Enhanced Maximum Ratio Combining for Mobile DVB-T Reception in Doubly Selective Channels

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***Abstract*—Time varying channels cause a loss of orthogonality among the subcarriers of an Orthogonal Frequency Division Multiplexing (OFDM) symbol and lead to inter-carrier interference (ICI) and degradation in performance. In this work, we propose an enhanced Maximum Ratio Combining (MRC) scheme, which takes into consideration the colored nature of ICI in a frequency selective channel. In DVB-T/T2, continual pilots (CoPs) are used at the receiver side to estimate the noise power per receive antenna. We show that a significant gain can be achieved, when the noise power per subcarrier is fed to the MRC block instead of the average noise power per antenna. We propose to use a window over the set of CoPs to estimate the localized noise power instead of the average noise power. The optimum size of the window depends on many factors such as the Doppler shift, additive white Gaussian Noise (AWGN) power and channel frequency selectivity. Motivated by the unequal pilot spacing of the CoPs as well as the frequency correlation of the noise power in a doubly selective channel we propose a simple weighting criterion for the noise window samples, which provides a good performance over a wide range of SNR and Doppler shift.**

I. Introduction

OFDM has emerged as an attractive modulation scheme in many digital communication standards. It has the advantage of being robust in frequency selective channels along with a simple transceiver structure. An OFDM data stream is de-multiplexed into N parallel low rate substreams, which are modulated over N subcarriers. The bandwidth of each modulated subcarrier is small enough to assume a frequency flat fading channel. A guard interval (cyclic prefix) is inserted between successive OFDM symbols. If the channel impulse response is shorter than the length of the guard interval, inter-symbol interference (ISI) is avoided. In a time invariant channel, a one-tap zero forcing (ZF) equalizer can then recover the transmitted symbol on each subcarrier. However, the orthogonality among the OFDM subcarriers is destroyed when the OFDM symbol is transmitted over time-varying channels, resulting in ICI. The longer the duration of the OFDM symbol, the more sensitive it is to a time-varying channel. In DVB-T [1] an OFDM size of up to 8k subcarriers is used, and such a symbol lasts for about 1ms when broadcasted in an 8MHz channel.

In a wireless environment, the channel can be time-varying because of statistical multi-path propagation and the user mobility. The maximum Doppler frequency $f_{D,max}$ and normalized maximum Doppler frequency $f_{D,norm}$ are given by

$$f_{D,max} = \frac{f_c v}{c}, \tag{1}$$

$$f_{D,norm} = \frac{f_{D,max}}{\Delta f} \tag{2}$$

where f_c is the RF carrier frequency, v is the velocity of the user, c is the speed of light and Δf is the subcarrier spacing. In the rest of this paper, we drop the term 'maximum' when referring to the maximum Doppler frequency or the normalized maximum Doppler frequency. For more information about modeling a time-varying multi-path channel, the interested reader can refer to [2].

In [3] and [4], conventional antenna receiver diversity methods were examined as methods to mitigate the effect of ICI in mobile channels. In [5] and [6], an expression for the optimum MMSE diversity combiner is derived. In such a receiver, however, the interference off-diagonal coefficients have to be computed for each antenna in order to compute the interference matrix. These coefficients can be estimated from observing the rate of change of the pilot assisted channel estimates (PACE) [7], which requires extra computations in addition to the need for the buffering of these OFDM symbols. In [8], the authors present a simple MRC scheme which is more suited for a time varying channel. In the modified scheme, referred to as RC-UMMSE solution, the SINR per antenna (γ_m), where m is the antenna index, is used to adjust the weights of the MRC. In our work, we show that a significant gain can be achieved if the colored nature of the ICI in a doubly selective channel is further taken into consideration. In other words, we consider the SINR per antenna and frequency bin ($\gamma_{m,k}$), where k is the frequency bin index. We also propose a practical method to estimate the SINR per subcarrier for a DVB-T system. In [9], the interference plus noise power per subcarrier is computed for the case of co-channel interference and is used to optimize the MRC weights. Since a static channel is considered in [9], several preamble symbols were considered to obtain an estimate of the interference plus noise power per subcarrier. However, in a time varying channel and especially with long FFT sizes, such a method for estimating interference plus noise power is not valid. Therefore, in our work we obtain our estimated interference plus noise power from averaging the noise samples in the frequency domain only.

In DVB-T/T2, CoPs are part of the transmitted OFDM symbol and are used at the receiver side for functions such as synchronization and noise estimation. The locations of the CoPs is known at the receiver side. Conventionally as in [8], the noise power is computed on all CoPs and averaged to obtain one value per OFDM symbol. In [10], the problem of noise estimation in the presence of strong interference is considered. The authors assume a fully known OFDM symbol at the receiver side and proposed to estimate the noise via a 2D moving average window in time and frequency. In this work, we propose to estimate the noise power per subcarrier in a DVB-T system via applying a moving average window which includes a subset of the neighboring CoPs, rather than all CoPs, resulting in a localized estimate of the colored noise power. In a second step, we use the enhanced noise estimate in the MRC block.

In our problem, the optimum window size depends well on parameters such as the SNR, Doppler shift and channel frequency selectivity. In [11], an MMSE filtering technique is proposed to deal with noise estimation in the presence of colored noise. The filtering coefficients are derived as a function of the noise covariance matrix and the AWGN noise power. In reality, such statistics need to be estimated which puts an overhead at the receiver side. In our work, we propose a simple weighting criterion which decreased the dependency of the performance on the abovementioned parameters and provided a good performance over a wide range of SNR and Doppler shifts.

This paper is organized as follows: in section II, we introduce the system model and the ICI problem due to a time varying channel. In section III, we describe the conventional MRC. In section IV, we describe the weighting criterion we follow. In section V, we present simulation results for DVB-T system which show the effectiveness of the proposed MRC approach for a doubly selective channel.

II. System Model

The signal vector after the FFT at the receiver side, $\boldsymbol{R}_{N\times 1}$, can be written in matrix form as

$$\boldsymbol{R} = \boldsymbol{H}\boldsymbol{S} + \boldsymbol{X}, \tag{3}$$

where $\boldsymbol{H}$ is the 2D Discrete Fourier Transform (DFT) of the $N \times N$ time domain channel matrix $\boldsymbol{H}_t$ [12], $\boldsymbol{S}$ is the vector of transmitted symbols and $\boldsymbol{X}$ is the effective frequency domain additive white Gaussian noise (AWGN) vector. The channel matrix $\boldsymbol{H}$ can thus be computed as

$$\boldsymbol{H} = \boldsymbol{F}\boldsymbol{H}_t\boldsymbol{F}^H, \tag{4}$$

where $\boldsymbol{F}$ is the DFT matrix and $()^H$ denotes the Hermitian operation. $\boldsymbol{H}$ is composed of the elements $H(k,l)$, $k,l = 1,...,N$. In case of a time-invariant channel delay profile, $\boldsymbol{H}_t$ is a circulant matrix, and hence its 2D DFT, $\boldsymbol{H}$, is a diagonal matrix. The k^{th} element $R(k)$ of $\boldsymbol{R}$, can then be written as

$$R(k) = H(k,k)S(k) + X(k), \tag{5}$$

where $S(k)$ and $X(k)$ are the sent symbol and the effective frequency domain noise for subcarrier index k, respectively. $H(k,k)$ is the k^{th} element on the main diagonal of $\boldsymbol{H}$. For a time-invariant channel, a one-tap ZF equalizer is sufficient to estimate $S(k)$. In case of a time-varying channel, the off-diagonal frequency coefficients in $\boldsymbol{H}$ are no longer zero. Equation (5) then can be written as

$$R(k) = H(k,k)S(k) + \sum_{d\neq 0} H(k,(k+d)_N)S(k+d) + X(k), \tag{6}$$

where $(x)_N$ is modulo operator of x with respect to N. The

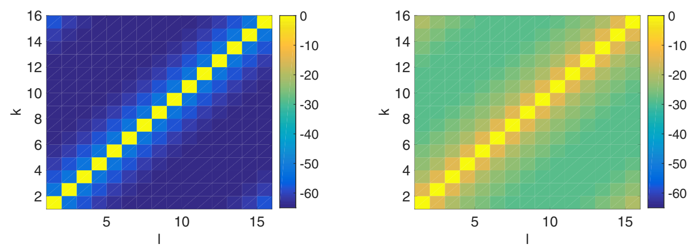


Fig. 1. The normalized matrix coefficients $\frac{|H(k,l)|}{|H_{max}|}$ in dB of a frequency flat fading channel matrix $\boldsymbol{H}$

larger the Doppler frequency, the larger the power leaking to the off-diagonal coefficients is. In Fig. 1, we can see the normalized magnitudes of the 16 point DFT channel matrix coefficients for a frequency flat Rayleigh fading channel at two different normalized Doppler frequency values, $f_{D,norm} = 0.001, 0.2$. The coefficients' magnitudes are normalized with respect to the magnitude of the largest main diagonal element $|H|_{max}$, $H_n(k,l) = \frac{H(k,l)}{|H_{max}|}$. We can see how the power is concentrated in the first few off-diagonal coefficients and decreases as the distance from the main diagonal increases. We can also see, how the edge subcarriers ($k = 1$ and $k = 16$) interfere with each other due to the circular nature of the FFT operation, in accordance with (6). As shown in Fig. 1, if the normalized Doppler frequency increases, the magnitudes of the interference off-diagonal coefficients increase. Consequently, ICI induces a higher error floor on the BER performance [4]. We define the interference signal as

$$\begin{aligned} I(k) &= R(k) - H(k,k)S(k) - X(k) \\ &= \sum_{d\neq 0} H(k,(k+d)_N)S(k+d). \end{aligned} \tag{7}$$

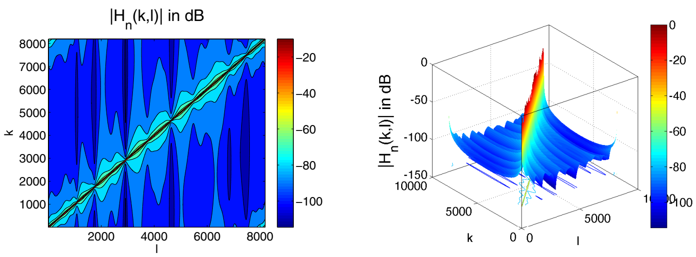
The expected value of the ICI power $I(k)$ can then be written as

$$\begin{aligned} E\left\{|I(k)|^2\right\} &= E\left\{I(k)I^*(k)\right\} \\ &= E\Bigg\{\sum_{d\neq 0} H(k,(k+d)_N)S(k+d) \\ &\quad \sum_{d\neq 0} H^*(k,(k+d)_N)S^*(k+d)\Bigg\}. \end{aligned} \tag{8}$$

Assuming an uncorrelated source, $E\{S(k)S^*(k+d)\} = 0$ $\forall d \neq 0$. Equation (8) can be written as

$$\begin{aligned} E\left\{|I(k)|^2\right\} &= E\left\{\sum_{d\neq 0} |H(k,(k+d)_N)|^2 |S(k+d)|^2\right\} \\ &= E_s \times E\left\{\sum_{d\neq 0} |H(k,(k+d)_N)|^2\right\}, \end{aligned} \tag{9}$$

where $E_s = E\left\{|S(k)|^2\right\}$. From (9), we can observe that the colored nature of the ICI power follows from the colored nature of the off-diagonal ICI coefficients. In other words, in a flat fading channel such as shown in Fig. 1, the ICI power is the same over all subcarriers, whereas in a frequency selective channel the coloring of the ICI follows the frequency selectivity of the channel. Fig. 2 depicts the magnitudes of the normalized coefficients $|H_n(k,l)|$ for a frequency selective channel, namely the Typical Urban 6 tap channel (TU6) as defined in [13], using an 8k DFT. As shown, the frequency selectivity is visible in the main diagonal, as well as for the off-diagonal coefficients. In Fig. 3, the normalized ICI power is plotted per subcarrier for 64QAM constellation in a flat fading channel as well in a frequency selective TU6 channel. We can see the frequency selectivity of the ICI power is clearly visible in the doubly selective channel, which supports our conclusion in (9).



(a) 2D plot of the channel matrix (b) 3D plot of the channel matrix

Fig. 2. 2D and 3D plot of normalized matrix coefficients $\frac{|H(k,l)|}{|H|_{max}}$ of a TU6 channel matrix $\boldsymbol{H}$ at a normalized Doppler frequency of $f_{D,norm} = 0.2$

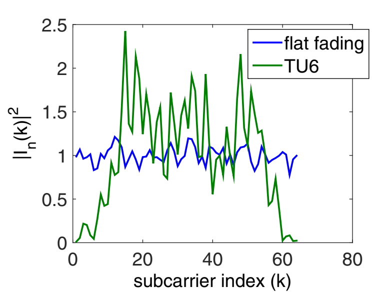


Fig. 3. normalized ICI power per frequency bin, $|I_n(k)|^2 = \frac{N}{\left(\sum_{k=0}^{N-1}|I(k)|^2\right)}|I(k)|^2$, for $N = 64$ point FFT at $f_{D,norm} = 0.2$

III. Maximum Ratio Combining in Doubly Selective Channels

With multiple antennas employed at the receiver side, (5) can be generalized to

$$\widetilde{\boldsymbol{R}}(k) = \widetilde{\boldsymbol{H}}(k,k)S(k) + \widetilde{\boldsymbol{X}}(k), \tag{10}$$

where received signal $\widetilde{\boldsymbol{R}}(k) = \begin{bmatrix} R_1(k) & \cdots & R_M(k) \end{bmatrix}^T$, main diagonal coefficients $\widetilde{\boldsymbol{H}}(k,k) = \begin{bmatrix} H_1(k,k) & \cdots & H_M(k,k) \end{bmatrix}^T$ and i.i.d. AWGN noise samples $\widetilde{\boldsymbol{X}}(k) = \begin{bmatrix} X_1(k) & \cdots & X_M(k) \end{bmatrix}^T$ are the $M \times 1$ vectors from M antennas at subcarrier position k respectively. The output of the diversity combiner is

$$Y(k) = \widetilde{\boldsymbol{W}}^H(k)\widetilde{\boldsymbol{R}}(k). \tag{11}$$

In [14], MRC was derived using an eigenfilter approach for the cases of equal and unequal noise power across the receive antennas. The coefficients for a MRC are derived as [14]

$$\widetilde{\boldsymbol{W}}(k) = \boldsymbol{C}_x^{-1}\widetilde{\boldsymbol{H}}(k,k), \tag{12}$$

where $\boldsymbol{C}_x$ is the real-valued noise covariance matrix

$$\boldsymbol{C}_x = E\left\{\widetilde{\boldsymbol{X}}(k)\widetilde{\boldsymbol{X}}^H(k)\right\} = \begin{bmatrix} \frac{1}{\sigma_1^2} & & 0 \\ & \ddots & \\ 0 & & \frac{1}{\sigma_M^2} \end{bmatrix}.$$

$$\begin{aligned} \widetilde{\boldsymbol{W}}(k) &= \begin{bmatrix} \frac{1}{\sigma_1^2} & & 0 \\ & \ddots & \\ 0 & & \frac{1}{\sigma_M^2} \end{bmatrix} \widetilde{\boldsymbol{H}}(k,k) \\ &= \begin{bmatrix} \frac{H_1(k)}{\sigma_1^2} & \cdots & \frac{H_M(k)}{\sigma_M^2} \end{bmatrix}^T \end{aligned} \tag{13}$$

In case of equal noise power, (12) can be written as $\widetilde{\boldsymbol{W}}(k) = \widetilde{\boldsymbol{H}}(k,k)$. Furthermore, normalization has to be applied on $Y(k)$ to obtain the final symbol estimate, $\hat{S}(k) = \frac{Y(k)}{\widetilde{\boldsymbol{W}}^H(k)\widetilde{\boldsymbol{H}}(k,k)}$.
In a mobile SIMO channel, (6) can be generalized by (10) to

$$\widetilde{\boldsymbol{R}}(k) = \widetilde{\boldsymbol{H}}(k,k)S(k) + \sum_{d \neq 0}\widetilde{\boldsymbol{H}}(k,(k+d)_N)S(k+d) + \widetilde{\boldsymbol{X}}(k). \tag{14}$$

In this work, we investigate the advantages of considering the frequency selectivity of the interference term $\widetilde{\boldsymbol{I}}(k) = \sum_{d \neq 0}\widetilde{\boldsymbol{H}}(k,(k+d)_N)S(k+d)$, such that we use $\boldsymbol{C}_x(k)$ rather than $\boldsymbol{C}_x$ in (12) i.e.

$$\widetilde{\boldsymbol{W}}(k) = \begin{bmatrix} \frac{H_1(k)}{\sigma_1^2(k)} & \cdots & \frac{H_M(k)}{\sigma_M^2(k)} \end{bmatrix}^T. \tag{15}$$

In the next section, we explain how this matrix can be computed for a DVB-T system. After combining, an estimate of the noise variance is required for subsequent blocks e.g. QAM demapper. The output noise variance can be computed using (12) as

$$\begin{aligned} \sigma_{out}^2(k) &= E\left\{\widetilde{\boldsymbol{W}}^H(k)\widetilde{\boldsymbol{X}}_t^H(k)\widetilde{\boldsymbol{X}}_t(k)\widetilde{\boldsymbol{W}}(k)\right\} \\ &= \widetilde{\boldsymbol{W}}^H(k)E\left\{\widetilde{\boldsymbol{X}}_t^H(k)\widetilde{\boldsymbol{X}}_t(k)\right\}\widetilde{\boldsymbol{W}}(k) \\ &= \widetilde{\boldsymbol{W}}^H(k)\boldsymbol{C}_x(k)\widetilde{\boldsymbol{W}}(k) \\ &= \widetilde{\boldsymbol{H}}^H(k,k)\left(\boldsymbol{C}_x^{-1}(k)\right)^H\boldsymbol{C}_x(k)\left(\boldsymbol{C}_x^{-1}(k)\right)^H\widetilde{\boldsymbol{H}}(k,k) \\ &= \widetilde{\boldsymbol{H}}^H(k,k)\boldsymbol{C}_x^{-1}(k)\boldsymbol{C}_x(k)\boldsymbol{C}_x^{-1}(k)\widetilde{\boldsymbol{H}}(k,k) \\ &= \widetilde{\boldsymbol{H}}^H(k,k)\boldsymbol{C}_x^{-1}(k)\widetilde{\boldsymbol{H}}(k,k) \\ &= \widetilde{\boldsymbol{H}}^H(k,k)\widetilde{\boldsymbol{W}}(k), \end{aligned} \tag{16}$$

where $\widetilde{\boldsymbol{X}}_t(k)$ is the total interference plus noise at subcarrier k, $\widetilde{\boldsymbol{X}}_t(k) = \sum_{d \neq 0}\widetilde{\boldsymbol{H}}(k,(k+d)_N)S(k+d) + \widetilde{\boldsymbol{X}}(k)$. As we see in (16), the output noise power depends on the accurate estimation of the input noise covariance matrix $\boldsymbol{C}_x(k)$.

IV. Windowed Noise Estimation for DVB-T system

Fig. 4 depicts the noise power estimated at the CoP positions in a TU6 channel at signal to noise ratio $SNR = 30dB$ and normalized Doppler shift of 0.13. Although the number of CoPs is significantly smaller than that of the scattered pilots, the former can still capture well the frequency selectivity of the colored component in the input noise (ICI) as shown in Fig. 4. It therefore makes sense to consider a window over the input CoPs such that a localized estimate of the noise power is targeted rather than an average noise power per OFDM symbol. Using a rectangular window of q input CoPs, the average input noise power at subcarrier position k is calculated as

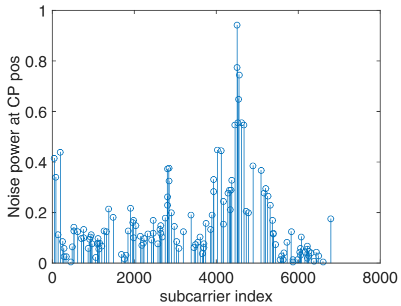


Fig. 4. Input noise power captured by CoPs in a TU6 channel, $SNR = 30dB$ and $f_{D,norm} = 0.13$.

$$\hat{\sigma}^2(k) = \frac{1}{q} \sum_{m \in \Gamma_k} \sigma^2(m) \ \ \forall k \notin \Gamma_{CoP} \tag{17}$$

where Γ_k is the window of q CoPs centered around the current subcarrier k and Γ_{CoP} is the set of all CoPs and $\sigma^2(m)$ is the noise variance computed at the CoP position at subcarrier m. In Fig. 5, we can see the MSE in noise estimation(normalized with respect to the maximum MSE, $\mu_i = \frac{mse_i}{max(mse_i)}$) for different Doppler shifts and channel lengths. We can see that, on the one hand, in a flat fading channel or in a static channel, the larger the window size the better the performance. This is intuitively satisfying since in these cases, the noise power is white. On the other hand, the higher the Doppler shift in the TU6 channel, the smaller the optimum window size is. As we can see, the optimum window size depends on the ratio of the colored noise to the total noise. The contribution of the colored noise increases as the Doppler shift increases or the frequency selectivity increases. Therefore, as we shall see in the next section, the optimum window size depends on the Doppler shift, channel delay profile and SNR used. One idea to decrease the dependency on the optimum window size is to apply a weighting function. We here propose a simple weighting criterion depending on the distance of the considered CoP from the current subcarrier position. We therefore modify (17) as

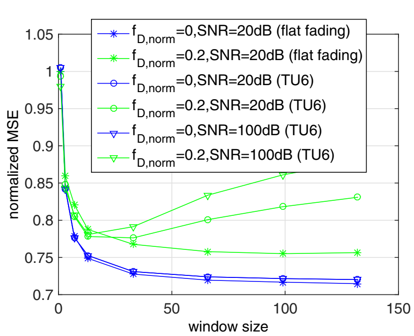


Fig. 5. Normalized MSE in noise estimation, $\mu_i = \frac{mse_i}{max(mse_i)}$, for normalized Doppler frequencies of 0 and 0.2 and 16QAM modulation for $N = 8K$ point DFT

$$\hat{\sigma}^2_w(k) = \frac{1}{\sum_{m \in \Gamma_k} v(k,m)} \sum_{m \in \Gamma_k} v(k,m)\sigma^2(m) \ \ \forall k \notin \Gamma_{CoP}, \tag{18}$$

where we choose the weighting function $v(k,m)$ as

$$v(k,m) = \frac{1}{|k-m|}. \tag{19}$$

Equation (17) is then a special case of (18) with $v(k,m) = 1$. In Fig. 6, we can see the actual input noise power (shown in blue). Using a window size of $q = 13$ CoPs (shown in red), the noise power can be better tracked than just setting it to the average noise power value $\bar{\sigma}^2 = \frac{1}{N_{CoP}} \sum_{m \in \Gamma_{CoP}} \sigma^2(m)$ (shown in black), where N_{CoP} is the number of CoPs per OFDM symbol. We can also see that even better tracking capability can be achieved using a window size of $q = 100$ samples combined with the proposed weighting criterion (shown in green).

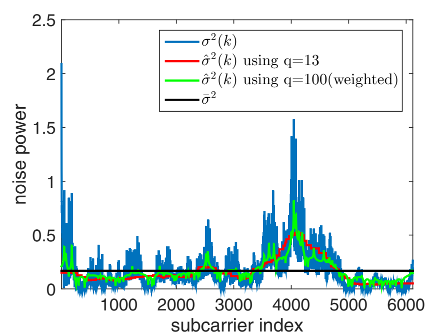
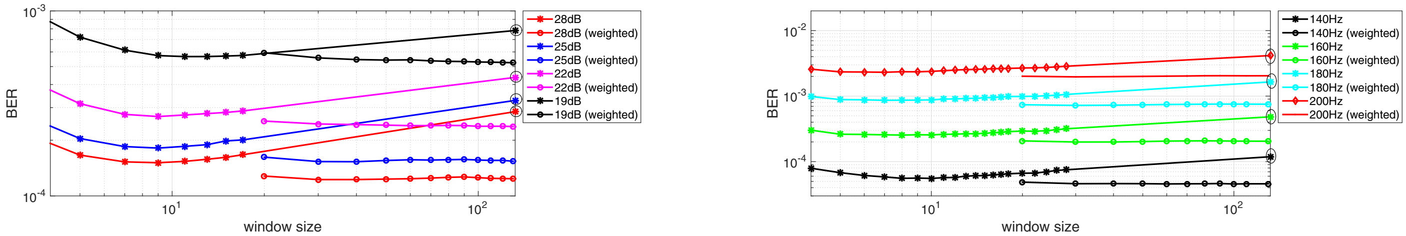


Fig. 6. Estimated noise power in a TU6 channel, 64QAM $SNR = 30dB$ and $f_{D,norm} = 0.13$.

V. Simulation Results

In this section, we present simulation results for a 2×1 SIMO DVB-T system [1], at a carrier frequency of 600MHz. An 8k FFT is used with subcarrier spacing $\Delta f = 1.116KHz$ and a guard interval of 2k samples. Out of the 8192 subcarriers, only 6048 carry data. A Viterbi convolutional channel decoder is used, with a code rate of 2/3. We consider a Doppler frequency of 150Hz, which correspond to a velocity of 270km/h. The simulated channel is a TU6 channel. Pilot assisted channel estimation is conducted to obtain the main diagonal channel coefficients, using frequency only linear interpolation.

In Fig. 7, we can see the BER performance vs the selected noise window size, at different values of SNR and Doppler shift. Compared to the conventional solution, where all CoPs



(a) $f_{D,max} = 150Hz$ (b) $SNR = 30dB$

Fig. 7. BER vs noise window size for different Doppler shift and SNR values and 16QAM modulation. (Conventional solution in circles)

are averaged to get one noise variance value per OFDM symbol (performance highlighted in circles), we can see how the noise windowing can lead to a lower error floor at $f_{D,max} = 150Hz$. However, at low SNR very small window sizes (such as $q = 5$) can lead to a worse performance as shown in Fig. 7. We can also see that the weighting proposed in (18) has significantly decreased the dependency of the BER performance on the selected noise window sizes, where the curves with the proposed weighting criterion in (18) have much flatter response than the other curves without special weighting. In addition, curves which follow the weighting criterion in (18) exhibit lower BER in comparison to their non-weighted counterparts. In DVB-T systems, quasi error-free transmission after Reed Solomon decoder is achieved when the BER after inner Viterbi convolutional decoder is below 2×10^{-4}. In Fig. 8, we can see BER performance after Viterbi for the TU6 channel as well as for the 2TU6 channel defined in [15] at $SNR = 30dB$, where the second TU6 channel is attenuated by $3dB$ and delayed by $\tau_{delay} = 0.15GI \approx 0.134msec$. As we can see, significant gain can still be achieved even for the more frequency selective 2TU6 channel by applying a weighted noise window size of $q = 100$ samples.

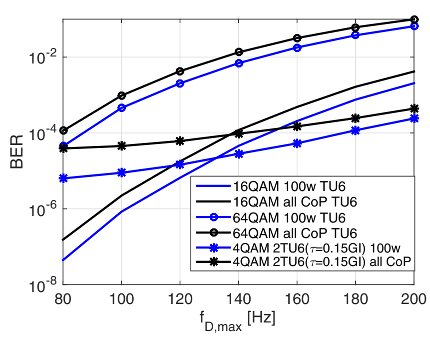


Fig. 8. BER performance after Viterbi for different channels and QAM constellations at $SNR = 30dB$

VI. Conclusion

In this work, we have proposed a low-complexity enhanced MRC scheme for mobile DVB-T systems. The proposed scheme uses the CoPs to capture the frequency selectivity of the noise information in a doubly selective channel and exhibits much lower error floor than the conventional scheme, which averages all CoPs for noise estimation. The optimum window size depends on the ratio of the colored noise component to the total noise. We have proposed a simple weighting criterion, which reduced the dependency of the BER on the optimum window size for a wide range of SNR and Doppler shift. Further results, not shown in this paper due to lack of space, show that using a weighted window size of 100 samples show a superior performance over a wide range of channel frequency selectivity. The proposed scheme can be applied for DVB-T2 systems as well as other systems which include CoPs in the transmitted symbols.

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