# A DOWNLINK ADAPTIVE TRANSMITTING ANTENNA METHOD FOR T/F/SDMA FDD SYSTEMS AVOIDING DOA ESTIMATION

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## ABSTRACT

In this paper, the problem of space only downlink processing for FDD ( Frequency Division Duplex ) radiocommunication systems is considered. It is shown that using criterions based on average C/I, one can avoid DOA estimation and deduce these criterions at downlink frequency from their equivalent at uplink frequency, under some assumptions about array topology. Moreover, here, we introduce a more generic approach than the former one, denoted as pattern synthesis.

## 1 INTRODUCTION

Exploiting spatial diversity with an antenna array at the base station seems to be a good way to enhance the capacity of current mobile communication systems. This can be achieved with SDMA ( Space Diversity Multiple Access ) schemes or, in a more simple manner, by reducing CCI ( Co Channel Interference ) and the size of cellular patterns. However, concerning space only downlink processing, a set of weights has to be choosen and applied to the concerned user, optimizing a criterion that involves the channels linking the base station to the desired and undesired sources.

Moreover, for FDD systems, these channels can differ significantly between up and downlink, making it difficult to exploit the available knowledge of the uplink channel for downlink beamforming. Only one basic assumption seems valid : the geometry of up and downlink channels are the same in terms of paths, DOAs ( Directions of Arrival ), and average powers.

To take advantage of this assumption, it was first suggested to detect and estimate the powers and DOAs of the paths, but this approach encounters some limitations. First, this has to be done in presence of jammers, localizing only one desired user [1]. Second, it has to be adapted if angular spreading or distributed sources are present [2] [3]. Third, it can lead to computationally exhaustive solutions. A new approach appeared with criterions based on mean powers (ie second order statistics) of the propagation channels, first optimized with the help of feedback [4]. In a recent paper [5] we showed that under certain assumptions and with given arrray topologies, frequency transposition was able to deduce these second order statistics at downlink frequency from their equivalent at uplink frequency. We shall here show that this transposition is a peculiar case of a more generic approach, which can be denoted as pattern synthesis.

## 2 SECOND ORDER STATISTICS OF THE PROPAGATION CHANNEL

### 2.1 Channel Modelisation

Consider a given user indexed by $k$. Its multidimensional uplink channel impulse response can be stored in an array $\mathbf{H}_k$ of dimension $M \times N$, where $M$ denotes the number of sensors and $N$ the maximum delay of multipath propagation for all users.

Let $\Theta_k(t)$ be the vector of the DOAs of the $P_k$ paths that link the base station to that user, which can be time-varying, and $\mathbf{D}(\Theta_k(t), f)$ the $(M \times P_k)$ steering matrix corresponding to those DOAs at frequency f. The double varying nature of the channel can then be explicited by the model :

$$\mathbf{H}_k(f,t) = \mathbf{D}(\Theta_k(t), f)\mathbf{B}_k(f,t) \tag{1}$$

where the $P_k \times N$ matrix $\mathbf{B}_k(f,t)$ summarizes all the phase shifts and attenuations encountered by the signal along the propagation paths, as well as an additional attenuation due to ISI introduced by the modulation into the channel. These coefficients appear to be highly dependent on time and frequency.

### 2.2 Temporal and Frequential Evolution of the Channel

Let us assume that the DOAs $\Theta_\mathbf{k}$ remain the same over the processing interval, a hundred of bursts for the GSM system, assumption grounded by experimental results [6]. Of course this assumption cannot be valid for the matrix $\mathbf{B}_k(f,t)$, as soon as the time considered exceeds one burst length, thus changing the fading state. But one can suppose that :

- $\mathbf{B}_k(f,t)$ is constant over one burst.
- the phase of the elements of $\mathbf{B}_k(f,t)$ are uniformly distributed over $[0, 2\pi]$.

- the different paths fade independently, and the average power of the paths is the same at both frequencies.

This results in :

$$\mathbf{E}[\mathbf{B}_k(f,t)\mathbf{B}_k^*(f,t)] = \mathbf{R}_k \quad (2)$$

### 2.3 Average Power Expression

The average power transmitted on the downlink to the desired user $k$ when applying a set of weights $\mathbf{w}_k$ can be written, supposing transmitted symbols are uncorrelated and with unitary power :

$$\begin{aligned} P^d_{k\to k} &= \mathbf{E}[\mathbf{w}_k^*\mathbf{H}_k(f_d,t)\mathbf{H}_k^*(f_d,t)\mathbf{w}_k] \\ &= \mathbf{w}_k^*\mathbf{D}(\Theta_k,f_d)\mathbf{E}[\mathbf{B}_k(f_d,t)\mathbf{B}_k^*(f_d,t)]\mathbf{D}^*(\Theta_k,f_d)\mathbf{w}_k \\ &= \mathbf{w}_k^*\mathbf{D}(\Theta_k,f_d)\mathbf{R}_k\mathbf{D}^*(\Theta_k,f_d)\mathbf{w}_k \\ &= \mathbf{w}_k^*\mathbf{\Gamma}_k^d\mathbf{w}_k \end{aligned} \quad (3)$$

where $\mathbf{\Gamma}_k^d = \mathbf{D}(\Theta_k,f_d)\mathbf{R}_k\mathbf{D}^*(\Theta_k,f_d)$.

In the same time, the average transmitted interfering power to users $j \neq k$ can be written :

$$P^d_{k\to j} = \sum_{j\neq k} \mathbf{w}_k^*\mathbf{\Gamma}_j^d\mathbf{w}_k = \mathbf{w}_k^*\left(\sum_{j\neq k}\mathbf{\Gamma}_j^d\right)\mathbf{w}_k \quad (4)$$

One can suggest to optimize the ratio between these two quantities, for each user, and thus to decide separately the sets of weights to apply by choosing :

$$\mathbf{w}_k^{opt} = arg\left(\begin{array}{c} max \\ \mathbf{w}_k \end{array} \frac{\mathbf{w}_k^*\mathbf{\Gamma}_k^d\mathbf{w}_k}{\mathbf{w}_k^*\mathbf{\Gamma}_{bb}^d\mathbf{w}_k}\right) \quad (5)$$

where $\mathbf{\Gamma}_{bb}^d = \sum_{j\neq k}\mathbf{\Gamma}_j^d$ denotes the covariance matrix of jammers related to user $k$ on the downlink. However, it is still necessary to have a good estimate of $\mathbf{\Gamma}_k^d$ and $\mathbf{\Gamma}_{bb}^d$. As was suggested in [4], the use of feedback may solve the problem. In [5], we introduced another way avoiding feedback : frequency transposition.

## 3 FREQUENCY TRANSPOSITION [5]

The first goal of frequency transposition was to deduce downlink channel second order statistics $\mathbf{\Gamma}_k^d$ and $\mathbf{\Gamma}_{bb}^d$ from its available corresponding values at uplink $\mathbf{\Gamma}_k^u$ and $\mathbf{\Gamma}_{bb}^u$.

It can be stated as follows : given two array manifolds $\mathbf{A}_u$ and $\mathbf{A}_d$ corresponding to $f_u$ and $f_d$, which dimensions are $M \times P$ (supposing the array manifold is sampled every $\frac{360}{P}$ degrees), suppose the existence of an $M \times M$ mathematical operator $\mathbf{T}_{u\to d}$ verifying ($(.)^*$ denotes transconjugate) :

$$\mathbf{T}_{u\to d}^*\mathbf{A}_u = \mathbf{A}_d \quad (6)$$

That is, $\mathbf{T}_{u\to d}$ turns any steering vector at $f_u$ into its equivalent at $f_d$.

Considering equations (3) and (4), one can then see that we have :

$$\begin{aligned} &\mathbf{T}_{u\to d}^*\mathbf{\Gamma}_k^u\mathbf{T}_{u\to d} \\ &= \mathbf{T}_{u\to d}^*\mathbf{D}(\Theta_k,f_u)\mathbf{R}_k\mathbf{D}^*(\Theta_k,f_u)\mathbf{T}_{u\to d} \\ &= \mathbf{D}(\Theta_k,f_d)\mathbf{R}_k\mathbf{D}^*(\Theta_k,f_d) = \mathbf{\Gamma}_k^d \end{aligned} \quad (7)$$

and $\mathbf{T}_{u\to d}^*\mathbf{\Gamma}_{bb}^u\mathbf{T}_{u\to d} = \mathbf{\Gamma}_{bb}^d$.

The existence of an operator $\mathbf{T}_{u\to d}^*$ with good performance in a Least Squares (LS) sense was stated for a circular array with $M = 10$ equispaced sensors as soon as its radius $R$ verifies $R/min(\lambda_u,\lambda_d) < 0.7$. Although, for $f_u = 900$ MHz and $f_d = 945$ MHz, $\mathbf{A}_u$ and $\mathbf{A}_d$ seem to be close, frequency transposition yields better results than approximating $\mathbf{A}_d$ with $\mathbf{A}_u$ [7] in terms of C/I.

## 4 PATTERN SYNTHESIS

### 4.1 Problem Statement

Let us consider another approach. Suppose first $f_u = f_d$, ie there is no FDD, and we still want to do a space only downlink processing based on average C/I. The direct availability of an estimate of $\mathbf{\Gamma}_k^u$ and $\mathbf{\Gamma}_{bb}^u$, injected in a criterion such as the one expressed in (5), would lead to a certain weight vector $\mathbf{w}_k^u$. Once applyied , this set of weights would give a certain pattern, that can be written :

$$P_u(\theta) = \frac{\mathbf{w}_k^{u^*}\mathbf{d}(\theta,f_u)\mathbf{d}^*(\theta,f_u)\mathbf{w}_k^u}{M}, \quad \forall\theta \quad (8)$$

Now, considering $f_u \neq f_d$, this function $P_u(\theta)$ that was optimized at $f_u$ can be considered as an objective function to achieve at frequency $f_d$. That means, we want to find a set of weights $\mathbf{w}_k^d$ minimizing the difference $e(\theta)$ between the two patterns :

$$e(\theta) = \frac{\mathbf{w}_k^{d^*}\mathbf{d}(\theta,f_d)\mathbf{d}^*(\theta,f_d)\mathbf{w}_k^d}{M} - P_u(\theta), \quad \forall\theta \quad (9)$$

### 4.2 Problem Solution

#### *4.2.1 Linear Solution*

Minimizing the error induced in equation (9) in a LS sense leads to a quadratic form with no trivial solution, which can be expressed :

$$w_k^d = arg\left(\begin{array}{c} min \\ \mathbf{w} \end{array} \sum_\theta \left|\frac{|\mathbf{w}^*\mathbf{d}(\theta,f_d)|^2}{M} - P_u(\theta)\right|^2\right) \quad (10)$$

Using an iterative Gauss-Newton procedure and separating $\mathbf{w} = [w_1 \ldots w_M]^T$ into its real and imaginary parts by writing :

$$\underline{\mathbf{w}} = [\Re(w_1)\ldots\Re(w_M) \quad \Im(w_2)\ldots\Im(w_M)]^T$$

where we have constrained $\Im(w_1) = 0$ because **w** is defined up to a phasis factor, the iterative procedure between step $l$ and $l+1$ can be written :

$$\underline{\mathbf{w}}^{(l+1)} = \underline{\mathbf{w}}^{(l)} + \mu\Delta\underline{\mathbf{w}} \qquad (11)$$

where $\Delta\underline{\mathbf{w}}$ is the solution in a Least Squares sense of :

$$\mathbf{F}^{(l)}\Delta\underline{\mathbf{w}} = \begin{pmatrix} P_u(\theta_1) - |\mathbf{w}^{(l)^*}\mathbf{d}(\theta_1, f_d)|^2/M \\ \vdots \\ P_u(\theta_P) - |\mathbf{w}^{(l)^*}\mathbf{d}(\theta_P, f_d)|^2/M \end{pmatrix} \qquad (12)$$

where $\mathbf{F}^{(l)}$ is given by :

$$\mathbf{F}^{(l)} = 2\left[\Re(\mathbf{G}^{(l)}) \quad -\Im(\mathbf{G}^{(l)})\right] \qquad (13)$$

with $\mathbf{G}^{(l)}$ given by :

$$\mathbf{G}^{(l)} = \begin{bmatrix} (\mathbf{w}^{(l)^*}\mathbf{d}(\theta_1))\mathbf{d}^*(\theta_1) \\ \cdots \\ (\mathbf{w}^{(l)^*}\mathbf{d}(\theta_P))\mathbf{d}^*(\theta_P) \end{bmatrix} \qquad (14)$$

In equation (13), the first column of the block matrix $\Im(\mathbf{G})$ is removed, which means that **F** has dimension $P \times (2M-1)$. The step $\mu$ is adapted in an iterative manner such that $\underline{\mathbf{w}}^{(l+1)}$ yields better performance than $\underline{\mathbf{w}}^{(l)}$.

#### *4.2.2 Logarithmic Solution*

In order to get deeper nulls in the beam, which are of drastic importance for jammers rejection, one can think of optimizing the logarithmic constraint given by :

$$w_k^d = arg\left(\begin{matrix} min \\ \mathbf{w}\end{matrix} \sum_\theta \left|10 * log10\left(\frac{|\mathbf{w}^*\mathbf{d}(\theta, f_d)|^2}{MP_u(\theta)}\right)\right|^2\right) \qquad (15)$$

This can be optimized in the same way as the linear constrained except that equations (12) and (14) have to be rewritten :

$$\mathbf{F}^{(l)}\Delta\underline{\mathbf{w}} = \begin{pmatrix} 10 * log10\left(\frac{P_u(\theta_1)M}{|\mathbf{w}^{(l)^*}\mathbf{d}(\theta_1, f_d)|^2}\right) \\ \cdots \\ 10 * log10\left(\frac{P_u(\theta_P)M}{|\mathbf{w}^{(l)^*}\mathbf{d}(\theta_P, f_d)|^2}\right) \end{pmatrix} \qquad (16)$$

$$\mathbf{G}^{(l)} = \begin{bmatrix} \frac{(\mathbf{w}^{(l)^*}\mathbf{d}(\theta_1))\mathbf{d}^*(\theta_1)}{|\mathbf{w}^{(l)^*}\mathbf{d}(\theta_1, f_d)|^2} \\ \cdots \\ \frac{(\mathbf{w}^{(l)^*}\mathbf{d}(\theta_P))\mathbf{d}^*(\theta_P)}{|\mathbf{w}^{(l)^*}\mathbf{d}(\theta_P, f_d)|^2} \end{bmatrix} \qquad (17)$$

#### *4.2.3 Initialisation*

It seems that the use of $\mathbf{w}_k^u$ as initial estimate is a good choice.

### 4.3 Suboptimal Solution

Considering equation (9), one can see that if we define a complex function $p_u(\theta)$ as follows :

$$p_u(\theta) = \mathbf{w}_k^{u^*}\mathbf{d}(\theta, f_u), \qquad \forall\theta \qquad (18)$$

we have $|p_u(\theta)|^2 = MP_u(\theta)$. Then we could try to find a vector $\mathbf{w}_k^d$ minimizing the difference :

$$\epsilon(\theta) = \mathbf{w}_k^{d^*}\mathbf{d}(\theta, f_d) - p_u(\theta), \qquad \forall\theta \qquad (19)$$

It is indeed obvious that (19) can be optimized in a much more simple manner than (9), but that the solution suffers more constraints than the former one, because it has constraints both in magnitude and phase. Thus we call it a suboptimal solution.

Moreover, considering frequency transposition, if we choose $\mathbf{w}_k^d = \mathbf{T}_{u\to d}^{-1}\mathbf{w}_k^u$, we get :

$$\mathbf{w}_k^{d^*}\mathbf{d}(\theta, f_d) = \mathbf{w}_k^{u^*}\mathbf{T}_{u\to d}^{-*}\mathbf{d}(\theta, f_d) = \mathbf{w}_k^{u^*}\mathbf{d}(\theta, f_u) = p_u(\theta) \qquad (20)$$

This shows clearly that frequency transposition can also be considered as a suboptimal way to do pattern synthesis.

## 5 Results and Conclusion

All simulations are conducted with four mobiles propagating through typical BU ( Bad Urban ) channels, each of them being constituted of one main and one secondary path distant of 45° ( see table 1 ). Each path has an angular spreading of 5° and is subject to rayleigh fading. The four mobiles impinge on the array with equal average power and the criterion (5) is optimized for each user with theoretical values of $\mathbf{\Gamma}_k^u$ and $\mathbf{\Gamma}_{bb}^u$, under the power constraint $\mathbf{w}_k^{u^*}\mathbf{\Gamma}_k^u\mathbf{w}_k^u = 1$. Simulations are going to be done first using an antenna array allowing frequency transposition, then we are going to prove that the pattern synthesis approach permits to override induced limitations of frequency transposition on array geometry.

### 5.1 Constrained Array

The constrained array is made of $M = 10$ sensors equispaced in a circular manner, with radius $R$ verifying $R/min(\lambda_u, \lambda_d) = 0.7$. Let us consider the case $f_u = 900$ MHz and $f_d = 945$ MHz, that corresponds to GSM. Figure 1 shows the pattern to achieve $P_u(\theta)$ for mobile 1 and the ones obtained with and without frequency transposition . The patterns obtained with patterns synthesis are not plotted there. Table 1 compares respectively, for each mobile, the theoretical (T) achievable C/I (dB) at $f_u$, the ones obtained at $f_d$ without frequency transposition (U), with frequency transposition (FT), with linear pattern synthesis (PS) and with logarithmic pattern synthesis (LPR).

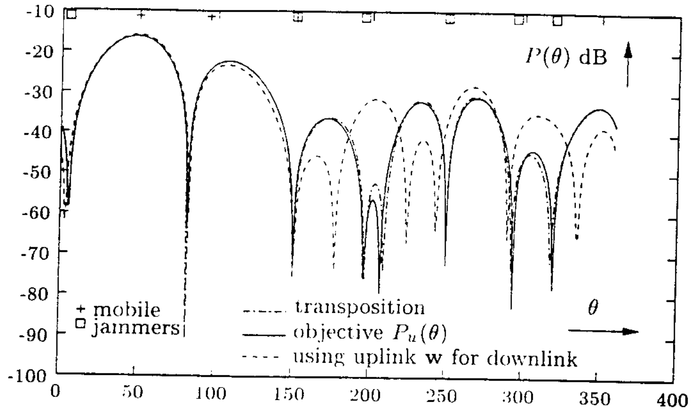


Figure 1: Patterns for mobile 1 at $f_u$ and $f_d$, $R/\lambda = 0.7$

| mobile | 1 | 2 | 3 | 4 |
|---|---|---|---|---|
| DOAs | 50 – 95 | 150 – 195 | 250 – 295 | 320 – 5 |
| C/I T | 30 | 32 | 26 | 30 |
| C/I U | 14 | 17 | 16 | 13 |
| C/I FT | 29 | 31 | 25 | 29 |
| C/I PS | 24 | 25 | 24 | 24 |
| C/I LPS | 29 | 31 | 26 | 30 |

Table 1: C/I with different approaches. $R/\lambda = 0.7$

### 5.2 Extended Array

Let us now increase the antenna radius $R$ ( and thus enhance its directivity ) such that $R/min(\lambda_u, \lambda_d) = 1$, and do the same simulations in that case. Figure 2 shows the pattern to achieve and the patterns obtained with the FT and LPS approaches. The pattern obtained using uplink weights for downlink processing are not plotted here. However, as shown in table 2, this technique does not yield good results, such as the FT approach which validity domain has been overriden. But one can see that the LPS approach, that has better respect to the nulls of the pattern, yields the best results. The PS approach seems not adapted to jammer rejection.

| mobile | 1 | 2 | 3 | 4 |
|---|---|---|---|---|
| DOAs | 50 – 95 | 150 – 195 | 250 – 295 | 320 – 5 |
| C/I T | 27 | 26 | 23 | 26 |
| C/I U | 11 | 12 | 10 | 13 |
| C/I FT | 19 | 15 | 16 | 16 |
| C/I PS | 12 | 12 | 13 | 14 |
| C/I LPS | 22 | 22 | 15 | 18 |

Table 2: C/I with different approaches, $R/\lambda = 1$

### 5.3 Conclusion

The optimal logarithmic solution enables to override the induced limitations of frequency transposition in terms of antenna size, but for a higher computational cost : a multidimensionnal criterion has to be optimized for every pattern realisation, while frequency transposition just involves one matrix multiplication.

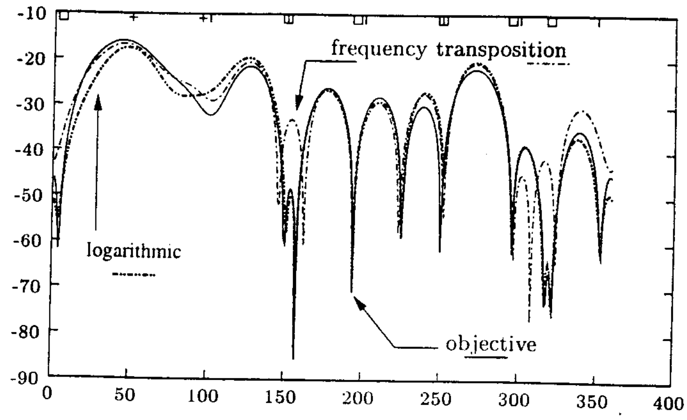


Figure 2: Patterns for mobile 1 at $f_u$ and $f_d$, $R/\lambda = 1$

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