DOPPLER-RESISTENT OFDM RECEIVERS FOR MOBILE MULTIMEDIA COMMUNICATIONS

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Abstract[1] - Mobile Multimedia Communication involves the transmission of high data rates over rapidly time varying channels. The paper proposes a new approach to mitigate the effect of time variations and simulates two new receiver structures. The simulation model is based on a Discrete-Frequency channel representation for the link between the input of the transmit I-FFT and the output of the receive FFT. It exploits a Taylor expansion of the time variations of the received subcarrier amplitudes. Our receiver method leads to a substantial improvement of the link performance.

1. Introduction

Orthogonal Frequency Division Multiplexing (OFDM) is a modulation method designed in the 1970's [1-6] in which multiple user symbols are transmitted in parallel using different subcarriers. Although these subcarriers have overlapping (sinc-shaped) spectra, the signal waveforms are orthogonal. Compared to other modulation methods, OFDM uses symbols with a relatively long time duration, but a narrow bandwidth. Mostly, OFDM systems are designed such that each subcarrier is small enough in bandwidth to experience frequency-flat fading. This also ensures that the subcarriers remain orthogonal when received over a (moderately) frequency selective but time-invariant channel. If the OFDM signal is received over such a channel, each subcarrier experiences a different attenuation, but no dispersion.

The relatively straightforward receiver architecture associated with this has been a prime motivation to use OFDM in several standards, such as Digital Audio Broadcasting (DAB), e.g. [5], Digital Video Broadcast - Terrestrial (DVB-T), and more recently HIPERLAN II. For Mobile Multimedia, i.e., for high data rate communication over rapidly fading channels, OFDM has the problem of being highly sensitive Doppler spreads.

Time variations are known to corrupt the orthogonality of the OFDM subcarrier waveforms [2,15]. In such case, Inter-Carrier Interference (ICI) also called "FFT leakage", occurs because signal components from one subcarrier spill into other, mostly to neighboring subcarriers. Previously the problem of OFDM with Doppler was addressed in [7, 8, 13, 16, 17, 18, and 19].

The basic idea to repair InterCarrier Interference by adaptively combining subcarrier signals is intuitively appealing and can work if the delay spread is negligibly small. In such case all subcarriers experience the same amplitude and phase shifts, thus the ICI arrives with same crosstalk coefficients. However, for channels with both a delay and a Doppler spread, a practical implementation is more complicated, as it would typically require a large matrix multiplication. This matrix has its dominant contributions along the main diagonal, but is not of a Toeplitz structure that would allow a delay-line filter structure. Moreover, extensive channel estimation poses problems in practice. So the prime challenge is to find a realistic channel representation which allows a computationally attractive implementation. We found this in the form of time-derivatives of amplitudes of subcarriers. This paper

- Further develops the channel modeling for OFDM and Multi-Carrier systems under Doppler conditions
- Proposes a new set of channel parameters on which the design of mobile OFDM receivers can be based
- Gives two examples of such receivers, and
- Provides simulation results for these receivers

Our approach allows substantial improvements of the link performance at limited receiver complexity. If used for DVB-T, it would not require requiring any modification to the transmit standard. If the solutions are considered in the definition of future mobile multimedia or fourth generation mobile systems (4G), it may allow the use of OFDM over a much wider range of parameters than hitherto believed

[1] Portions of this paper will be presented at PIMRC 2000, London, September 2000.

2. Channel Model

This section develops a model for a channel representation starting from the classic multipath description, using as a collection of I_w reflected waves. Each wave has its particular Doppler frequency offset ω_i, path delay T_i and amplitude D_i. The frequency offset lies within the Doppler spread $-2\pi f_\Delta < \omega_i < 2\pi f_\Delta$, with $f_\Delta = v_s f_c/c$. Here v_s is the velocity of the mobile antenna, c is the speed of light and the carrier frequency is $\omega_c = 2\pi f_c$. More precisely, the Doppler shift of the i-th wave is $\omega_i = (2\pi v_s/c)\, f_c \cos(\theta_i)$ with θ_i the angle of arrival. Let f_s denote the spacing between the adjacent subcarriers and $\omega_s = 2\pi f_s$. The received signal equals (1)

$$r(t) = \sum_{n=0}^{N-1} \sum_{i=0}^{I_w-1} a_n D_i \exp\{ j(w_c + nw_s)(t - T_i) + jw_i t\} + n(t)$$

here $n(t)$ is additive white Gaussian noise. We will denote the vector of modulation symbols as $\boldsymbol{A} = [a_0, a_1, \ldots a_{N-1}]^T$. The transmit energy per subcarrier is $E_N = \mathrm{E}\,|a_n|^2$. Not all reflected waves are individually 'resolvable', that is, a receiver sampling in a time-window of finite duration will only see the collective effect of multiple reflected waves within a certain time-frequency window.

The narrowband mobile channel model can be compacted into a complex received amplitude that is time varying. We rewrite (1) as (2a)

$$r(t) = \sum_{n=0}^{N-1} a_n V_n(t) \exp\{ j(w_c + nw_s)t\} + n(t)$$

where the time-varying channel amplitude $V_n(t)$ at the n-th subcarrier is (2b)

$$V_n(t) = \sum_{i=0}^{I_w-1} D_i \exp\{ -j(w_c + nw_s)T_i + jw_i t\}$$

A Tayler expansion is $V_n(t) = v_n^{(0)}(t_0) + v_n^{(1)}(t_0)\ (t - t_0) + v_n^{(2)}(t_0)\ (t - t_0)^2 / 2 + \ldots$. Here $v_n^{(q)}(t_0)$ denotes the q-th derivative of the amplitude with respect to time, at instant $t = t_0$. If the Doppler spread is much smaller than the frequency resolution of the FFT grid, we may restrict our analysis to zero and first order effects. In a Rayleigh channel, $v_n^{(q)}$ is zero-mean complex Gaussian for any n and q. To characterize the channel, we are interested in the covariance of variables $v_n^{(p)}$ and $v_m^{(q)}$, viz., $\mathrm{E} v_n^{(p)} v_m^{*(q)}$, where * denotes the complex conjugate. In [20], we have derived this for an exponential delay spread and a uniform angle of arrival. The covariance becomes, for $p + q$ even, (3)

$$\mathrm{E}\, v_n^{(p)} v_m^{*(q)} = \left(2pf_\Delta\right)^{p+q} \frac{(p+q-1)!!}{(p+q)!!} \frac{(-1)^q j^{p+q}}{1 + j\Delta T_{rms} w_s}$$

and $\mathrm{E} v_n^{(p)} v_m^{*(q)} = 0$ for $p + q$ is odd.

3. Discrete-Frequency Link Model

In OFDM, a frame of N symbols is detected by taking N samples and performing an FFT. In other words, detection of the signal at subcarrier m occurs by correlation with a complex sinusoid having the frequency of the m-th subcarrier, viz., $\exp\{-j\omega_c t - j(m-\Delta_f)\omega_s t\}$ during the symbol duration NT. Here Δ_f is the frequency offset normalized to the subcarrier spacing ω_s. In the sampling process, the receiver makes a timing error Δ_t, where Δ_t is the time offset normalized to a sample interval $T = (N f_s)^{-1}$. That is, it samples at $t = \Delta_t, \Delta_t + T, \Delta_t + 2T, \ldots$. We assume that cyclic prefixes avoid any interframe interference. We use $\boldsymbol{Y}$ as the vector of length N to denote the output of the FFT in the receiver. The m-th output of the FFT is found as (4)

$$y_m = \frac{1}{N} \sum_{k=0}^{N-1} r(kT + \Delta_t) \exp\{ -j(w_c + w_s(m + \Delta_f))(kT + \Delta_t)\} + n_m$$

Here, $m = 0, 1, \ldots, N - 1$ and $\boldsymbol{Y} = [y_0, y_1, \ldots, y_{N-1}]^T$. We observe that $\omega_s T = 2\pi/N$. We introduce the OFDM system parameter $\xi_\Delta^{(q)}$, defined as

$$\hat{\imath}_\Delta^{(q)} = \frac{1}{N} \sum_{k=0}^{N-1} k^q \exp\left\{ \frac{2pjk\Delta}{N} \right\}, \qquad (5)$$

to describe the signal transfer over the q-th derivative of the amplitude at subcarier n to the $(n+\Delta)$-th receive subcarrier. This allows us to rewrite y_m as follows: (6)

$$y_m = \frac{1}{N} \sum_{n}^{N-1} a_n e^{j2\eth\left(f_c - \frac{\Delta_f}{N} + nf_s \right)\Delta_t} \sum_{q=0}^{\infty} \frac{v_n^{(q)} \hat{\imath}_{n-m-\Delta_f}^{(q)} T^q}{q!} + n_m$$

In particular, we will consider the first two terms of the expansion, and we denote $\chi_\Delta = \xi_\Delta^{(0)}$ and $\zeta_\Delta = \xi_\Delta^{(1)}$. So, (7)

$$c_\Delta = \frac{1}{N} \sum_{k=0}^{N-1} \exp\left\{ \frac{2pjk\Delta}{N} \right\} = \frac{1}{N} \frac{1 - \exp\{j2p\Delta\}}{1 - \exp\{j2p\Delta / N\}}$$

For integer Δ, χ_Δ reduces to δ_Δ, which is just a confirmation that subcarriers (with a nonfading amplitude) are orthogonal. Note further that

$$z_\Delta = -\frac{jN}{2p} \frac{c_\Delta}{d\Delta} \qquad (8)$$

for integer and non-zero Δ, we see that

$$z_\Delta = \begin{cases} (N-1)/2, & \Delta = 0 \bmod N \\ -\left(1 - \exp\{j2p\Delta/N\}\right)^{-1}, & \Delta \neq 0 \bmod N \end{cases} \qquad (9)$$

Roughly speaking $\zeta_\Delta / \zeta_0 \approx (j\pi\Delta)^{-1}$, with $\Delta = 1, 2, 3, \ldots$, so the ICI reduces slowly with increasing subcarrier separation. Relatively many subcarriers make a significant contribution to the ICI. We define vector $\boldsymbol{V} = [v_0, v_1, .., v_{N-1}]$ for the subcarrier amplitudes and $\boldsymbol{V'} = [v_0^{(1)}, v_1^{(1)}, .., v_{N-1}^{(1)}]^T$ for the derivatives. For an ideally synchronizing receiver, i.e., with $\Delta_t = 0$ and $\Delta_f = 0$,, the received signal $\boldsymbol{Y}$ can compactly be written in Discrete Frequency domain as, (10)

$$\boldsymbol{Y} = (\mathrm{DIAG}(\boldsymbol{V}) + T\,\Xi\,\mathrm{DIAG}(\boldsymbol{V}'))\,\boldsymbol{A} + \boldsymbol{N}$$

with DIAG(X) the diagonal matrix composed of the elements of vector X, and (11)

$$\Xi = \begin{bmatrix} z_0 & z_1 & .. & z_{N-1} \\ z_{-1} & z_0 & .. & z_{N-2} \\ .. & .. & .. & .. \\ z_{-N+1} & z_{-N+2} & .. & z_0 \end{bmatrix}$$

Not all terms in Ξ address ICI: In (11) the diagonal terms ζ_0 carry signal components from the n subcarrier to the n-th subcarrier, so in fact the receiver sees a signal amplitude of $(\chi_0 v_n + \zeta_0 T v_n^{(1)}) a_n$. To address this in the following calculations we define $\Xi^* = \Xi - \zeta_0 \boldsymbol{I}_N$. The wanted signal component in a conventional receiver equals $\boldsymbol{V}+\zeta_0\boldsymbol{V'}T$, which in practice closely approximates $\boldsymbol{V}$, except in deep fades.
Figure 1 depicts the channel and receiver in the discrete frequency domain. Thus, the FFT is not drawn explicitly. In a conventional system, $\boldsymbol{W}$ represents the equalizer, or automatic gain control per subcarrier, to compensate for fading on the subcarriers.

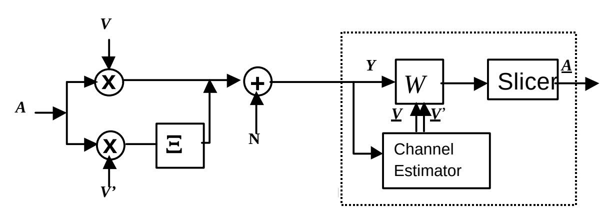


Figure 1: Discrete-Frequency representation for the Doppler multipath channel and a (feedforward) OFDM receiver

4. Receiver I: MMSE ICI canceller

The OFDM receiver sees the signal $\boldsymbol{Y} = \boldsymbol{Q}\,\boldsymbol{A} + \boldsymbol{N}$, with matrix $\boldsymbol{Q} = \mathrm{DIAG}(\boldsymbol{V}) + T\,\Xi\,\mathrm{DIAG}(\boldsymbol{V'})$. Our OFDM receiver is extended such that it can not only reliably estimate amplitudes (as conventional receivers do), but also complex valued derivatives (which is not common for normal OFDM receivers), with $\underline{\boldsymbol{V}}$ and $\underline{\boldsymbol{V}}'$ the estimate of amplitudes and derivatives, respectively. Then the data can be recovered as follows:

- Create the matrix $\underline{\boldsymbol{Q}} = \mathrm{DIAG}(\underline{\boldsymbol{V}}) + T\,\Xi\,\mathrm{DIAG}(\underline{\boldsymbol{V}}')$.
- Compute an MMSE filter $\boldsymbol{W}$ according to $\boldsymbol{W} = \underline{\boldsymbol{Q}}^H [\,\underline{\boldsymbol{Q}}\,\underline{\boldsymbol{Q}}^H + N_0 I_N]^{-1}$.

For a receiver that perfectly estimates the channel, the covariance matrix of the residual ICI plus noise, normalized to unity signal power, becomes

$$C = \left[I_N + \frac{E_n}{N_0} \boldsymbol{Q}^H \boldsymbol{Q} \right]^{-1} \qquad (12)$$

The vector of the SINR at the N subcarriers is (13)

$$\mathrm{diag}(\boldsymbol{Q}) .* \mathrm{diag}(\boldsymbol{Q}^H)\, E_N ./ \mathrm{diag}(\boldsymbol{C}),$$

where ./ is a subcarrier-by-subcarrier division and .* a subcarrier-by-subcarrier multiplication. A conventional OFDM link with known channel characteristics ($\boldsymbol{V}$, $\boldsymbol{V'}$) would have a noise plus ICI contribution described by covariance matrix: (14)

$$E_N\, T^2 \Xi^* \boldsymbol{V'} \boldsymbol{V'}^H \Xi^{*H} + N_0\, I_N,$$

The vector of SINR values is found as (15)

$$\mathrm{diag}(\boldsymbol{Q}).*\mathrm{diag}(\boldsymbol{Q}^H)\, E_N./\mathrm{diag}(T^2 \Xi^* \boldsymbol{V'} \boldsymbol{V'}^H \Xi^{*H} E_N + N_0 \boldsymbol{I}_N).$$

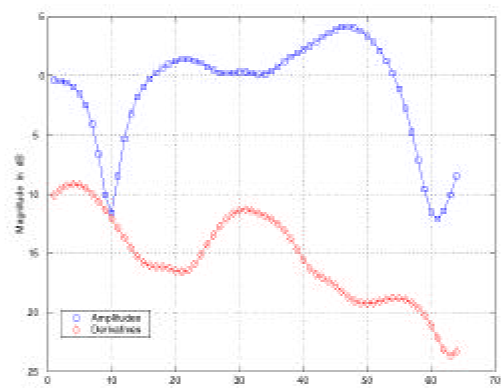
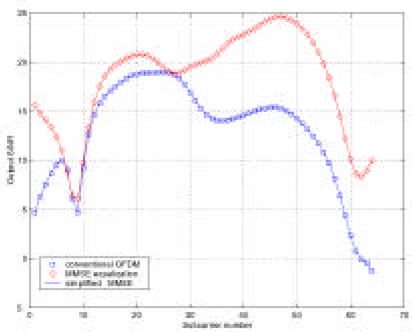


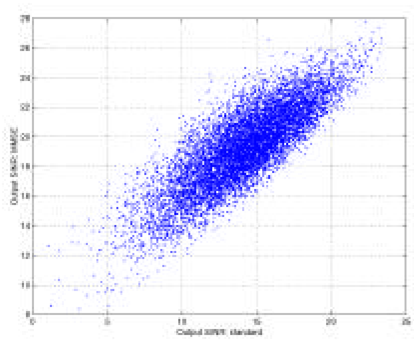
Figure 2: Amplitudes $|\boldsymbol{V}+\zeta_0\boldsymbol{V'}T|$ and derivatives $\zeta_0 T|\boldsymbol{V'}|$, as a function of subcarrier number for a sample channel

We simulated this receiver in Discrete-Frequency domain, extending the method presented in [12]: We generate complex fading amplitudes of a Rayleigh channel with Doppler and delay spread from an i.i.d vector of two complex Gaussian random variables $\boldsymbol{G}$ and $\boldsymbol{G}'$, both with unity variance and length N. The vector $\boldsymbol{G}$ was then multiplied by a precomputed N-by-N matrix $\boldsymbol{U}$, such that $\boldsymbol{U}\boldsymbol{U}^H$ is the channel covariance matrix Γ with elements $\Gamma_{n,m} = \mathrm{E} v_n^{(0)} v_m^{*(0)}$, to create $\boldsymbol{V} = \boldsymbol{U}\,\boldsymbol{G}$. Similarly the derivatives are generated from $\boldsymbol{G'}$ according to $\boldsymbol{V'} = 2\pi f_\Delta T\,\boldsymbol{U}\,\boldsymbol{G'}$.

We simulated a HIPERLAN II type of system under extreme Doppler conditions. A carrier frequency of f_c = 17 GHz, 200 km/h, N = 64 subcarriers, transmit bandwidth of 2 MHz (T = 0.5 µs) and a delay spread of T_{RMS} = 1 µs. E_N / N_0 = 100 (20 dB). The Doppler spread is f_Δ = 3.148 kHz and f_s = 31.25 kHz. So the average SIR for a conventional OFDM system would be around 18 dB. These parameters correspond to the level of ICI experienced in an 8k DVB-T system at normal vehicle speeds. Figure 2 plots an example of the signal amplitude $|\boldsymbol{V}+\zeta_0 T\boldsymbol{V'}T|$ and $\zeta_0 T|\boldsymbol{V'}|$ as a function of the subcarrier number. Figure 3 gives the SINR after the equalization filter for a conventional OFDM system and our proposed solution.



Figures 3: (ð) Signal-to-ICI-plus-noise ratio in conventional system, (◊) SINR in new system as a function of subcarrier number for the channel in Figure 2.



Figures 4: Comparison of SNR in conventional system and SNR in a system with ICI canceller. Every dot corresponds to one sample channel with *N* subcarriers.

6. Receiver II: DFE

We used a DFE receiver architecture inspired by the LSE structure. It allows iterative computing to minimize the variance of the error between input and output of the slicer.

The iteration method contains the following steps for iteration round *i*:

- Input: observation $\boldsymbol{Y}_0$, as well as the i-1-th estimate of amplitudes $\underline{\boldsymbol{V}}(i-1)$, derivatives $\underline{\boldsymbol{V}}'(i-1)$, and data $\underline{\boldsymbol{A}}(i-1)$
- Calculation of $\boldsymbol{Y}_2(i)$, using previous estimates of derivatives $\underline{\boldsymbol{V}}'(i-1)$ and data $\underline{\boldsymbol{A}}(i-1)$, using $\boldsymbol{Y}_2(i) = \boldsymbol{Y}_0 - \Xi\ (\underline{\boldsymbol{V}}'(i-1)\ .*\ \underline{\boldsymbol{A}}(i-1))$
- New estimate amplitudes $\underline{\boldsymbol{V}}(i)$ from $\boldsymbol{Y}_2(i)$
- New estimate of data $\underline{\boldsymbol{A}}(i)$, and the corresponding values of $1/\underline{\boldsymbol{A}}(i)$, and $\underline{\boldsymbol{V}}(i)\ .*\ \underline{\boldsymbol{A}}(i)$
- Calculation of $\boldsymbol{Z}_6(i)$, $\boldsymbol{Z}_7(i)$, $\boldsymbol{Z}_8(i)$, $\boldsymbol{Z}_9(i)$, Here $\boldsymbol{Z}_9(i)$ acts as the estimate of $\underline{\boldsymbol{V}}'$
- Integration of $\boldsymbol{Z}_9(i)$ over various rounds, for instance $\underline{\boldsymbol{V}}'(i) = \alpha\underline{\boldsymbol{V}}'(i-1) + (1-\alpha)\boldsymbol{Z}_9(i)$
- Output: new estimate of amplitudes $\underline{\boldsymbol{V}}(i)$, derivatives $\underline{\boldsymbol{V}}'(i)$, and data $\underline{\boldsymbol{A}}(i)$

Starting condition is the all-zero vector for $\underline{\boldsymbol{V}}(0)$, $\underline{\boldsymbol{V}}'(0)$ and $\underline{\boldsymbol{A}}(0)$.

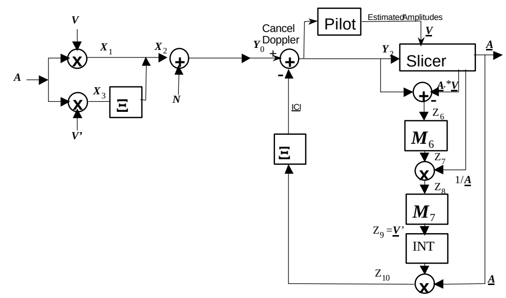


Figure 5: Discrete-frequency domain representation of channel and DFE receiver

The following paragraph addresses the selection of useful choices for the filter in the DFE loop. Filter $\boldsymbol{M}_6$ attempts to recover $\boldsymbol{X}_3$ from $\boldsymbol{Z}_6$ by filtering $\boldsymbol{Z}_7 = \boldsymbol{M}_6\,\boldsymbol{Z}_6$. Using the orthogonality principle, an appropriate (MMSE-like) choice for $\boldsymbol{M}_6$ follows from the requirement

$$\mathrm{E}[(\boldsymbol{Z}_7 - \boldsymbol{X}_3)\,\boldsymbol{Z}_6^{\mathrm{H}}] = 0_N$$

We define $\boldsymbol{e}$ as the vector of decision errors, with $\boldsymbol{e} = \boldsymbol{V}.*\boldsymbol{A} - \underline{\boldsymbol{V}}.*\underline{\boldsymbol{A}}$ so for $\boldsymbol{Z}_6$ we write $\boldsymbol{Z}_6 = \Xi\,\boldsymbol{X}_3 + \boldsymbol{N} + \boldsymbol{e}$, and insert this in

$$\boldsymbol{M}_6\,\mathrm{E}[\boldsymbol{Z}_6\,\boldsymbol{Z}_6^{\mathrm{H}}] = \mathrm{E}[\boldsymbol{X}_3\,(\Xi\,\boldsymbol{X}_3 + \boldsymbol{N} + \boldsymbol{e})^{\mathrm{H}}]$$

Modeling of the terms $\mathrm{E}\boldsymbol{e}_1\boldsymbol{e}_1^{\mathrm{H}}$, $\mathrm{E}\boldsymbol{X}_3\boldsymbol{e}_1^{\mathrm{H}}$ and $\mathrm{E}\boldsymbol{X}_3^{\mathrm{H}}\Xi^{\mathrm{H}}\boldsymbol{e}_1$ is not straightforward, so next we will search for simplifying approximations. If $\boldsymbol{e}$ were statistically independent of $\boldsymbol{X}_3$, one can significantly simplify these expressions. Although such assumption is questionable, it can lead to a simple and practical receiver design. We simplify the resulting $\boldsymbol{M}_6$ as

$$\boldsymbol{M}_6 = \Xi^{\mathrm{H}}\,[\,\mathrm{E}[\Xi\,\Xi^{\mathrm{H}} + \boldsymbol{G}\,]^{-1}$$

where $\boldsymbol{G}$ is chosen as $\boldsymbol{G} = c_1\,\boldsymbol{I}_{\mathrm{N}}$. Exploiting results from a derivation of the OFDM error probability in fading channels [20] which is omitted here, a suitable estimate appears to be

$$c_1 = 2\,E_N^{-1}(2\pi f_\Delta T)^{-2}\,(N_0 + E_N\,\sigma_{\mathrm{error}}^{2} / (4(E_N / N_0 + 1))\,)$$

where σ_{error} is the standard deviation for the error in the event that the slicer makes a symbol error, E_N / N_0 is the average (local-mean) signal to noise ratio (per subcarrier)

As far as $\boldsymbol{M}_7$ is concerned, $\boldsymbol{Z}_8$ approximates $\boldsymbol{V}$' however it contains error contributions, due to AWGN and estimation error in $\underline{\boldsymbol{A}}$ and $\underline{\boldsymbol{V}}$. An MMSE filter to estimate $\boldsymbol{Z}_9 = \underline{\boldsymbol{V}}'$ follows from E $(\boldsymbol{Z}_9 - \underline{\boldsymbol{V}}')\, \boldsymbol{Z}_8^{H} = 0_N$, thus $\boldsymbol{M}_7 = \mathrm{E}\underline{\boldsymbol{V}}'\underline{\boldsymbol{Z}}_8^{H}\, [\mathrm{E}\, \boldsymbol{Z}_8\, \boldsymbol{Z}_8^{H}]^{-1}$. We crudely approximate the error terms by assuming that these are independent of $\boldsymbol{V}$' and we only consider the main effects, by postulating $\boldsymbol{R}_3$ in

$$\boldsymbol{M}_2 = \mathrm{E}\underline{\boldsymbol{V}}'\underline{\boldsymbol{V}}'^{H}\, [\mathrm{E}\underline{\boldsymbol{V}}'\underline{\boldsymbol{V}}'^{H} + \boldsymbol{R}_3]^{-1}$$

Experiments revealed that $\boldsymbol{R}_3$ a constant times $\boldsymbol{I}_N$ is a workable solution.

We simulated the channel in discrete-frequency domain, as described before. We took the signal-to-ICI ratio equal to 1200 (31 dB). Local-mean noise is at –40 dB ($\sigma_n = 0.01\chi_0$). We considered a channel with normalized rms delay spread $T_{rms} / T_s = 0.03$. This determines the correlation between (derivatives of) amplitudes at different subcarriers. The modulation method is 16 QAM with signal levels at {–3, –1, 1, 3}. In the simulation, we assumed perfect estimates the subcarrier amplitudes. The estimation of derivatives is done iteratively by the DFE loop. The integration constant was set $\alpha = 0.9$, so $\underline{\boldsymbol{V}}'(i) = 0.9\ \underline{\boldsymbol{V}}'(i\text{-}1) + 0.1\ \boldsymbol{Z}_9(i)$. The constants have been chosen after a short series of experiments with different values.
In the example run of Figure 6, we see that a conventional system would make errors at subcarrier 3, 4 and 5. The DFE system corrects these errors. In Figure 7, every dot corresponds to one simulation for one particular channel. We see a significant improvement, often of about 10 dB.

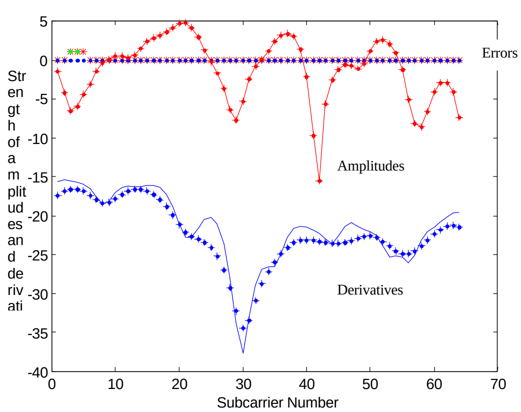


Figure 6: strength in dB of amplitudes 10 $\log_{10}$ diag($\boldsymbol{VV}^{H}$) and derivatives 10 $\log_{10}$ diag($\boldsymbol{V'V'}^{H}$) (upper and lower solid line) and their estimates (+,•) after one and many iterations respectively, versus subcarrier number. Also, on same vertical axis, the number of errors per subcarrier: *:errors by conventional system. (•) errors by proposed system.

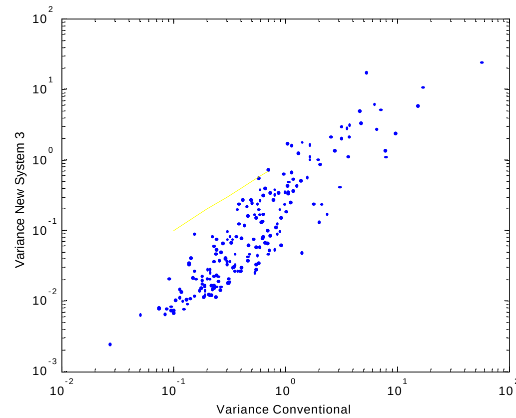


Figure 7: variance of signal before slicer, horizontal: conventional system. vertical: remaining variance after processing in a DFE receiver.

6 Conclusions

Doppler spread is well recognized as a major problem in mobile OFDM reception. Previously contermeasures to mitigate this problem have been searched mainly by considering (i.e., estimating and separating) individual resolvable frequency-shifted components. Estimation, separation and cancellation of these components can become a computationally intensive task, and the effect of signal recovery is sensitive to estimation errors. This paper took a different approach by modeling the time-varying subcarrier amplitude of each OFDM subcarrier channel as a Taylor expansion with amplitude and derivatives.

Our receiver uses estimated derivatives. This simplifies the computational burden on the receiver as it limits the number of channel parameters to be estimated. In the receiver evaluated here, we see that the ICI is largely eliminated.

References

1. S.B. Weinstein and P.M. Ebert, "Data Transmission by Frequency Division Multiplexing Using the Discrete Fourier Transform," IEEE Tans. Commun. Technol., Vol. COM-19, No.5, pp. 628-634, Oct. 1971.
2. L. Cimini, Jr. "Analysis and Simulation of a Digital Mobile Channel Using Orthogonal Frequency Division Multiplexing," IEEE Trans. Commun., Vol. COM-33, No.7, pp. 665-675, July 1985.
3. M. Alard and R. Lassalle, "Principles of Modulation and Channel Coding for Digital Broadcasting for Mobile Receivers," EBU Technical Review, No. 224, pp.168-190, Aug. 1987.
4. J.P. Linnartz and S. Hara, Special Issue on Multi-Carrier Communications, Wireless Personal Communications, Vol. 2 No. 1 & 2, 1995, pp. 1-7.

5. B. Le Floch, R. Halbert-Lassalle and D. Castelain, “Digital Sound Broadcasting to Mobile Receivers," IEEE Trans. Consumer Electronics., Vol. 73, No.4, pp.30-34, Aug. 1989.
6. Wireless Communication, The Interactive Multimedia CD-ROM, 3rd edition, Kluwer Academic Publ. http://www.baltzer.nl/wicom
7. P. Robertson, S. Kaiser, “Effects of Doppler Spreads in OFDM(A) Mobile Radio Systems” VTC Fall ’99, Amsterdam.
8. M. Russel and G. Stuber, “Terrestrial digital video broadcasting for mobile reception using OFDM”, Special Issue on MCC, Wireless Personal Communication, Vol. 2, No., 1&2, 1995, pp. 45-66.
9. W.C. Jakes, "Microwave Mobile Communications", John Wiley and Sons, New York, 1978. New edition 1994.
10. R.H. Clarke, “A statistical theory of mobile radio reception”, Bell Syst. Tech. J., Vol. 47, 1968, pp. 957-1000.
11. T. Aulin, “A modified model for the fading signal at a mobile radio channel”, IEEE Tr. on Veh. Tech., Vol. VT-28, 1979, No. 3, pp. 182-203.
12. N. Yee and J.P.M.G. Linnartz, "Wiener filtering for Multi-Carrier CDMA", conference on Personal Indoor Mobile Radio Communications (PIMRC), The Hague, September 19-23, 1994, Vol. 4, pp. 1344-1347.
13. J.P. Linnartz, “Performance analysis of MC-CDMA in Rayleigh channel with Doppler and delay spread”, IEEE Benelux Symp. on DSP, Hilvarenbeek, 2000.
14. G. Fettweis et al.,"On multi-Carrier CDMA modem design", Proc. of IEEE VTC'94, pp.1665-1669, Stockholm, Sweden, June 1994.
15. http://www.eecs.berkeley.edu/~linnartz/issue.html
16. J.P.M.G. Linnartz, "Performance analysis of synchronous MC-CDMA in mobile Rayleigh channel with both delay and Doppler spreads", submitted in 2000.
17. W.G. Jeon, K.H. Chang and Y.S. Cho, “An equalization technique for Orthogonal Frequency-Division Multiplexing in Time-variant multipath channels”, IEEE Tr. on Comm. Vol. 47, No. 1, Jan. 1999, pp. 27-32.
18. Y. Li, N. Seshadri, and S. Ariyavistakul, “”, IEEE JSAC, Vol. 17, No. 3, March 1999, pp. 461-471
19. Y. Li, L.J. Cimini, and N.R. Sollenberger, “Robust channel estimation for OFDM systems with rapid dispersive fading channels”, IEEE, Tr. on Comm. vol. 46, pp. 902-915, July 1998.
20. J.P.M.G. Linnartz and A. Gorokhov, "New Equalization approach for OFDM over dispersive and rapidly time varying channel", PIMRC 2000, London.