stream is first 1-to-2 zero packed and then shaped by a square root raised cosine (SRRC) filter $h^{tx}(n)$; and assume perfect digital to analog conversion, the emitted signal has power spectral density $\frac{1}{2}\sigma_s^2|H^{tx}(\Omega)|^2$, where Ω is the analog frequency; and $H_{tx}(\Omega)$ is the fourier transform of the shaping pulse. Denote the complex base-band equivalent channels for N-branch receiver as $h_l^c(t)$, for $l = 1,2,\dots,L$. The additive white noise $n_l(n) = n_l^I(t) + jn_l^Q(t)$ of two-sided power spectral density $N_0/2$ W/Hz per complex components is introduced at the output of each independent channel. The received l^{th} branch continuous signal is written as:

$$r_l(t) = \sum_{k=-\infty}^{\infty} S_k h_l^{tc}(t - kT - t_l) + n_l(t) \qquad (1)$$

where, $h_l^{tc}(t) \triangleq h^{tx}(t) * h_l^c(t)$; and t_l is the channel delay or the sampler phase. In this paper, we assume the channel $h_l^c(t)$ is free of linear phase component, and set $t_l = 0$. Digitizing the received signal at sampling speed $T_s = T/2$, i.e., 2 samples-per-symbol, we find digitized signal as:

$$r_l(n) = \sum_{k=-\infty}^{\infty} S_k h_l^{tc}(n - 2k) + n_l(n) \qquad (2)$$

Here we assume the noise variance on all L-branches are the same and equal to σ_n^2. The decimated by D and frequency translated output signal observed at the m^{th} AFB output on the l^{th} diversity branch is:

$$x_{m,l}(n) = (\downarrow D)\left[\sum_k S_k h_{m,l}^a(n-2k)e^{-j\theta_m n} + v_{m,l}(n)e^{-j\theta_m n}\right] \qquad (3)$$

where, $h_{m,l}^a(n) \triangleq h_l^{tc}(n) * a_m(n)$, and $v_{m,l}(n) \triangleq n_l(n) * a_m(n)$. The Discrete Time Fourier Transform (DTFT) of $h_{m,l}^a(n)$ is denoted as $H_{m,l}^a(\theta)$, and $H_{m,l}^a(\theta) = H^{tx}(\theta)H_l^c(\theta)A_m(\theta)$. Assuming the number of AFB is large, i.e., $M \to \infty$, the bandwidth of BPF $A_m(\theta)$ becomes arbitrarily narrow. The DTFT of $h_{m,l}^a(\mathrm{n})$ can be rewritten as:

$$H_{m,l}^a(\theta_m) \approx \gamma_m\beta_{m,l}A_m(\theta) \xleftrightarrow{IDTFT} \gamma_m\beta_{m,l}a_m(n) \approx h_{m,l}^a(n) \quad (4)$$

where, $\gamma_m = H^{tx}(\theta_m)$, $\beta_{m,l} = H_l^c(\theta_m)$. And, Eq (3) is written as:

$$x_{m,l}(n) = e^{-j\theta_m nD}\left[\sum_k S_k\gamma_m\beta_{m,l}\bar{a}_m(n-2k) + \bar{v}_{m,l}(n)\right] (5)$$

where, $\bar{a}_m(n) = a_m(nD)$, $\bar{v}_{m,l}(n) = v_{m,l}(nD)$.

4. DIVERSITY COMBINING AND EQUALIZER

Examine Eq. (5), the AFB has transformed wideband SC signal onto a collection of narrow band signals and this is true when M is sufficiently large for a given frequency selective channel. With this assumption, the multipath channel becomes a complex scalar gain $\beta_{m,l}$ for the m^{th} AFB output on the l^{th} diversity branch. The diversity combining for frequency selective channel can now be readily defined based on the existing narrow band diversity concepts.

For the l^{th} branch, define $X_l(n) \triangleq [x_{0,l}(n), x_{1,l}(n), \ldots x_{M-1,l}(n)]^T$ as the AFB outputs; $B_l \triangleq [\beta_{0,l}, \beta_{1,l}, \ldots \beta_{M-1,l}]^T$ as the channel gain; $\mathbb{W}_l = diag\{w_{0,l} \ldots w_{M-1,l}\}$ as the complex scalar weights applied to AFB outputs. Thus, the diversity combined signal is $X(n) \triangleq [x_0(n), x_1(n), \ldots x_{M-1}(n)]^T$ is written as:

$$X(n) = \sum_l \mathrm{W}_l^H X_l(n) \tag{6}$$

Based on Eq. (6), one immediately recognizes from narrow band combining concepts that setting $\mathbb{W}_l = diag\{\beta_{0,l} \ldots \beta_{M-1,l}\}$ produces the MRC [8]. However, the MRC requires precise channel knowledge, which is difficult to obtain for non-CP SC systems. Therefore, our focus naturally turns to the FB based selection diversity. Take the m^{th} AFB output for example; one can select the signal with the highest power among the L available branches. And the combiner's weights can be represented as:

$$w_{m,l} = \begin{cases} 1 \ , if\ |\beta_{m,l}| = \max(|\beta_{m,1}|, \ldots, |\beta_{m,L}|) \\ 0 \ , otherwise \end{cases} \tag{7}$$

It is clear that, regardless of the combining rule,the mechanism i.e., Fig.2,essentially generates a new channelthat is supposed to enhance the overall system performance. This new channel $\hat{B}$, expressed as M-by-1 vector, can be expressed as:

$$\hat{B} = \sum_l \mathrm{W}_l^H B_l = [\hat{\beta}_0, \hat{\beta}_1 \ldots \hat{\beta}_{M-1}]^T \tag{8}$$

For the MRC, the m^{th} entry of $\hat{B}$ is: $\hat{\beta}_m = \sum_l |\beta_{m,l}|^2$; and for the selection diversity the m^{th} entry of $\hat{B}$ is $\hat{\beta}_m = \beta_m^{Max}$, where β_m^{Max} is the channel gain associated with the signal that has the highest power across the L branches. The combined signal observed on the m^{th} AFB path is written as

$$x_m(n) = e^{-j\theta_m nD}\left[\sum_k S_k \gamma_m \hat{\beta}_m \bar{a}_m(n-2k) + \bar{u}_m(n)\right] \tag{9}$$

where the noise term $\bar{u}_m(n) = \sum_l w_{m,l}^* \bar{v}_{m,l}(n)$.

Assuming a perfect copy of the matched filtered signal free of channel and noise is available, denoted as $x_m^{ref}(n)$:

$$x_m^{ref}(n) = e^{-j\theta_m nD} \sum_k S_k |\gamma_m|^2 \overline{a_m}(n-2k) \tag{10}$$

And we denote reference signal column vector $X^{ref}(n) \triangleq [x_0^{ref}(n), x_1^{ref}(n), \ldots x_{M-1}^{ref}(n)]^T$. In practice, the reference signal is produced via decision directed process, a standard process in equalizer design [8]. The MMSE linear equalizer is produced by solving the following optimization problem:

$$\arg\min_K \mathbf{J}(K) = \arg\min_K E\left\|\mathbf{K}^H X(n) - X^{ref}(n)\right\|_2^2$$

where, the equalizer's coefficient $\mathbf{K} \triangleq diag\{k_0, k_1, \ldots k_{M-1}\}$ are to be determined. Since the AFBs has decoupled the input signal into M outputs. The optimization problem can be solved by examining one channel, say the m^{th} channel. In the case $M \to \infty$, the equalizer weight for the m^{th} channel can be solved by minimizing the following function:

$$\begin{aligned} \arg\min_K J_m(k_m) &= \arg\min_K E\left\|k_m^* x_m(n) - x_m^{ref}(n)\right\|_2^2 \\ &= \arg\min_K E\left[\varepsilon_m^f \varepsilon_m^{f*}\right] \end{aligned} \tag{11}$$

where $\varepsilon_m^f \triangleq k_m^* x_m(n) - x_m^{ref}(n)$. After simple steps, the optimum coefficient k_m^{opt} can be written as:

$$k_m^{opt} = \frac{E\{x_m(n) x_m^{ref*}(n)\}}{E\{x_m(n) x_m^*(n)\}}$$

And, one may calculate quantities:

$$E\{\bar{v}_m(n)\bar{v}_m^*(n)\} = \sigma_n^2 \sum_k |\bar{a}_m(n-k)|^2 = \sigma_n^2 \eta_1,$$

$$\sigma_u^2 = E\{\bar{u}_m(n)\bar{u}_m^*(n)\} = \begin{cases} \sigma_n^2 \eta_1 & , Selection\ Combining \\ \sigma_n^2 \eta_1 \sum_l |\beta_{m,l}|^2 & , MRC \end{cases}$$

$$E\{x_m(n)x_m^*(n)\} = |\gamma_m|^2 |\hat{\beta}_m|^2 \sigma_s^2 \eta_2 + \sigma_u^2,$$

$$E\{x_m(n)x_m^{ref*}(n)\} = |\gamma_m|^2 \gamma_m \hat{\beta}_m \sigma_s^2 \eta_2$$

where $\eta_1 = \sum_k |\bar{a}_m(n-k)|^2$; $\eta_2 = \sum_k |\bar{a}_m(n-2k)|^2$. Note, η_1 is the decimated AFB prototype filter norm; and we also have $\eta_2 = \eta_1/2$, i.e., $\bar{a}_m(2k)$ is the two path polyphase partition of $\bar{a}_m(k)$. We then find the optimum k_m^{Opt} as

$$k_m^{opt} = \frac{\gamma_m \hat{\beta}_m}{|\hat{\beta}_m|^2 + \dfrac{\sigma_u^2}{\sigma_s^2 \eta_2 |\gamma_m|^2}} \tag{13}$$

Examine Eq. (13), the term γ_m is the fixed SRRC part; and the rest is the MMSE equalizer based on channel $\hat{\beta}_m$. The path-wise MMSE can be found by plugging Eq. (13) into Eq. (11), we have

$$J_m(k_m^{Opt}) = \frac{\sigma_u^2 |\gamma_m|^2}{|\hat{\beta}_m|^2 + \dfrac{\sigma_u^2}{\sigma_s^2 \eta_2 |\gamma_m|^2}} \tag{14}$$

And, one can use corresponding $\hat{B}$ and σ_u^2 to produce the equalizer coefficient for selection combining and MRC.

$$J_m^{SEL}(k_m^{Opt}) = \frac{\sigma_n^2 \eta_1 |\gamma_m|^2}{|\beta_m^{Max}|^2 + \dfrac{\sigma_n^2 \eta_1}{\sigma_s^2 \eta_2 |\gamma_m|^2}} \quad J_m^{MRC}(k_m^{Opt}) = \frac{\sigma_n^2 \eta_1 |\gamma_m|^2}{\sum_l |\beta_{m,l}|^2 + \dfrac{\sigma_n^2 \eta_1}{\sigma_s^2 \eta_2 |\gamma_m|^2}}$$

Clearly, $J_m^{SEL}(k_m^{opt}) \geq J_m^{MRC}(k_m^{opt})$, since $|\beta_m^{Max}|^2 \leq \sum_l |\beta_{m,l}|^2$. And, the total MMSE is simply defined as: $J_{min}^{SEL} = \sum_m J_m^{SEL}(k_m^{opt})$; and $J_{min}^{MRC} = \sum_m J_m^{MRC}(k_m^{opt})$.

5. NUMERICAL RESULTS

We have derived the selection combining scheme in the last section; and we have demonstrated that the selection combining is suboptimal compared to MRC in the MMSE sense. A detailed simulation on MMSE based on single channel realization will be conducted to verify the derived results and BER result averaged over statistical channel model will also be provided.

5.1. MMSE Study over Single Channel Realization

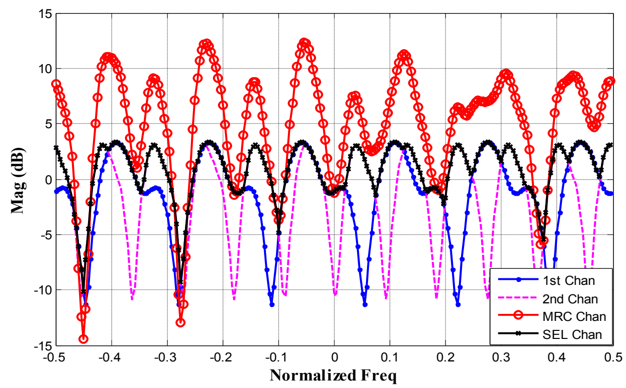

Fig. 4.Magnitude response 1st Chan, 2nd Chan and Combined Channels

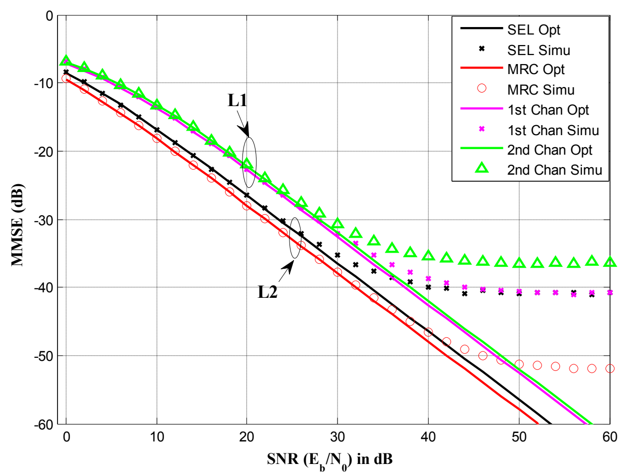

Fig. 5.MMSE after Linear Equalization for 1st Chan, 2nd Chan and Combined Channels

A 240-path;D = 80;8 taps per polyphase arm LPPF; PR-NMDFB is constructed to perform diversity combining and linear equalization. The input signal is SRRC shaped QPSK sampled at 2 samples per symbol with roll off factor being 25%. The channel on the 1st branch, denoted as "1st Chan" has impulse response 0.8638 + 0.4319i Z^{-6} + 0.2591i Z^{-12} + 0.008638i Z^{-18}; and the channel on the 2nd branch, denoted as "2nd Chan" has impulse response 0.8352 - 0.5429Z^{-11} - 0.0835iZ^{22} + 0.0251iZ^{-33}. Fig . 4 shows the magnitude response of the two channels along with the combined channel via both MRC and selection criteria based on Eq. (8). Fig. 5 shows the theoretically achievable MMSE (Eq. (14)) and simulated MMSE for selection, MRC, 1st Chan and 2nd Chan. Clearly the 2-branch diversity techniques outperform non diversity receivers by 4dB to 5 dB. In the realistic E_b/N_0 range, i.e., from 0 dB to 30 dB, we found the MMSE of selection diversity is only 1.25 dB worse than MRC. The result shows the FB based selection technique is more than acceptable given the fact it does not require any channel knowledge.

5.2. BER Results over Statistical Channel Model

Consider QPSK signal with 50 MHz symbol rate communicating over ITU-R M.1225 indoor office channel B [9]. The ITU channel B has RMS delay spread of 100 ns, meaning the normalized delay spread is$\tau_{rms}/T = 5$. Fig.6 shows the BER performance averaged over 10^3 channel realizations. And, we can see the MRC has approximately 1 dB E_b/N_0 advantage over FB selection approach. And, the diversity technique outperforms single branch receiver by more than4dB.We also included the flat fading QPSK BER for 1 and 2 branches for comparison, which equals to conventional OFDM BER[10]. We can see that given perfect linear equalization, the SC has lower BER than conventional OFDM.

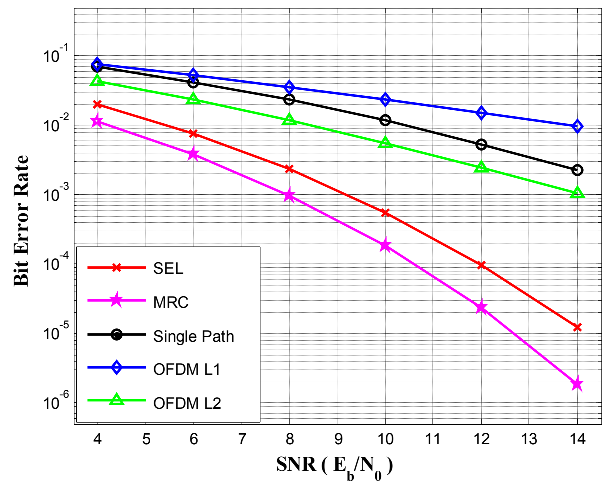

Fig. 6.BER over ITU-B Channel with 50 MHz Symbol Rate

5.3. Implementation Complexity

The FB based diversity receiver has L PACs and 1 PSC. It has L IPEs associated with each PAC performing the combining; and 1 IPE perform equalization and SRRC filtering. The M =240, D = 80, PR-NMDFB has LPPF of 1920 taps, or 8 taps per polyphase arm, which supports 90 dB dynamic range. For every 80 complex inputs, all PAC, PSC operate once ($L+1$ polyphase filters and 240-pt FFT) and uses $\frac{(L+1)}{80}[1920 \times 2 + 1100 + 240 \times 4] = (L+1)73.75$ real multiplies, where the 240-pt complex FFT costs 1100 real multiplies [Table 2-6, 11]. Setting L = 2, the count is 222, corresponding to an FIR filter with 111 taps processing complex data; yet we have implemented combining and equalization!

6. CONCLUSION

We have proposed the FB based selection diversity over frequency selective channel. Detailed MMSE performance analysis of the proposed and MRC approach is presented, which well matches the numerical results.Both theory andsimulation show the FB based selection diversity only suffers marginal performance loss compared to MRC. However, it does not require channel knowledge, which permits practical implementation. The authors believe FB selection approach and PR-NMDFB receiver is the key solution for future broadband SC transmission.

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