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ABSTRACT

Branch power equalization and increased system robustness against antenna orientation in transmit polarization diversity systems is achieved through the precoded transmission of elliptical polarization. Elliptical polarization offers two main benefits over traditional polarization diversity systems. Firstly, it is able to facilitate power coupling between orthogonal polarizations and reduce the system sensitivity to antenna orientation. Secondly, when the operation parameters for elliptical polarization are chosen carefully based on channel information, branch power equalization is achieved for arbitrary transmitter orientation. In this paper, we articulate the parameters which need to be precoded at the transmitter to achieve power balance. The closed form solutions for optimal cross polarization discrimination are presented, assuming the full knowledge of channel statistics.

Categories and Subject Descriptors
C.2.1 [Computer-Communication Networks]: Network Architecture and Design—Wireless communication

General Terms
Transmit Polarization Diversity, Cross Polar Discrimination

1. INTRODUCTION

Antenna based diversity techniques have been well documented in the literature [3]. In typical systems using spatial diversity at the base station, antenna separations in the order of 30 wavelengths is prerequisite to achieve sufficient decorrelation for diversity action. Due to the limitations in space availability, this type of diversity may be difficult and even impossible to implement. Antenna polarization diversity has been considered as a more attractive alternative because antenna elements can be co-located [5, 6]. Furthermore, an advantage of antenna polarization diversity is its ability to recover from a polarization mismatch, which occurs when the polarization of the receiver antenna becomes mis-aligned with the dominant polarization as a result of random antenna orientations [4].

Traditional polarization diversity systems have been realized with the transmission of a principally linearly polarized signal and then using a pair of antennas to capture the signal in orthogonal polarizations [1, 9]. As the signal propagates through the mobile medium and interacts with channel obstacles, some of the signal power residing in the co-polarization is coupled into the cross-polarization. Ideally, half the power in the co-polarized plane couples into the cross polarized plane, and signals propagating through the orthogonal polarizations suffer uncorrelated fading. When these conditions are not satisfied, system robustness against multipath fading deteriorates.

One of the main issues in traditional implementations of polarization diversity transmitting a principally vertically polarized signal and then using a pair of antennas to capture the signals in the vertical polarization (Vpol) and the horizontal polarization (Hpol) is the large power discrepancy between the received signals. Even within an urban environment, the density of the obstacles can not guarantee adequate power coupling [9], giving rise to a dominant polarization dictated by the orientation of the transmitter. Furthermore, due to the nature of electromagnetic propagation, attenuation of the orthogonal polarizations are not symmetrical, with the polarization perpendicular to the channel obstacles suffering greater attenuation than the polarization parallel to the obstacles [4]. Despite adequate power coupling, the asymmetric attenuation between polarizations will still give rise to branch power imbalance.

The power imbalance between the Vpol and Hpol is quantified by the Cross Polarization Discrimination (XPD), and it is defined as the ratio between the total power available in the Vpol and the total power in the Hpol. When there is large discrimination, diversity performance is sub-optimal because signal contributions from the branch with lower power is limited.

One method to equalize the power discrepancy was introduced by Kozono et al. in [5]. This features a pair of linearly polarized antennas, which have been aligned at an angle ±α relative to the vertical axis. However, this technique of power equalization is at the expense of increased fading correlation. Fading correlation could be controlled by changing α to allocate different amounts of signal contribution from the Vpol and Hpol, but this causes the mean signal levels to drop as a result of polarization mismatch. Furthermore, this configuration assumes the transmission of a principally vertically polarized signal, which may not al-
always be the case in the mobile uplink.

To encourage power coupling between polarizations and compensate for the asymmetric attenuation at the same time, the transmission of an elliptically polarized signal is considered. The use of elliptical polarization is an advantage because firstly, the rotation of polarization states is artificially introduced at the transmitter to encourage power coupling. Secondly, assuming that accurate channel estimates are readily available at the transmitter, the signal could be preconditioned to inject more power into the polarization expected to suffer higher attenuation.

In this paper, we show that by adaptively precoding the parameters of elliptical polarization at the transmitter, assuming accurate estimations of the channel statistics, branch power equalization is obtained. The optimal operation parameters are orientation specific, therefore a knowledge of the transmitter orientation must be available during the adaptive precoding phase prior to transmission. The availability of the transmitter orientation in the uplink will not be a problem, as the emerging mobile handsets available on the market include inbuilt accelerometers suitable for this application.

The structure of the paper is as follows. In Section 2 we detail the transmit and receiver systems, and explain the generation of an elliptically polarized signal. The channel characteristics are also presented. The expressions of the diversity parameters XPD and power correlation as a function of antenna orientation are derived in Section 3. The optimal parameters and the method to achieve branch power equalization given arbitrary orientation is presented in Section 4. Section 5 presents a discussion on the theoretical and simulated results. Finally, Section 6 concludes the paper.

2. SYSTEM MODEL

2.1 Transmitter System

At the transmitter, a pair of linearly polarized antennas $V_1$ and $V_2$, which are always orthogonal to each other is considered. The transmitter orientation is described by $\alpha$, defined as the angle between antenna $V_1$ and the $y$ axis. Azimuthal dependence is not considered and the configuration is assumed to be broadside to the receiver. This is illustrated in Fig. 1(a).

An elliptically polarized signal is produced when the signals feeding into the two antennas take the following form:

$$V_1 = A_1 e^{j(\omega t - \phi)}$$
$$V_2 = A_2 e^{j(\omega t - \theta)},$$

where $A_1$ and $A_2$ are the signal amplitudes of $V_1$ and $V_2$ respectively. $\theta$ controls the phase offset between the two antennas, and in particular, when $\theta = 0$, it corresponds to linear polarization as both antennas are transmitting inphase.

Despite the random orientation at the transmitter, the analysis is conducted with respect to the Vpol and Hpol. A transformation matrix is used to standardize the field components back into the Vpol and Hpol, written here as $V_y$ and $V_z$ respectively.

$$
\begin{bmatrix}
V_y \\
V_z
\end{bmatrix} =
\begin{bmatrix}
\sin(\alpha) & -\cos(\alpha) \\
\cos(\alpha) & \sin(\alpha)
\end{bmatrix}
\begin{bmatrix}
V_1 \\
V_2
\end{bmatrix}
\quad (3)
$$

2.2 Channel

The modelling of the channel takes a similar approach to that of [6], and describes the channel between the transmitter and receiver with four communications links. This is illustrated in Fig. 2.

Each of the communication links has been given a complex channel response in the form $\Gamma e^{j\omega t}$, where $\Gamma$ represents the short term fading as a result of the multipath effect while $\phi$ represents the random phase offset introduced by the channel. To remain consistent with the channel measurements reported in [6, 7], weak fading correlation is assumed between $\Gamma_{11}$ and $\Gamma_{21}$ as well as $\Gamma_{22}$ and $\Gamma_{12}$. All other combinations of $\Gamma$ are assumed independent. The random phase offsets are all modelled as independent random variables uniformly distributed between $[0, 2\pi)$. The communication medium is also assumed to be linear and passive, so that reciprocity can be applied [6].

After the signal has propagated through the channel, the resultant signal present inside the Vpol and the Hpol at the receiver, written here as $R_y$ and $R_z$ is calculated by

$$
\begin{bmatrix}
R_y \\
R_z
\end{bmatrix} =
\begin{bmatrix}
\Gamma_{11} e^{j\phi_{11}} & \Gamma_{12} e^{j\phi_{12}} \\
\Gamma_{21} e^{j\phi_{21}} & \Gamma_{22} e^{j\phi_{22}}
\end{bmatrix}
\begin{bmatrix}
V_z \\
V_y
\end{bmatrix}.
$$
\quad (4)

2.3 Receiver System

The receiver configuration is based on the one considered in [9]. Two linearly polarized diversity antennas $R_1$ and $R_2$, broadside to the transmitter, which are always orthogonal but rotatable together are considered. The rotation of the receiver is modelled by $\psi$ and is defined as the angle be-
between antenna $R_1$ and the vertical axis. This configuration is illustrated in Fig. 1(b).

The signals received in the diversity antennas are obtained by summing the projections from $R_x$ and $R_y$.

$$\begin{bmatrix} R_1 \\ R_2 \end{bmatrix} = \begin{bmatrix} \sin(\psi) & \cos(\psi) \\ -\cos(\psi) & \sin(\psi) \end{bmatrix} \begin{bmatrix} R_x \\ R_y \end{bmatrix}.$$  

(5)

3. DIVERSITY PARAMETERS

The performance of diversity systems is described by the diversity gain, and is defined as the improvement in the signal to noise ratio with diversity relative to a single branch without diversity combining, at a given level of cumulative probability. For polarization diversity, diversity gain is influenced by the XPD and the fading correlation between antenna branches.

3.1 Cross Polarization Discrimination

The cross polarization discrimination (XPD) is the measure of the difference in signal power between the Vpol and Hpol. It is defined as the ratio between the total power available at the receiver in the Vpol and the total power available in the Hpol. Firstly, to calculate their respective powers in the Vpol and Hpol, we start with equation (4) and proceed in the Hpol. Firstly, to calculate their respective powers in the Vpol and the total power available between antenna branches.

$$XPD = \frac{E[R_2^2]}{E[R_2^2]} = \frac{K_d E[\Gamma_{21}^2] + K_s E[\Gamma_{22}^2]}{K_d E[\Gamma_{11}^2] + K_s E[\Gamma_{12}^2]}.$$  

(6)

where $K_d$ and $K_s$ are

$$K_d = A_1^2 \sin^2(\alpha) + A_2^2 \cos^2(\alpha)$$

$$K_s = A_1^2 \cos^2(\alpha) + A_2^2 \sin^2(\alpha)$$

(7)

When the operating conditions $A_1 = 1$, $A_2 = 0$ and $\alpha = 0$ is considered, the expression of the XPD simplifies down to

$$XPD_{Vpol} = \frac{E[\Gamma_{22}^2]}{E[\Gamma_{12}^2]}.$$  

(9)

which represents the degree of power coupling from the Vpol to the Hpol. This is the traditional definition of the XPD in the literature [5, 9] assuming the transmission of a vertically polarized signal. A similar interpretation for the power coupling form the Hpol into the Vpol is obtained by setting $A_1 = 1$, $A_2 = 0$ and $\alpha = \pi/2$.

$$XPD_{Hpol} = \frac{E[\Gamma_{21}^2]}{E[\Gamma_{11}^2]}.$$  

(10)

There is a high level of discrimination between the Vpol and the Hpol, the power contributions from $E[\Gamma_{11}^2]$ and $E[\Gamma_{21}^2]$ becomes insignificant. An expression of the relative attenuation between the Vpol to Vpol and Hpol to Hpol branches is obtained when $A_1 = 1$, $A_2 = 0$ and $\alpha = \pi/2$.

$$XPD_{Aym} = \frac{E[\Gamma_{22}^2]}{E[\Gamma_{11}^2]}.$$  

(11)

The asymmetric attenuation across the Vpol and Hpol has been measured to fluctuate between 0dB and 15dB, with the local means from the polarizations to be within $\pm$3dB for almost 90% of the time [6].

3.2 Power Correlation Coefficient

The power correlation coefficient between receiver antennas $R_1$ and $R_2$ is an indication of the degree of similarity between fading signatures, and is calculated by

$$\rho_p = \frac{E[R_1^2 R_2^2] - E[R_1^2] E[R_2^2]}{\sqrt{E[R_1^4] - E^2[R_1^2]} \sqrt{E[R_2^4] - E^2[R_2^2].}$$  

(12)

The required moments and joint moments of $R_1^2$ and $R_2^2$ are obtained from (5) and substituted into (12). The calculation assumes correlated channels as described in Section 2.2. Further, it is assumed that the envelope fading follows Rayleigh distribution, as it is considered to be the worst case scenario for short term fading [8]. As a result, it follows that all of the fourth order moments of $\Gamma$ can be rewritten as second order moments through the property $E[\Gamma^4] = 2E[\Gamma^2]$.

The simplified expression of the power correlation is

$$\rho_p = \left\{ \begin{array}{c} \tan^2(\psi) \left[ E[R_2^2] - E[R_1^2] \right]^2 + 2\beta_{ep} \tan^2(\psi) \right\}$$

$$\sqrt{\left(\tan^2(\psi) E[R_2^2] + E[R_1^2]\right)^2 + 4\beta_{ep} \tan^2(\psi).}$$  

(13)

where

$$\beta_{ep} = K_2^2 \text{Cov}(\Gamma_{21}, \Gamma_{21}) + K_2^2 \text{Cov}(\Gamma_{22}, \Gamma_{22}).$$  

(14)

and $\text{Cov}(X, Y)$ is the covariance between $X$ and $Y$. For analysis purposes, it is often more convenient to express the channel covariance in terms of the channel correlation coefficients $\rho_{11,21}$ and $\rho_{12,22}$.

$$\beta_{ep} = K_2^2 \rho_{11,21} E[\Gamma_{11}^2] E[\Gamma_{21}^2] + K_2^2 \rho_{12,22} E[\Gamma_{12}^2] E[\Gamma_{22}^2].$$  

(15)

4. OPTIMAL PARAMETERS

Branch power equalization is achieved from the transmitter for a known arbitrary transmitter orientation, given the perfect knowledge of channel conditions. The parameters $A_1$, $A_2$ and $\theta$ are precoded to give the ideal eccentricity and rotation of the ellipse, so that power imbalance in a particular polarization may be compensated. The optimization procedure is to achieve two conditions. Firstly, the value of the XPD=0dB at the given transmitter orientation $\alpha$,

$$XPD = \frac{E[R_2^2]}{E[R_2^2]} = 1,$$  

(16)

secondly, the resulting XPD as a function of $\alpha$ must be at a global minimum for the defined $\alpha$

$$\frac{d \text{XPD}}{d \alpha} = 0.$$  

(17)

To find the values of $A_1$, $A_2$ and $\theta$ for optimum XPD, these two conditions are used to form two equations. A solution is first derived in terms of $T$, which represents the ratio between the antenna amplitudes $T = A_1/A_2$, and $\theta$. A third constraint is then introduced to specify the total transmitted power to finally obtain values for $A_1$ and $A_2$.

From (6) and (16), when represented in terms of $T$, it
follows that optimal XPD=0dB is achieved when
\[ G = \frac{T^2 \sin^2(\alpha) + \cos^2(\alpha) - 2T \sin(\alpha) \cos(\alpha) \cos(\theta)}{T^2 \cos^2(\alpha) + \sin^2(\alpha) + 2T \sin(\alpha) \cos(\alpha) \cos(\theta)}, \] (18)

where
\[ G = \frac{E \left[ \Gamma_{22}^2 \right] - E \left[ \Gamma_{12}^2 \right]}{E \left[ \Gamma_{11}^2 \right] - E \left[ \Gamma_{21}^2 \right]}. \] (19)

Rewriting (18) to make \( \cos(\theta) \) the subject, we get
\[ \cos(\theta) = \frac{\sin(\alpha) \{ T^2 - G \} + \cos^2(\alpha) (1 - T^2) G}{2T \sin(\alpha) \cos(\alpha) (G + 1)}, \] (20)

for \( \alpha \neq \frac{\pi}{2} \) where \( n \) is integer.

When the second condition is used to derive the second equation, \( \cos(\theta) \) and \( \sin(\theta) \) are obtained.

Finally, by combining (20) and (22) the expression for \( T \) is obtained.
\[ T^2 = \frac{G \sin^2(\alpha) + \cos^2(\alpha)}{\sin^2(\alpha) + G \cos^2(\alpha)} \] (23)

The general algorithm to compute the optimal transmission parameters is to first obtain knowledge on the transmitter orientation \( \alpha \) and the channel conditions for calculating \( G \). The ratio of \( A_1 \) to \( A_2 \) is then obtained through \( T \) in (23). Following this, depending on the value of \( \alpha \), the phase offset \( \theta \) is calculated from (20) and (22). Finally, a constraint is placed on the total transmit power and \( A_1 \) and \( A_2 \) are then calculated. The parameters of the elliptical polarization should be computed periodically, determined by the channel coherence time, to ensure that the XPD is adaptively optimized in a time varying channel.

An exhaustive approach was employed to verify the parameters for optimal XPD over a range of different \( \alpha \). An urban environment with characteristics consistent with those reported in the literature [1, 6, 9] was assumed, and this translates to the conditions where \( \text{XPD}_{\text{Hpol}} = \text{XPD}_{\text{Vpol}} = 6\text{dB} \) and \( \text{XPD}_{\text{Asym}} = 3\text{dB} \). Table 1 summarizes the results.

The analytical solutions for the optimal parameters were compared with the simulated results. The comparisons are illustrated in Fig. 3 and Fig. 4. In both figures, it can be seen that the simulated parameters closely agree with the analytical results.

5. PERFORMANCE ANALYSIS

In this section, we demonstrate the ability of elliptical polarization to achieve optimal XPD, given arbitrary transmitter orientation. An urban environment was assumed in the analysis, where \( \text{XPD}_{\text{Vpol}} = \text{XPD}_{\text{Hpol}} = 6\text{dB} \) and \( \text{XPD}_{\text{Asym}} = 3\text{dB} \). The channels are also assumed to be correlated, such that \( \rho_{11,21} = \rho_{12,22} = 0.4 \). Following the analysis of the XPD, the power correlation coefficient across antenna \( R_1 \) and \( R_2 \) are also presented. Computer simulations were used to verify validity of the development, and the generation of correlated Rayleigh channels was achieved using the method described in [2]. Fifty thousand independent trials were conducted and all of the simulated results agree to the theoretical calculations.

5.1 Cross Polarization Discrimination

Fig. 5 shows the results for the elliptical polarization systems which have been tuned for optimal performance at \( \alpha \).
The result verifies our design, and demonstrates that an optimal XPD, the maximum correlation coefficient will remain suitably low with a maximum value no larger than 0.15.

6. CONCLUSION

Branch power equalization is achieved through the transmission of a precoded elliptical polarization assuming the perfect knowledge of the channel statistics and transmitter orientation. For arbitrary transmitter orientations, the operation parameters of elliptical polarization can be precoded for optimal XPD. The branch power correlation is also shown to be very low, suitable for diversity action. Dynamic adaptation of the operation parameters using the proposed general algorithm could guarantee effective and improved diversity action irrespective of transmitter orientation, which may not be possible with the traditional space or polarization diversity schemes.

7. REFERENCES


