A SIMPLE ANTENNA COMBINING FRAMEWORK FOR DOPPLER COMPENSATION IN MOBILE OFDM SYSTEMS

Semih Serbetli

Philips Research Laboratories, Eindhoven, The Netherlands
Email: semih.serbetli@philips.com

ABSTRACT

In OFDM systems, Doppler spreading due to the mobility of the receiver distorts the orthogonality among the subcarriers, and results in intercarrier interference that degrades the performance. In this paper, we investigate how Doppler spreading can be mitigated by using multiple antennas. In this context, we propose a novel antenna combining framework exploiting the correlation among the time varying channels seen by the multiple antennas. Depending on the computational complexity requirements, the scheme can take the form of beamforming, beamforming with frequency offset correction and simple time-varying combining schemes. We derive the optimum combining scheme in each context, and show that by using the proposed combining schemes, the performance of the mobile OFDM systems can be greatly enhanced.

***Index Terms*—** OFDM, Time-varying channels, Antenna arrays

1. INTRODUCTION

Orthogonal Frequency Division Multiplexing (OFDM) is an effective way to mitigate the multipath spread of the wireless channels by a simple equalization. Due to its robustness against the multipath spread of the channel and its high spectral efficiency, it has been widely adopted in different wireless standards, e.g., DVB-T/H [1], and it is also being considered as the most valuable candidate for future cellular radio systems. Though its advantages, the reception quality of OFDM systems is hampered by the intercarrier interference (ICI) due to the Doppler spread of the signal in high mobility scenarios [2–4].

The ICI problem in mobile OFDM systems is a well known and widely studied problem, e.g., [2–6]. The proposed solutions can be grouped under three categories, i.e., signal processing based ICI cancellation schemes, self ICI cancelling coding schemes, and multiple antenna techniques. Signal processing based ICI cancellation schemes require estimation of the time varying characteristics of the channel to estimate the interference of each subcarrier to its adjacent subcarriers to cancel the ICI, e.g., [4], whereas self ICI cancelling coding schemes provide robustness to the mobility at the expense of reduced spectral efficiency, e.g., [6]. The multiple antenna approaches are based on exploiting antenna diversity and spatial processing to mitigate the ICI, e.g., [7–10].

Recently, the use of multiple antennas at both transmitter and receiver side has recently attracted a lot of attention due to its inherent capability to increase the spectral efficiency. Especially, using maximum ratio combing (MRC) at the receiver is shown to provide enhanced reception quality in time invariant channels. However, in the presence of high Doppler spread, MRC can not mitigate the ICI [7, 8]. Thus, there have been several different approaches for mitigating Doppler spread by using multiple receive antennas. The first group of solutions is based on combining the received signals from each antenna considering both the channel gain and the ICI power level for each subcarrier, e.g., [7]. The second group of solutions is based on forming virtual stationary antennas by using spatial interpolation to mitigate the Doppler spread [10]. These two approaches perform quite well, but they require complex receiver structures. The third group of solutions is based on exploiting the directivity of Doppler spread by using directional antennas [8, 9]. In this paper, we propose a novel antenna combining framework that utilizes the correlation among the receive antennas to mitigate the Doppler spread. We investigate three special cases namely, *beamforming* (BF), *beamforming with frequency offset correction* (BFOC) and *simple time-varying combining* (STVC) and demonstrate the performance of the proposed solutions.

2. SYSTEM DESCRIPTION

We consider a conventional OFDM system with M receive antennas where N complex symbols, $\mathbf{s} = [s_1, s_2, ..., s_N]$, are modulated onto N orthogonal subcarriers by using an N-point IFFT. We assume that a cyclic prefix longer than the length of the channel impulse response is added to the signal to prevent inter-block-interference. The transmitted signal goes through a wide sense stationary time varying multipath channel consisting of uncorrelated paths with complex attenuation $\{h_{i,l}(t)\}$ and delay of $\{\tau_l\}$ with an impulse response

$$h_i(t,\tau) = \sum_{l=0}^{L-1} h_{i,l}(t)\delta(\tau - \tau_l), \quad (1)$$

for the i^{th} receive antenna. Following the same notation in [4, 5], we denote the baseband received signal at the i^{th} receive antenna in time domain as $r_i(t)$ and express it as

$$r_i(t) = \sum_{n=0}^{N-1} H_{i,n}(t)e^{j2\pi n f_s t}s_n + \nu_i(t), \quad (2)$$

where $H_{i,n}(t) = \sum_l h_{i,l}(t)e^{-j2\pi n f_s \tau_l}$ is the channel frequency response of subcarrier n at time t at the i^{th} antenna, f_s is the subcarrier spacing and $\nu_i(t)$ is the AWGN with variance σ_n^2 at the i^{th} antenna. Following the approach in [4, 5], $H_{i,n}(t)$ can be approximated by using Taylor series expansion around t_0 up to the first-order term as

$$H_{i,n}(t) \approx H_{i,n}(t_0) + H'_{i,n}(t_0)(t - t_0). \quad (3)$$

Using (3), $r_i(t)$ can be approximated as [4, 5]

$$r_i(t) \approx W_i(t) + I_i(t) + \nu_i(t), \quad (4)$$

where $W_i(t) = \sum_{n=0}^{N-1} H_{i,n}(t_0)e^{j2\pi n f_s t}s_n$ is the wanted part of the signal, and $I_i(t) = \sum_{n=0}^{N-1}(t-t_0)H'_{i,n}(t_0)e^{j2\pi n f_s t}s_n$ is the ICI generating part of the signal. In the sequel, we use (4) to evaluate the performance of the proposed combining schemes.

3. ANTENNA COMBINING FRAMEWORK

We consider the mobile reception of OFDM signals with closely spaced multiple receive antennas. Each antenna observes a different time varying channel. However, there exists some correlation among these channels depending on the angle-of-arrival (AOA) distribution of the scattered signals, the antenna spacing between the antenna elements and the electromagnetic coupling among them. In this section, we exploit the correlation among both fixed and time-varying parts of the channels to find a good antenna combining scheme to compensate for the Doppler spread. We assume that the receiver has the information of correlation among the fixed parts of the channels, $E\big[H_{i,n}(t_o)H^*_{j,n}(t_o)\big]$, the time-varying parts of the channels, $E\big[H'_{i,n}(t_o){H'}^*_{j,n}(t_o)\big]$, and also the cross-correlation among them, $E\big[H_{i,n}(t_o){H'}^*_{j,n}(t_o)\big]$ where the expectation is taken over the time and subcarriers, and they form the $M \times M$ matrices, $\mathbf{R}_{HH}$, $\mathbf{R}_{H'H'}$ and $\mathbf{R}_{HH'}$, respectively. In this context, different from the conventional antenna combining schemes [7,8], we choose to do combining in time domain for each OFDM symbol with possibly time-varying combining weights of $f_1(t), f_2(t), \cdots f_M(t)$ for M antennas as in Figure 1. Thus, the combined signal can be expressed as

$$r_c(t) = \sum_{i=1}^{M} f_i(t) r_i(t). \tag{5}$$

The choice $\{f_i(t)\}$ depends on the correlation among the channels and the computational complexity requirements induced by $\{f_i(t)\}$. Note that the combining weights, $\{f_i(t)\}$, can also be approximated by using Taylor series expansion as

$$f_i(t) = f_i(t_0) + f_i'(t_0)(t-t_0) + f_i''(t_0)(t-t_0)^2 + \cdots, \tag{6}$$

which form the general combining weight vector of

$$\mathbf{f}(t) = [f_1(t), f_2(t), \cdots, f_M(t)]^T = \mathbf{f}_0 + \mathbf{f}_1(t-t_0) + \cdots. \tag{7}$$

Using (7), appropriate combining weights can be investigated up to a desired precision and complexity. In this paper, we restrict ourselves to the approximation up to the first-order term. Thus, the combined signal can be approximated as

$$r_c(t) \approx W_c(t) + I_c(t) + N_c(t) \tag{8}$$

where

$$W_c(t) = \sum_{n=0}^{N-1} \mathbf{f}_0^T [H_{1,n}(t_o), H_{2,n}(t_o), ..., H_{M,n}(t_o)]^T e^{j2\pi n f_s t} s_n,$$

$$I_c(t) = \sum_{n=0}^{N-1} (t-t_0)[\mathbf{f}_0^T [H'_{1,n}(t_o), H'_{2,n}(t_o), ..., H'_{M,n}(t_o)]^T + \mathbf{f}_1^T [H_{1,n}(t_o), H_{2,n}(t_o), ..., H_{M,n}(t_o)]^T] e^{j2\pi n f_s t} s_n,$$

$$N_c(t) = [\mathbf{f}_0^T + (t-t_0)\mathbf{f}_1^T][\nu_1(t), \nu_2(t), \cdots, \nu_M(t)]^T$$

For a given $\mathbf{f}_0$ and $\mathbf{f}_1$, the power of the wanted part and ICI generating part of the signal and noise can be expressed as

$$P_W = E\{|W_c(t)|^2\} = \mathbf{f_0}^T \mathbf{R}_{HH} \mathbf{f_0}^*, \tag{9}$$

$$P_I = E\{|I_c(t)|^2\} = \frac{1}{6f_s^2} \begin{bmatrix} \mathbf{f_0}^T & \mathbf{f_1}^T \end{bmatrix} \begin{bmatrix} \mathbf{R}_{H'H'} & \mathbf{R}^\dagger_{HH'} \\ \mathbf{R}_{HH'} & \mathbf{R}_{HH} \end{bmatrix} \begin{bmatrix} \mathbf{f_0}^* \\ \mathbf{f_1}^* \end{bmatrix}, \tag{10}$$

$$P_{noise} = \sigma_n^2 (|\mathbf{f}_o|^2 + \frac{|\mathbf{f}_1|^2}{6f_s^2}), \tag{11}$$

when $t_0 = 1/2f_s$ is used, resulting a wanted part-to-ICI generating part ratio (WIR) of $WIR = \frac{P_W}{P_I}$ and a wanted part-to-ICI generating part-and noise ratio ($WINR$) of $WINR = \frac{P_W}{P_I + P_{noise}}$. In the following subsections, we investigate how the combining weights should be chosen to compensate the Doppler spread with the required computational complexity.

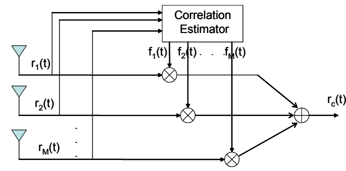

Fig. 1. General Antenna Combining Framework

3.1. Beamforming (BF)

So far, we proposed a general antenna combining mechanism to compensate the Doppler spread. When the complexity of the receiver is a concern, one may choose to combine the signals from multiple antennas with static weights, i.e., $\mathbf{f}_1 = 0$. In this case, the proposed scheme becomes similar to the case of conventional beamforming with a beamforming vector $\mathbf{f}_0$. However, the choice of the beamforming weights should be different than the conventional beamforming since the aim is to mitigate the Doppler spread rather than increasing the received signal power. In this context, one can choose WIR or $WINR$ as the performance metric which can be expressed in terms of beamforming weight, $\mathbf{f}_0$ as

$$WI(N)R = \frac{\mathbf{f_0}^T \mathbf{R}_{HH} \mathbf{f_0}^*}{\mathbf{f_0}^T \mathbf{R}_\star \mathbf{f_0}^*}, \tag{12}$$

where $\mathbf{R}_\star$ can be replaced by $\frac{1}{6f_s^2}\mathbf{R}_{H'H'}$ or $\frac{1}{6f_s^2}\mathbf{R}_{H'H'} + \sigma_n^2 \mathbf{I}$ for WIR and $WINR$ performance metrics, respectively. It can be shown that the optimum beamforming vector that maximizes WIR or $WINR$ are the generalized eigenvector with the largest eigenvalue of $\mathbf{R}_{HH}$ and $\mathbf{R}_\star$ as

$$\mathbf{f}_0^T \mathbf{R}_{HH} = \lambda_{max} \mathbf{f}_0^T \mathbf{R}_\star. \tag{13}$$

Note that the optimum beamforming vector that maximizes the $WINR$ or WIR is different than the conventional beamforming vector that aims to maximize the signal power, i.e., $\tilde{\mathbf{f}}_0^T \mathbf{R}_{HH} = \tilde{\lambda}_{max} \tilde{\mathbf{f}}_0^T$. It is also important to note that when the channel is changing very fast, i.e., the ICI terms are the dominant factor that determines the performance, the effect of the AWGN is negligible. Thus, in the high mobility scenarios, the beamforming vector that maximizes the $WINR$ converges to the beamforming vector that maximizes the WIR. Similarly, in low mobility case, the effect of ICI terms become negligible, and the beamforming vector maximizing the $WINR$ converges to the beamforming vector maximizing the received signal power.

3.2. Beamforming with Frequency Offset Correction (BFOC)

Another special form of the proposed antenna combining scheme is the combination of beamforming with frequency offset correction. In this case, the combining weights have a specific structure of $\mathbf{f}(t) = e^{-j2\pi f_\triangle t}\mathbf{w}$ where $\mathbf{w}$ is the beamforming weight and $f_\triangle$ is the frequency offset correction applied to the beamformed signal. Using Taylor series expansion up to the first order terms, $\mathbf{f}(t)$ can be approximated as

$$\mathbf{f}(t) \approx \mathbf{f}_0 + \mathbf{f}_1(t-t_0) = \mathbf{f}_0 - j2\pi f_\triangle (t-to)\mathbf{f}_0. \tag{14}$$

Observe that beamforming with frequency offset correction creates a special form of the antenna combining technique in (7) where $\mathbf{f}_1$ is a scaled version of $\mathbf{f}_0$. Using this specific structure, the WIR and $WINR$ can be expressed as in (12) with

$$\mathbf{R}_\star = \frac{1}{6f_s^2} \begin{bmatrix} \mathbf{I} & -j2\pi f_\triangle \mathbf{I} \end{bmatrix}^T \begin{bmatrix} \mathbf{R}_{H'H'} & \mathbf{R}^\dagger_{HH'} \\ \mathbf{R}_{HH'} & \mathbf{R}_{HH} \end{bmatrix} \begin{bmatrix} \mathbf{I} \\ j2\pi f_\triangle \mathbf{I} \end{bmatrix};$$

Table 1. Beamforming and Frequency Offset Correction Algorithm

Step 1: Initialize $\mathbf{f}_0$ Step 2: Optimize $f_\triangle$ for given $\mathbf{f}_0$ Step 3: Optimize $\mathbf{f}_0$ for given $f_\triangle$ Step 4: If $WI(N)R_{new} > (1+\epsilon)WI(N)R_{old}$ go to Step 2 else end.

for WIR, and

$$\mathbf{R}_\star = \frac{1}{6f_s^2}\begin{bmatrix}\mathbf{I} & -j2\pi f_\triangle \mathbf{I}\end{bmatrix}\begin{bmatrix}\mathbf{R}_{H'H'} & \mathbf{R}_{HH'}^\dagger \\ \mathbf{R}_{HH'} & \mathbf{R}_{HH}\end{bmatrix}\begin{bmatrix}\mathbf{I} \\ j2\pi f_\triangle \mathbf{I}\end{bmatrix} + \sigma_n^2\mathbf{I};$$

for $WINR$ performance metrics. Note that frequency offset correction does not affect the noise level for this approach. For a given $f_\triangle$, one can use the same approach in simple beamforming scheme to find the optimum beamforming vector as in (13). Similarly, for a given beamforming vector $\mathbf{f}_0$, it can be shown that the optimum frequency offset correction is

$$f_{\triangle,opt} = \frac{Re\{j\mathbf{f}_0^T\mathbf{R}_{HH'}\mathbf{f}_0^*\}}{2\pi\mathbf{f}_0^T\mathbf{R}_{HH}\mathbf{f}_0^*}. \quad (15)$$

Thus, one can devise an iterative algorithm for finding the jointly optimum beamforming scheme and the frequency offset correction method by optimizing each term assuming that the other parameter is fixed. However, the iterative approach is not guaranteed to converge the global optimum solution. One can choose to initialize the beamforming vector as the one that maximizes the signal power and fine-tune the parameters with the iterative algorithm. Another option is to run the iterative algorithm several times with different initial beamforming vectors and frequency offset correction terms and choose the best performing one. The outline of the proposed iterative algorithm is summarized in Table 1.

3.3. Simple Time-Varying Combining (STVC)

We investigated the proposed antenna combining scheme for two special cases and optimized the parameters for these specific structures. In this subsection, we study the antenna combining scheme without any structural constraints. In this context, we define the parameters to be optimized as

$$\mathbf{f}^T = [\mathbf{f}_0^T, \mathbf{f}_1^T]. \quad (16)$$

When $\mathbf{f}$ is used as the combining parameter vector, P_W, P_I and P_{noise} can be reformulated as

$$P_W = \mathbf{f}^T\begin{bmatrix}\mathbf{R}_{HH} & 0^\dagger \\ 0 & 0\end{bmatrix}\mathbf{f}^* = \mathbf{f}^T\mathbf{R}_W\mathbf{f}^*, \quad (17)$$

$$P_I = \frac{1}{6f_s^2}\mathbf{f}^T\begin{bmatrix}\mathbf{R}_{H'H'} & \mathbf{R}_{HH'}^\dagger \\ \mathbf{R}_{HH'} & \mathbf{R}_{HH}\end{bmatrix}\mathbf{f}^* = \mathbf{f}^T\mathbf{R}_I\mathbf{f}^*, \quad (18)$$

$$P_{noise} = \sigma_n^2\mathbf{f}^T\begin{bmatrix}\mathbf{I}_{M\times M} & 0 \\ 0 & \frac{1}{6fs^2}\mathbf{I}_{M\times M}\end{bmatrix}\mathbf{f}^* = \mathbf{f}^T\mathbf{R}_n\mathbf{f}^* (19)$$

The optimum $\mathbf{f}$ can be found by using linear algebra, and (17-19). WIR maximizing $\mathbf{f}$ is the generalized eigenvector of $\mathbf{R}_W$ and $\mathbf{R}_I$ with the largest eigenvalue, i.e., $\mathbf{f}^T\mathbf{R}_W = \beta_{max}\mathbf{f}^T\mathbf{R}_I$. Similarly, $WINR$ maximizing $\mathbf{f}$ is the generalized eigenvector of $\mathbf{R}_W$ and $\mathbf{R}_I + \mathbf{R}_n$ with the largest eigenvalue, i.e., $\mathbf{f}^T\mathbf{R}_W = \tilde{\beta}_{max}\mathbf{f}^T(\mathbf{R}_I + \mathbf{R}_n)$. Thus, one can find the optimum STVC simply by using the antenna correlations.

4. MUTUAL COUPLING EFFECT

When several antenna elements are closely spaced, as in the proposed antenna combining schemes, the electromagnetic field generated by each antenna element affects the distribution of the current and voltage on the others. Thus, the current/voltage at each antenna element does not depend only on the incident electromagnetic field on the antenna itself but also on the field of the other antenna elements. This effect is called mutual coupling effect, and it is widely studied in the context of MIMO systems, e.g., [11]. The coupling effect can be simply modelled by a multi-port network. Following the same approach as in [11], the coupled channels seen by each antenna can be expressed as

$$\begin{bmatrix}h_1(t,\tau) \\ h_2(t,\tau) \\ \vdots \\ h_M(t,\tau)\end{bmatrix} = \mathbf{C}\begin{bmatrix}h_{u,1}(t,\tau) \\ h_{u,2}(t,\tau) \\ \vdots \\ h_{u,M}(t,\tau)\end{bmatrix};\ \mathbf{C} = (Z_A + Z_L)(\mathbf{Z} + Z_L\mathbf{I})^{-1}, \quad (20)$$

where $h_{u,i}(t,\tau)$ is the channel impulse response of the i^{th} antenna in an uncoupled case, $\mathbf{C}$ is the mutual coupling matrix with Z_A, Z_L and $\mathbf{Z}$ as the antenna impedance, loading impedance and mutual impedance matrix, respectively.

Mutual coupling affects both the channels that each antenna experiences and the correlation among them. Thus, it should be considered during the proposed antenna combining scheme. If the channel correlations used to derive the combining weights are estimated in a coupled case, then there is no change required in the proposed antenna combining approach since the estimated correlations of the antennas represent the actual correlations in which all antennas are operating and coupling. However, if the correlations are estimated in an uncoupled case, i.e., each antenna take turns to receive and use this signal for correlation estimation, then, correlation matrices should be multiplied with the coupling matrices to consider the effect of coupling on the correlations, i.e., $\mathbf{R}_{coupled} = \mathbf{C}\mathbf{R}_{uncoupled}\mathbf{C}^\dagger$. Thus, the proposed antenna combining scheme can be still used for the mutually coupled antennas where coupling can be compensated automatically by the correlation estimation process or explicitly by updating the correlation matrices considering the coupling effect.

5. NUMERICAL RESULTS AND CONCLUSIONS

In this section, we present numerical results to demonstrate the Doppler compensation performance of the proposed antenna combining framework with different complexity requirements. We consider an OFDM system with a carrier spacing, $f_s = 1$ KHz, an SNR of 23 dB and a maximum Doppler frequency of $f_d = 100$ Hz. We assume that the receiver has two dipole antennas that are aligned with the direction of movement, and the antennas are loaded with 50Ω. The receiver has the perfect information of the correlation matrices, $\mathbf{R}_{HH}$, $\mathbf{R}_{H'H'}$ and $\mathbf{R}_{HH'}$ considering the effects of mutual coupling. We investigate the WIR and $WINR$ improvements obtained by using the proposed antenna combining framework with different computational complexities, i.e., BF, BFOC and STVC, for different antenna spacings in different scattering environments. For the sake of comprehensibility of the graphs, we only present the performance of the proposed schemes that maximizes the $WINR$ which is a more practical performance metric than WIR.

For simulation results, we consider three different scattering scenarios, namely, uniform AOA distribution, truncated Gaussian AOA distribution with 0 mean and a variance of $\pi/16$, and bimodal truncated Gaussian AOA distribution with means 0 and π

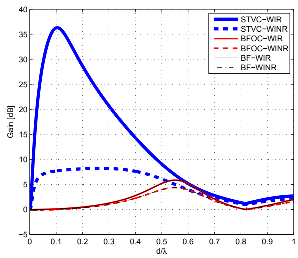

Fig. 2. Uniformly Distributed AOAs

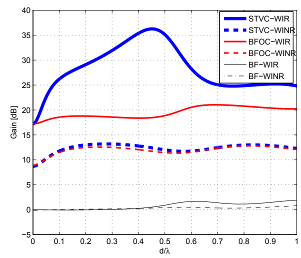

Fig. 3. Truncated Gaussian Distributed AOAs

and variances of $\pi/16$. For each scenario, we present the WIR and $WINR$ gains obtained by using two antennas with respect to single antenna receiver. In Figure 2, we present the performance in uniformly distributed AOAs. We observe that all of the antenna combining approaches improve the performance in terms of WIR and $WINR$. The performance of BF and BFOC is quite similar where both schemes provide a maximum gain of 5.8 dB in WIR and 4.4 dB in $WINR$ at the antenna spacing $d = 0.55\lambda$. However, their performance is quite sensitive to the antenna spacing and the carrier frequency. Thus, to obtain good Doppler compensation performance by using these methods, appropriate antenna design is required. It is observed that the STVC scheme cancels the ICI terms almost completely, and provides a very good Doppler compensation mechanism for closely spaced antennas where the correlation among the antennas is high. We observe that STVC is more robust to antenna spacing and carrier frequency than the BF and BFOC schemes.

Figure 3 shows the Doppler compensation performance in a scattering environment with a truncated Gaussian angle spread having a mean 0 and variance of $\pi/16$. This is a scenario where the signals are arriving the receiver with a narrow angle spread that are severely affected by Doppler spreading. We observe that BF can not improve the performance much since there is no direction that the incoming signals will lead to low ICI while providing a good received signal power. Since, BFOC has the advantage of mitigating most of the Doppler spread by a simple frequency shift and also improve the signal strength by beamforming in that direction, it cancels most of the ICI terms and provides a good performance improvement in both WIR and $WINR$. The STVC scheme provides the best performance also in this scenario. It cancels most of the ICI terms and provides a very high WIR gain. However, with 23 dB SNR and maximum Doppler frequency of 100 Hz, the performance improvement in $WINR$, is not much different than the BFOC scheme. Next, in Figure 4, we present the performance in a scattering environment with bimodal truncated Gaussian angle spread having means 0 and π and variances of $\pi/16$. As expected, the BF scheme does not provide much improvement in such a scattering environment. The BFOC scheme can still mitigate the Doppler spread by a simple frequency shift since it forms the beamforming vector that focuses on one of the mean AOAs of the bimodal distribution. The STVC scheme

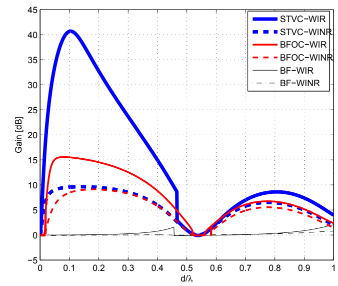

Fig. 4. Bimodal Truncated Gaussian Distributed AOAs

provides a good combining mechanism also in this scenarios. We observe that the STVC scheme is the most promising Doppler compensation mechanism. It is observed that it is less sensitive to the antenna spacing and channel models than the other simpler schemes, and can provide OFDM systems robustness against mobility.

In this paper, we proposed a simple antenna combining framework for Doppler compensation, and showed that it becomes similar to some conventional receiver mechanism in some special cases. We show that as the complexity of the combining mechanism is increased, the performance of mitigating the Doppler spread becomes better and more robust to different channel models. The framework can also be used to devise new types of combining schemes. We observe that even with limited complexity, STVC approach provides a robust and effective Doppler compensation performance. Thus, it can serve as a stand-alone Doppler compensation mechanism or can be combined with existing mechanisms to provide more robustness against Doppler spreading.

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