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Polarization Distortion as a Means for Securing Wireless Communication

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Abstract—This paper presents a simple polarization encoding strategy that operates using only single element dual port transmit and receive antennas in such a way that selective spatial scrambling of QPSK data can be effected. The key transmitter and receiver relationships needed for this operation to occur are derived. The system is validated using a cross dipole antenna arrangement. Unlike all previously reported physical layer wireless solutions the approach developed in this paper transfers the security property to the receive side resulting in very simple transmit and receive side architectures thus avoiding the need for near field modulated array technology. In addition the scheme permits, for the first time, multiple spatially separated secured receive sites to operate in parallel.

Index Terms—Constellation pattern, directional modulation, polarization distortion.

I. INTRODUCTION

Recently the Directional Modulation (DM) scheme has received attention as a potential means for physical layer wireless security. DM is a technique that enables, through the use of near field encoding, digital signal mappings in IQ space that are dependent on spatial direction. Radio systems with this added capability can reduce probability of intercept of a radio transmission beyond that achievable for a conventional wireless broadcast.

Attempts to date [1]-[6] encode wanted digital data directly into the individual scatterers comprising the DM transmission array. In all cases the design process is very complicated due to the complex interactions in the near field and their spatial dependent transformation into the far field. In [7] quadrature modulated I and Q data streams were separately baseband encoded then up-converted to radio frequency (RF), transmitted, and then combined in the far-field so that IQ data de-modulation could occur along a pre-defined direction. Unfortunately the approach does not lend itself to applications where the prescribed direction needs to vary, e.g., in a mobile application scenario.

In the DM architectures proposed to date wave polarization has not been considered as a means for spatially selective physical layer data scrambling. By introducing polarization pre-distortion at the transmitter and appropriate post-distortion correction at the receiver we will show in this paper that a robust DM system with very simple architecture can be constructed based on a single two port antenna at the transmit side and a single two port antenna at the receive side. Furthermore the system operates in such a way that spatial

direction selectivity is determined by the receiver positioned in the far-field and not by the transmitter as has been the case in all previously reported art. As a consequence the receiver position relative to the transmitter for optimal reception does not need to be known, or even specified, a-priori. In addition the proposed scheme makes it possible for the first time for simultaneous multi-direction secure wireless communications in a straightforward manner without recourse to any modification to the transmitter settings.

In Section II of this paper, the architecture of the proposed system is described. In Section III an investigation, for QPSK modulation, of the polarization relationships between a transmit and receive antenna pair is given. This is followed by an example design for a single antenna QPSK DM system. In Section IV a modified version of error vector magnitude (EVM) is used in order to assess the performance of the system. In Section V the proposed architecture is experimentally verified for a cross dipole pair. Finally conclusions are drawn in Section VI.

II. SYSTEM ARCHITECTURE WITH POLARIZATION DISTORTION CAPABILITY

The proposed system architecture is illustrated in Fig. 1. Dual-port cross dipole are chosen as transmit and receive antennas due to their excellent port circuit isolation and capability for generating orthogonally polarized electromagnetic waves in the x - z plane with low EM radiation cross polarization in the far-field.

For the dual-port cross dipole transmit antenna the $\hat{\theta}$ and $\hat{\phi}$ components of the far-field \vec{E} in the x - z plane, denoted as \vec{E}_θ and \vec{E}_ϕ , are determined by excitations at port 1 and port 2 respectively. Each port is modulated by separate I and Q data streams. An additional phase shift of β is added to the Q data path prior to transmission. At the receiver the signal at port 1 is combined with the signal at port 2, which itself is phase delayed by η . After suitable receiver configuration at the required spatial direction, these combined data streams appear at the receiver output as well formatted IQ data that can be recovered by a standard quadrature demodulator.

The mathematical derivation necessary to reveal the interrelationships among the transmitter and receiver settings for spatially scrambled QPSK modulation in free space is presented in the next section.

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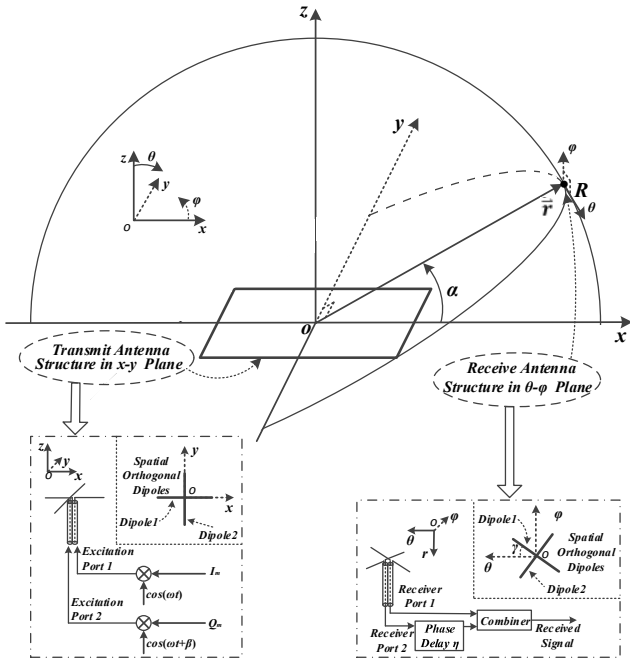


Fig. 1. System architecture and co-ordinate system.

III. MATHEMATICAL DERIVATION AND DESIGN PROCEDURE FOR CROSS DIPOLE QPSK DM SYSTEM

In order to model the system in Fig. 1 we assume:

- That the receive antenna locates in x - z plane in the far-field, and that the receive cross dipole structure lies in the θ - ϕ plane and can be rotated to angle γ relative to the positive θ axis, Fig. 1.
- I_m and Q_m are the m^{th} input digital IQ data streams. Under the QPSK modulation scheme we assign states ‘-1’ or ‘+1’ according to the symbol transmitted. For example, symbol ‘01’ indicates that ‘ $I_m = -1$ ’ and ‘ $Q_m = +1$ ’.
- As described in Section II and illustrated in Fig. 1, β is the LO phase shift applied when up-converting the Q data stream at the transmitter relative to the phase of the I data stream;
 η is the phase delay inserted at receive antenna port 2;
 γ is the rotation angle of the dual port cross dipole receive antenna relative to the positive θ axis in θ - ϕ plane.
- $F_\theta(\alpha)$ and $F_\phi(\alpha)$ are the far-field magnitudes of \vec{E}_θ and \vec{E}_ϕ in x - z plane measured at the receiver location, and α is the spatial projection angle in this cut.
- \vec{k} is the wavenumber vector along the spatial transmission direction.
- \vec{r} represents the location vector of the receive antenna relative to the transmit antenna.

With the definitions above, while transmitting the m^{th} data symbol, \vec{E} at the far-field point \mathbf{R} in θ - ϕ coordinate, denoted as \vec{E}_{Rm} , can be expressed as (1)

$$\vec{E}_{Rm} = \underbrace{-I_m \cdot F_\theta(\alpha) \cdot \cos(\omega t - \vec{k} \cdot \vec{r}) \cdot \hat{\theta}}_{\vec{E}_{\theta Rm}} + \underbrace{-Q_m \cdot F_\phi(\alpha) \cdot \cos(\omega t - \vec{k} \cdot \vec{r} + \beta) \cdot \hat{\phi}}_{\vec{E}_{\phi Rm}} \quad (1)$$

where $\hat{\theta}$ and $\hat{\phi}$ are the unit vectors along the corresponding directions. With reference to Fig. 1 the signs in the front of terms $\vec{E}_{\theta Rm}$ and $\vec{E}_{\phi Rm}$ in (1) are, respectively, a result of the inner conductor to dipole arm connection direction of each dipole in the receive cross dipole antenna. When the directions are opposite, these signs need to be changed. This is included in the mathematical derivation below.

Applying (1) to the arrangement in Fig. 1 wherein the receive antenna at the spatial location \mathbf{R} is illuminated by an incident wave arriving from the broadside direction, the received signal S_m at the output of the receive side combiner can, after inclusion of the phase delay η at receiver port 2, be expressed as in (2). S_{mij} ($i, j = 1, 2$) occurs as a consequence of the m^{th} symbol detected at the receiver port i due to the \vec{E} field generated by the excitation at transmitter port j . For instance, as can be seen in Fig. 2, signals S_{m11} and S_{m21_p} , which is the signal before the phase delay at receiver port 2, at the receiver ports 1 and 2 respectively are the projections of $\vec{E}_{\theta Rm}$ onto the corresponding port vectors \vec{V}_{port1} and \vec{V}_{port2} in θ - ϕ plane, i.e., ‘ $\vec{E}_{\theta Rm} \cdot \cos(\gamma)$ ’ and ‘ $\vec{E}_{\theta Rm} \cdot \cos(\pi/2 + \gamma)$ ’.

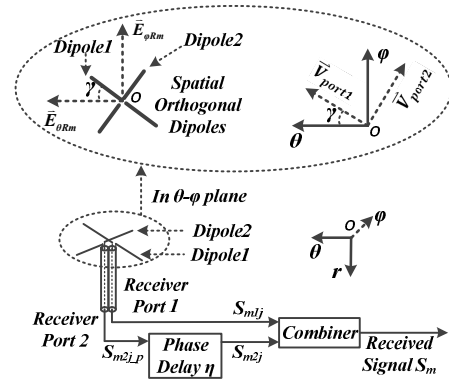


Fig. 2. The receive cross dipole antenna showing the projections of $\vec{E}_{\theta Rm}$ and $\vec{E}_{\phi Rm}$ onto port vectors of \vec{V}_{port1} and \vec{V}_{port2} .

$$S_m = \underbrace{-I_m \cdot F_\theta(\alpha) \cdot \cos(\omega t - \vec{k} \cdot \vec{r}) \cdot \cos(\gamma)}_{S_{m11}} + \underbrace{-Q_m \cdot F_\phi(\alpha) \cdot \cos(\omega t - \vec{k} \cdot \vec{r} + \beta) \cdot \cos\left(\frac{\pi}{2} - \gamma\right)}_{S_{m12}} + \underbrace{-I_m \cdot F_\theta(\alpha) \cdot \cos(\omega t - \vec{k} \cdot \vec{r} - \eta) \cdot \cos\left(\frac{\pi}{2} + \gamma\right)}_{S_{m21}} + \underbrace{-Q_m \cdot F_\phi(\alpha) \cdot \cos(\omega t - \vec{k} \cdot \vec{r} + \beta - \eta) \cdot \cos(\gamma)}_{S_{m22}} \quad (2)$$

Equation (2) can be rewritten as (3), wherein A_m and ξ_m are the amplitude and phase of the received signal. These correspond to the constellation point of the m^{th} symbol in IQ space. The spatial path phase delay $\vec{k} \cdot \vec{r}$ represents a constellation phase rotation, and is set to zero since it has no influence on constellation pattern shape.

$$S_m = \underbrace{\begin{Bmatrix} -I_m \cdot F_\theta(\alpha) \cdot \cos(\gamma) \\ -Q_m \cdot F_\phi(\alpha) \cdot \cos(\beta) \cdot \sin(\gamma) \\ +I_m \cdot F_\theta(\alpha) \cdot \cos(-\eta) \cdot \sin(\gamma) \\ -Q_m \cdot F_\phi(\alpha) \cdot \cos(\beta - \eta) \cdot \cos(\gamma) \end{Bmatrix}}_{g_m} \cdot \cos(\omega t) - \underbrace{\begin{Bmatrix} -Q_m \cdot F_\phi(\alpha) \cdot \sin(\beta) \cdot \sin(\gamma) \\ +I_m \cdot F_\theta(\alpha) \cdot \sin(-\eta) \cdot \sin(\gamma) \\ -Q_m \cdot F_\phi(\alpha) \cdot \sin(\beta - \eta) \cdot \cos(\gamma) \end{Bmatrix}}_{h_m} \cdot \sin(\omega t) \quad (3)$$

$$= A_m \cdot \cos(\omega t + \xi_m)$$

where $A_m = \sqrt{g_m^2 + h_m^2}$, and $\tan(\xi_m) = \frac{h_m}{g_m}$

In this paper, QPSK modulation with Gray coding is adopted. Thus the four symbols '11', '01', '00' and '10' in a standard QPSK system should lie in the first to the fourth quadrants respectively. In the system under consideration here, by substituting I_m and Q_m in (3) with '-1' or '+1' according to the transmitted symbol 'pq', where $(pq) \in \{(11), (01), (00), (10)\}$, the four unique received QPSK symbols, denoted as $S^{(pq)}$ with magnitude $A^{(pq)}$ and phase $\xi^{(pq)}$, can be obtained. $S^{(pq)}$, $A^{(pq)}$ and $\xi^{(pq)}$ are functions of α , β , γ and η .

$$\left(\frac{A^{11}}{A^{01}} \right)^2 = \frac{\begin{Bmatrix} F_\theta^2(\alpha) + F_\phi^2(\alpha) \\ + [F_\phi^2(\alpha) - F_\theta^2(\alpha)] \cos(\eta) \sin(2\gamma) \\ + 2F_\theta(\alpha) F_\phi(\alpha) [\sin(\beta) \sin(\eta) + \cos(\beta) \cos(\eta) \cos(2\gamma)] \end{Bmatrix}}{\begin{Bmatrix} F_\theta^2(\alpha) + F_\phi^2(\alpha) \\ + [F_\phi^2(\alpha) - F_\theta^2(\alpha)] \cos(\eta) \sin(2\gamma) \\ - 2F_\theta(\alpha) F_\phi(\alpha) [\sin(\beta) \sin(\eta) + \cos(\beta) \cos(\eta) \cos(2\gamma)] \end{Bmatrix}} \quad (4)$$

$$\tan(\xi^{01} - \xi^{11}) = \frac{\tan(\xi^{01}) - \tan(\xi^{11})}{1 + \tan(\xi^{01}) \tan(\xi^{11})} = \frac{2F_\theta(\alpha) F_\phi(\alpha) [\sin(\beta) \cos(\eta) \cos(2\gamma) - \cos(\beta) \sin(\eta)]}{[F_\phi^2(\alpha) - F_\theta^2(\alpha)] + [F_\phi^2(\alpha) + F_\theta^2(\alpha)] \cos(\eta) \sin(2\gamma)} \quad (5)$$

Since $A^{11} = A^{00}$, $A^{01} = A^{10}$, $\xi^{11} = \xi^{00} + \pi$ and $\xi^{01} = \xi^{10} + \pi$, then symbols '11' and '00' are always central-symmetric in IQ space, ditto symbols '01' and '10'. From the above observation we see that two degrees of freedom for receive side constellation pattern re-construction exist, i.e., (A^{11}/A^{00}) and

$(\xi^{01} - \xi^{11})$. In order to achieve standard QPSK symbol patterns in IQ space only along spatial direction α , these quantities have to be 1 and $\pi/2$ respectively. For this to occur equations (4) and (5) derived from (3) are bounded by the three requirements cited in (6), which need to be simultaneously satisfied.

$$\begin{cases} \cos(2\gamma) \cos(\beta) \cos(\eta) = -\sin(\beta) \sin(\eta) \\ \cos(\eta) \sin(2\gamma) = \frac{F_\theta^2(\alpha) - F_\phi^2(\alpha)}{F_\theta^2(\alpha) + F_\phi^2(\alpha)} \\ \sin(\beta) \cos(\eta) \cos(2\gamma) > \cos(\beta) \sin(\eta) \end{cases} \quad (6)$$

The solutions of (6) can be sorted into two cases:

- When $\cos(\eta) = 0$,
$$\begin{cases} F_\theta(\alpha) = F_\phi(\alpha) \\ \beta = 0 \text{ and } \eta = -\frac{\pi}{2}; \text{ or } \beta = \pi \text{ and } \eta = \frac{\pi}{2} \end{cases} \quad (7)$$

γ can take any values, and linear polarization occurs while transmitting each symbol.

Equation (7) indicates that the communication direction is fixed to the angle where $|\vec{E}_\theta|$ and $|\vec{E}_\phi|$ are identical, i.e., the broadside direction of the dual port cross dipole transmit antenna.

- When $\cos(\eta) \neq 0$, equations (8-10) can be derived from (6),

$$\begin{cases} \sin(\eta) = -\frac{2F_\theta(\alpha) F_\phi(\alpha)}{F_\theta^2(\alpha) + F_\phi^2(\alpha)} \cos(\beta) \\ \sin(2\gamma) = \frac{\text{sign}[\cos(\eta)] \cdot [F_\theta^2(\alpha) - F_\phi^2(\alpha)]}{\sqrt{F_\theta^4(\alpha) + F_\phi^4(\alpha) - 2F_\theta^2(\alpha) F_\phi^2(\alpha) \cos(2\beta)}} \\ \text{sign}[\cos(2\gamma)] = \text{sign}[\cos(\eta) \cdot \sin(\beta)] \end{cases} \quad (8) \quad (9) \quad (10)$$

here the function $\text{sign}[x]$ returns '+1' if the corresponding value of x is greater than zero, otherwise it is '-1'.

Based on (8)-(10) an operating procedure for the proposed system is now described:

- Choose an arbitrary β at the transmit side;
- Sample $F_\theta(\alpha)$ and $F_\phi(\alpha)$ at the receive side by using a linear polarized receive antenna positioned along direction(s) α ;
- At the receiver(s) calculate η by equation (8) and set this at the receiver(s): β having been communicated by normal cryptographic means;
- Calculate receiver(s) γ using equations (9) and (10).

It is acknowledged that there is a chance for eavesdroppers to intercept the β , and hence to design appropriate receive antennas if the transmitter architecture and modulation scheme are known. However, compared with classical encryptions at the application layer, an extra hardware modification of the receive antenna at the physical layer is required. This is usually difficult to apply at the eavesdropper sides. The incorrect

receive structure will lead to severely distorted signal, and thus hamper the useful information decoding.

IV. METRIC FOR ASSESSING SYSTEM PERFORMANCE

Next the transmitter's capability for scrambling the constellation diagrams at directions other than the desired communication direction needs to be quantified. To do this we base our discussion on the *EVM* definition [8], [9], modified to reflect the properties of the system under discussion. Here we call the modified *EVM* form, $FOM(\alpha)$, and we use it as a figure of merit for describing the distortion of the received constellation diagram:

$$FOM(\alpha) = \left[\frac{\frac{1}{M} \sum_{m=1}^M |S_{DM_m} - S_{st_m}|^2}{\frac{1}{M} \sum_{m=1}^M |S_{st_m}|^2} \right]^{\frac{1}{2}} \quad (11)$$

Here S_{DM_m} is the received noiseless symbol when the m^{th} unique symbol, i.e., $m = 1, 2, 3, 4$ for QPSK scheme, is transmitted in our proposed system, whereas the S_{st_m} is the corresponding m^{th} standard received noiseless symbol. M is the total number of symbol states, i.e. 4 for QPSK case. At each specified spatial direction, the total power of four symbols received from the transmitter is normalized to be identical to that received from a standard QPSK system, namely

$$\sum_{m=1}^4 |S_{DM_m}|^2 = \sum_{m=1}^4 |S_{st_m}|^2 \quad (12)$$

Furthermore, as would happen during the standard QPSK synchronization process, each received constellation pattern is rotated to align the phase of the reference symbol. Hence $FOM(\alpha)$, describes the average normalized distance from the received symbols relative to those of the standard constellation points along a prescribed direction α .

With the above definition an $FOM(\alpha)$ value of zero means that all received symbols lie exactly on top of the ideal QPSK constellation patterns while higher values indicate that the received symbols are more grossly distorted in amplitude and/or phase relative to ideal QPSK constellation symbol position. Thus $FOM(\alpha)$ is used to quantify symbol displacement in IQ space due to spatially dependent scrambling. It should be noted that in a conventional system *EVM* is used to quantify symbol displacement along a fixed direction accounting for effects due to noise and system non-linearities. Whereas, unlike *EVM*, due to the extra capability of the constellation pattern manipulation possessed by the proposed system described here, $FOM(\alpha)$ does not link directly to system BER e.g., via the relationship stated in [9].

V. EXPERIMENT RESULTS AND DISCUSSIONS

To validate the proposed system Pawsey balun fed dipoles operating at 1 GHz were orthogonally arranged, Fig. 3, to give a two port antenna. The reflection coefficient of each port and port 1 to 2 cross coupling properties are measured and depicted in Fig. 4.

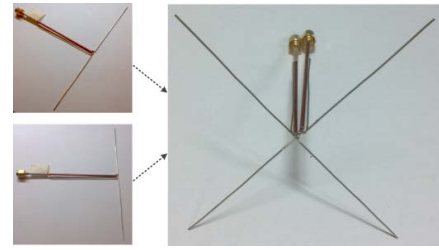


Fig. 3. Fabricated two-port cross-dipole for 1 GHz operation.

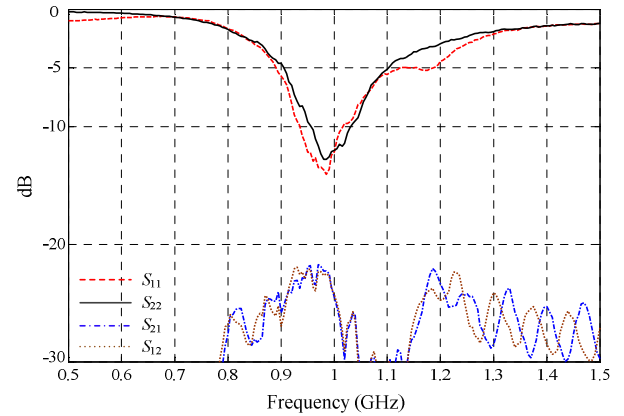


Fig. 4. Measured S-parameters for cross dipole antenna.

Fig. 5 shows the normalized far-field power patterns of the dual-port cross dipole measured in an anechoic chamber. It can be seen that \bar{E}_θ and \bar{E}_ϕ are dominant at the corresponding ports of excitation. This, together with the isolation between two ports shown in Fig 4, means that I and Q data streams can be independently modulated onto the $\hat{\theta}$ and $\hat{\phi}$ components of electric field.

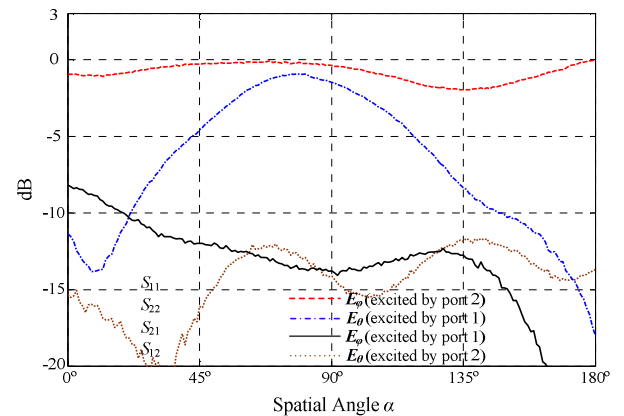


Fig. 5. Measured far-field power patterns for \bar{E}_θ and \bar{E}_ϕ in x - z plane as excited by each cross dipole port.

Following the operating procedure summarized at the end of the Section III, the settings of the cross dipole DM systems for secure wireless communication reception at an example angular elevation direction of 45° are listed in Table I. Cases 1 and 2 are for systems with different phase LO settings β imposed in the Q path at the transmit side, which leads to the different receiver parameters η and γ .

TABLE I. EXAMPLES OF CROSS DIPOLE DM SYSTEM SETTINGS FOR 45° SECURED WIRELESS COMMUNICATION

		$\alpha = 45^\circ$			
		$F_\theta(\alpha) = 0.613$		$F_\phi(\alpha) = 1.001$	
Cross Dipole DM system	Case 1	$\beta = 30^\circ$			
		$\eta = 309^\circ$		$\eta = 231^\circ$	
	$\gamma = 157^\circ$	$\gamma = 337^\circ$	$\gamma = 67^\circ$	$\gamma = 247^\circ$	
	Case 2	$\alpha = 45^\circ$			
$F_\theta(\alpha) = 0.613$		$F_\phi(\alpha) = 1.001$			
$\eta = 334^\circ$		$\eta = 206^\circ$			
$\gamma = 165^\circ$		$\gamma = 345^\circ$	$\gamma = 75^\circ$	$\gamma = 255^\circ$	

Taking Case 1 as an example, and assuming a noiseless channel, if an eavesdropper is equipped with a linearly polarized (LP), or, with a left-hand, or, right-hand circular polarized (L/RHCP) antenna, the $FOM(\alpha)$ curves obtained by using (11) are shown in Fig. 6. The different η and γ pairs in Case 1 lead to the same results. The relatively high values of the $FOM(\alpha)$ across the x - z plane, always > 0.19 , indicate that in all spatial directions only distorted QPSK constellation diagrams are present. In Fig. 6 it should be noted that the value of unity obtained for the LP receiver antennas aligned for polarization direction $\hat{\theta}$ and $\hat{\phi}$ reveals, as expected, that a pair of QPSK symbols is not being recovered in the received data streams.

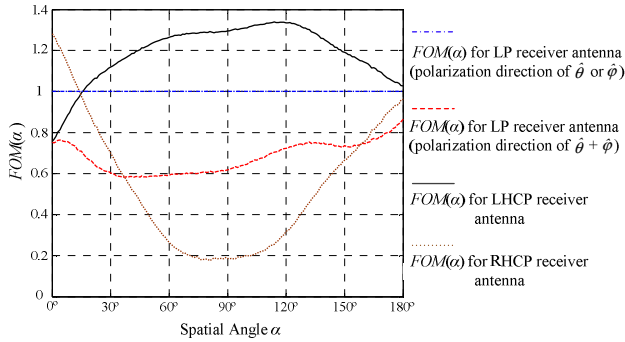


Fig. 6. The $FOM(\alpha)$ for the cross dipole transmitter DM system with settings in Case 1 Table I with LP or CP receiver antennas.

Fig. 7 shows that once the receiver is correctly imprinted, see Table I, for 45° in this example, significantly increased probability for modulation recovery exists along the prescribed 45° direction. The secondary dip around 135° is due to the symmetric property of the far-field patterns. However non-flat far-field phase patterns prevents $FOM(\alpha)$ from reaching zero. The received constellation diagrams along several spatial directions, including the desired direction, 45° in this example, are also illustrated in Fig. 7. It can be observed that the standard QPSK constellation pattern is preserved only along the 45° direction, while patterns are distorted in all other directions. In order to better contextualize the results we now show by way of example the BER performance of the system when the signal to noise ratio (SNR) over the Gaussian channel is set to 6 dB, 9 dB, then 12 dB. The resulting BER distributions are given in Fig. 7. The slight shift in the BER minimum positions towards boresight (90°) is due to the rapid roll-off of the \vec{E}_θ excited by port 1 of the cross dipole transmit antenna, see Fig. 5.

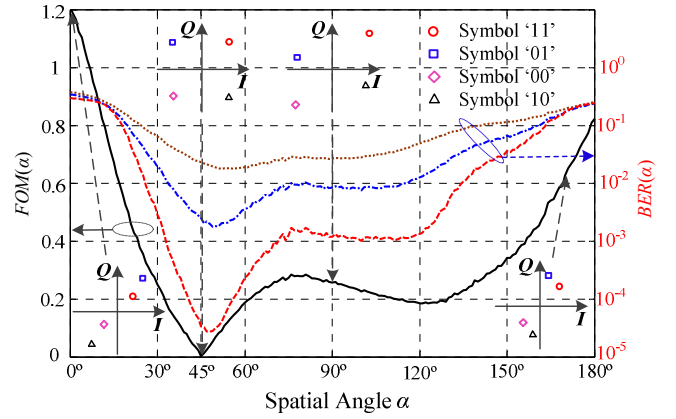


Fig. 7. $FOM(\alpha)$ for dual-port cross dipole equipped receivers imprinted for 45° secure communication using the settings for Case 1 in Table I, and its BER performance, $BER(\alpha)$, for SNR of 6 dB, 9 dB, and 12 dB respectively ('.....': for 6 dB SNR; '-.-.-': for 9 dB SNR; '-.-.-': for 12 dB SNR).

VI. CONCLUSION

In this paper a strategy was presented which allows suitably imprinted receivers of simple construction to recover elliptically polarized encoded information modulated on a per symbol basis whilst ensuring that conventional receivers of identical sensitivity at the same spatial locations would be unable to decode the same signal. The technique developed in this paper should be useful in systems where additional physical layer data security is required in a wireless system.

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