CELLULAR INTERNET SERVICE: ADVANCED FRONT-END FOR FDSS*

Ana I. Perez-Neira, Miguel A. Lagunas

TSC Dept., Modulo D5,Campus Nord UPC
C/ Gran Capita s/n
08034 Barcelona, SPAIN

Abstract

Technologies for achieving reliable high-speed transmission to wide-area mobile and portable cellular subscribers with high spectrum efficiency are more and more necessary. One example is a basic service as Advanced Cellular Internet Service; such a wireless service has to be optimized to meet the needs of a client-server model for information retrieval and Web browsing. One possibility in the radio link design is the combination of multicarrier schemes, both OFDM and FDSS (Frequency Diversity Spread Spectrum) with antenna diversity to overcome the link budget and dispersive fading limitations of the cellular mobile radio environment. This work proposes an advanced front-end to obtain blind spatial processing for FDSS in order to enhance optimum detection in presence of band-limited partial-time interference. The procedure is adaptive and, therefore, suitable to be used in highly time varying scenarios. The algorithm is derived based on the fact of the presence of the desired signal in every component, together with an automatic gain control constraint. The optimum detector for FDSS is described in order to show its full compatibility with spatial processing techniques.

1. Introduction.

Interest in wireless Internet access was initially generated by the opportunities for e-mail and messaging, but the recent explosion in popularity of the World Wide Web suggests broader long-term opportunities for wireless data. One alternative [1] is an asymmetric wireless packet data service for macro-cellular environment, with peak down-link bit rates of 1-2Mbps and channelization of about 1 MHz. Thus, service can be deployed with a limited amount of spectrum. The up-link works with a lower speed of 50-100 Kbps. In order to minimize the effects of multipath delay spread, multicarrier techniques are used. Since, no equalization is required these techniques alleviate the need of long training periods. In the up-link OFDM is used. To ensure flat frequency response and achieve the desired bit rate, 100-200 sub-channels are required, each modulated at 5-10 Kbauds. With 5-10 Kbauds sub-channels and guard periods of 20-40 μs, delay spreads as large as 40 μs can be accommodated with little or no ISI. The up-link, or terminal-to-base transmission direction, could also use OFDM by dividing the wideband channel into clusters of 10-20 tones for use by individual terminals to achieve bit rates of about 100 kbps while requiring acceptable transmit power levels. In order to alleviate the severe problem of the sensitivity to transmission non-linearities at the mobile terminal and also of partial band interferences in the up-link, an alternative to OFDM is the use of FDSS. That is, multiple narrow carriers, which must then be demodulated using a filter bank of receivers with disjoint frequency support.

FDSS was reported by G.Kaleh [2-3] as an effective alternative that outperforms the traditional Spread Spectrum (SS) techniques, namely Direct Sequence and Frequency Hopping

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whenever band-limited partial-time interference is present and spectrum savings are required. This work will contain a summary of references [2-5] including concepts that describe the existing relationship between channel encoders and SS modulation techniques. Also, a general architecture [2] for SS systems is described. Both aspects enlarge the applicability of the spatial diversity procedures, to be reported in this work. The main advantages of FDSS are based on the repetition of the desired signal along all the frequency diversity branches. Since the jammer cannot extend its bandwidth, yet preserving significant levels of interference density power, this allows the presence of several diversity components where the desired signal is unjammed providing optimum and suboptimum receivers of easy implementation. The basic architecture of the transmitter is shown in Figure 1. Every branch contains a pass-band filter which do not overlap in frequency without substantial attenuation. This is the key feature that makes the difference between FDSS and OFDM. The orthogonality between the diversity components is made from non-overlapping filters instead of using other transformation like DFT. The filters are implemented by a polyphase network [7] and the difference with OFDM traduces in the fact that, whenever a partial band interference is present, still some frequency bands are free of interference. The presence of the desired signal free of interference in several bands allows two key features of FDSS namely sub-optimum receiver and blind spatial processing.

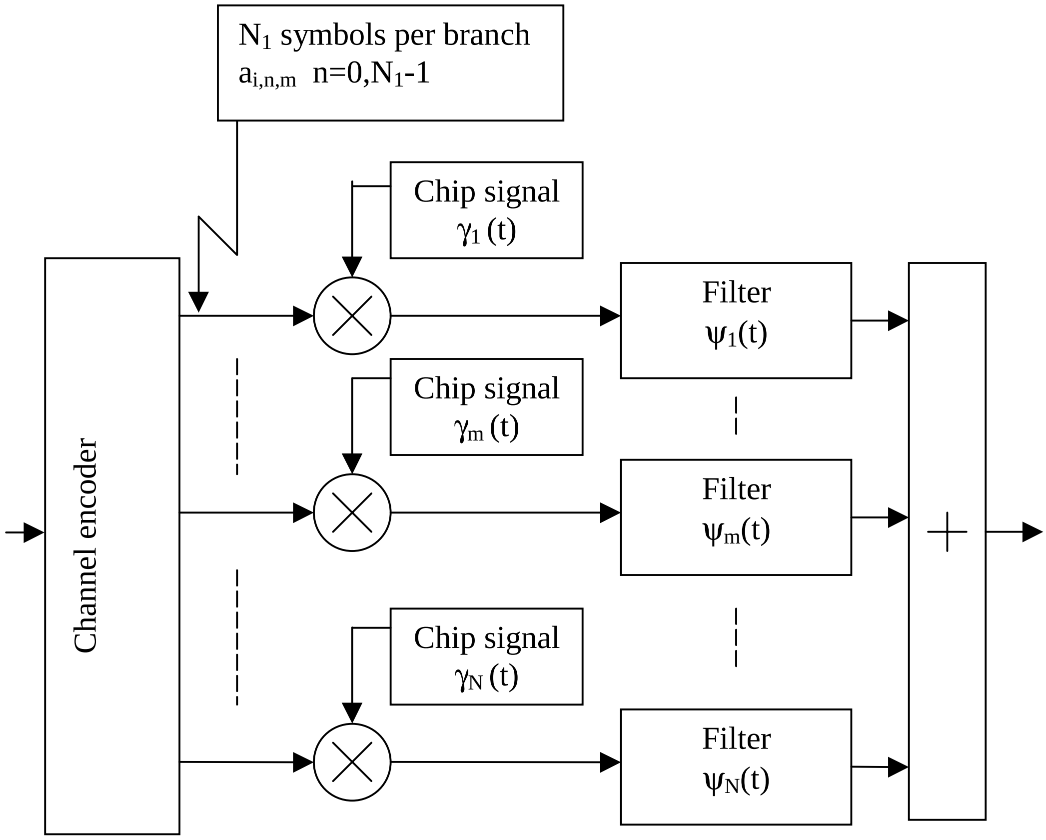

Fig. 1: General scheme for FDSS transmitter

This work also describes the adaptive design of the optimum beamforming for FDSS in the presence of partial band jamming. The adaptive algorithm designs the proper beamforming for every diversity branch, allowing optimum receiver performance under any jamming condition and scenario, without hard degradation. This includes the case of severe shadowing of the jammer when its aperture signature coincides with the desired signal. The adaptive algorithms provides an alternative to the block design which is more suitable for highly time varying scenarios.

Without loss of generality, it will be assumed that a pure repetition code (N,1) is used, thus the information symbol is repeated in every branch simultaneously. Also, it is assumed that a single chip per coded symbol is used. The advantage of the chip modulation is that, at the receiver, the interference will show up uncorrelated in every diversity branch jammed. In summary, the transmitted signal will be

$$X_T(t) = \sum_i \sum_{m=1}^{N} a_i\ \gamma_{i,m}\ \psi_m\left(\frac{t-iT_S}{T_S}\right) \tag{1}$$

where a_i is the information symbol, repeated in every frequency diversity branch, $\gamma_{i,m}$ are the chip symbols used at branch m, and $\Psi_m(.)$ is the temporal response of the filter at the same branch including roll-off. It is required that the bandwidth must be equal to the minimum bandwidth expected for interferences.

2. The FDSS receiver.

This section is a brief summary of the nice works reported by G.K. Kaleh ([2],[3] and [5]). It has been included here for the sake of presentation. Interested readers are highly recommended to go to the mentioned references. The sufficient statistic for the receiver is given by (2), where $z_{i,m}$ are the received samples and F_m is a set of weighting factors which depend on the relative strength of the jammer and the front-end noise,

$$\Lambda_i = \sum_{m=1}^{N} z_{i,m}\gamma^{*}_{i,m}F_m \tag{2}$$

being the factors equal to (3).

$$F_m = 1/(1+\left[\frac{J_o}{N_o}\right]) \quad \text{for bands hit by the jammer} \tag{3}$$

$$F_m = 1 \qquad \text{for free bands}$$

This formulation provides the final symbol error probability as a function of the average symbol energy and a factor β, which accounts for the jammer loss.

$$P_e = Q\left(\sqrt{\left(\frac{2E_S}{N_O}\right)\beta}\right) \tag{4}$$

The loss factor is equal to (5), where it is assumed that both the jammer and the front-end noise are spectrally flat in their respective frequency band support and η is the number, out of N, of the bands hit by the jammer. This expression reveals that optimum detection requires the estimation of the jammer strength.

$$\beta = \left\{\left[\left(\eta \Big/ (1+\left(J_O / N_O\right)) \right) + (1-\eta)\right]\right\} \tag{5}$$

A suboptimum detector is formed taking into account only the free of jamming bands. The resulting symbol error rate is

$$P_e = Q\left(\sqrt{\left(2E_S / N_O\right)(1-\eta)}\right) \tag{6}$$

The spatial processing will enhance the features of the optimum detector under any jamming condition. Even the case of severe shadowing can be incorporated to the detector as it can be viewed in [6].

3. Advanced front-end: Spatial Processing for FDSS.

Taking advantage of the frequency diversity available in FDSS systems, we propose a procedure to design optimum spatial diversity processing that is blind to the desired signal waveform. In fact, the procedure is valid for all the SS schemes which implement the architecture of Figure 1 for the transmitted signal generation, providing N is greater than one.

It is assumed that the desired signal is present in all the frequency bands. It is also assumed that the labeling of hit bands is properly done (This assumption is latter relaxed). All the air interfaces, down conversion, and filtering channels are calibrated. Full coherence is assumed for the desired signal across spatial diversity channels. Let us assume that $\underline{X}_{f,n}$ and $\underline{X}_{h,n}$ are the snapshots, at time n, of the free and hit bands, respectively. Both bands are selected to start up the design of the respective optimum spatial combiners $\underline{w}_f$ and $\underline{w}_h$. Once the chip symbol is removed for the two bands selected, i.e. $\underline{X}_{f,n} \Rightarrow \gamma^*_{f,n} . \underline{X}_{f,n}$ and $\underline{X}_{h,n} \Rightarrow \gamma^*_{h,n} . \underline{X}_{h,n}$, the desired signal, which is present in both bands can act as the reference signal. The design of both combiners can then be performed through the quadratic cost function minimization:

$$\xi = \mathrm{E} \left\{ \left| \underline{w}_f^H \underline{X}_{f,n} - \underline{w}_h^H \underline{X}_{h,n} \right|^2 \right\} \tag{7}$$

Assuming that the labeling is correct, the design procedure minimizes in the MSE between the spatial combiners' outputs of the free and hit bands. In addition, the cross-correlation of these two outputs is constrained to be different from zero in order to avoid trivial solutions

$$\mathrm{Re} \left\{ E \left[\underline{w}_f^H \underline{X}_{f,n} \underline{X}_{h,n}^H \underline{w}_h \right] \right\} = \phi \tag{8}$$

where ϕ is some constant different from zero.

This solution keeps a close resemblance with the so-called cross-score procedure reported in reference [8]. The major differences come from the fact that the above procedure is derived from the exact prediction property, existing due to the presence of the desired signal in every frequency branch, without resorting cyclostationarity [6].

The solution for the reference beamformer $\underline{w}_f$ is given as a generalized eigenvalue problem by (9.a). Once the reference beamformer has been derived, its output can be used as a time reference for the rest of the bands. Doing this temporal reference beamforming, the solution for the rest of the bands is (9.b).

$$\underline{\underline{R}}_f \underline{w}_f = (1+\lambda)^2 \underline{\underline{P}}_{f,h} \underline{\underline{R}}_h^{-1} \underline{\underline{P}}_{f,h}^H \underline{w}_f \tag{9.a}$$

$$\underline{w}_h = (1+\lambda) \underline{\underline{R}}_h^{-1} \underline{\underline{P}}_{f,h}^H \underline{w}_f \quad \begin{matrix} h = 1, N \\ h \neq f \end{matrix} \tag{9.b}$$

Note that for narrowband beamforming only two beamformers need to be derived, i.e. the reference for the free bands and a single one, following (9.b) for the hit bands. Nevertheless, the direct use of (9) is highly recommended since, doing in this way, the procedure is blind, without requiring the proper labeling of free and hit bands. Note that due to the chip demodulation, the jammer is uncorrelated from band to band and, in cosequence any couple of them can be used in (9).

An important aspect, usually overpassed in the literature, is the impact of the spatial processing in the design of the detector. One of the major advantages of the reported procedure is that easily incorporates the beamformer in the sufficient statistics. Furthermore, and estimation for the optimum factors F_m can be derived [6]. The factor for every branch is given by

$$F_m = \frac{\dfrac{\underline{w}_m^H \underline{\underline{P}}_{mf} \underline{w}_f}{\underline{w}_f^H \underline{\underline{R}}_f \underline{w}_f}}{\left(\left[MSE(m) \Big/ N_o \left|\underline{w}_f\right|^2 \right]^{-1}\right)} \tag{10}$$

This factor accounts for the weighting of the spatial processing front-end. Its formulation depends basically in the reference beamformer $\underline{w}_f$. The MSE residual, the background noise density N_o and the cross-correlation with the reference snapshot are the rest of the terms involved in the likelihood factor for every band. This estimate, derived in [6], proves to be efficient even under severe shadowing of the transmitter by the jammer. This severe shadowing occurs, in a point source model, when the jammer is just in the same DOA that the transmitter.

4. Adaptive algorithm.

The adaptive algorithm is based on the instantaneous gradient of the Lagrangian. It should be emphasized that the fundamental step is to find the response of the free-band combiner $\underline{w}_f$; once this combiner is found at each iteration, the update of the rest of the combiners, including $\underline{w}_h$, is done in a time reference combining framework, where the output of $\underline{w}_f$ acts as the reference waveform. The basic LMS updates are

$$\begin{aligned} \underline{w}_{f,n+1} &= \underline{w}_{f,n} - \mu_{f,n} \underline{X}_{f,n} \left(y_{f,n}^* - \lambda_n y_{h,n}^* \right) \\ \underline{w}_{h,n+1} &= \underline{w}_{h,n} - \mu_{h,n} \underline{X}_{h,n} \left(y_{h,n}^* - \lambda_n y_{f,n}^* \right) \end{aligned} \tag{11}$$

Factor λ_n is the Lagrange multiplier and its role in the algorithm, for low missadjusments usually far below 10%, is like an automatic control gain. The optimum setting for this parameter depends on the running power at the beamformers output (12) and the cross-correlation value set in the constrained minimization.

$$\lambda_n = \frac{\phi_n}{\left(P_{h,n} + P_{f,n}\right)} \tag{12}$$

The step-sizes are the standard ones for a LMS algorithm, which depends on the snapshot power (14) and α is the desired missadjusment factor.

$$\mu_{f,n} = \frac{\alpha}{W_{f,n}} \ ; \ \mu_{h,n} = \frac{\alpha}{W_{h,n}} \tag{14}$$

This new algorithm, reported in [6] and detailed in [9], outperforms the alternative, reported in [10] for the quadratic constrained minimization in a FH communication system.

5. A simulation example.

Figure 2 shows the performance of the algorithm when the FDSS front-end was formed by 7 diversity branches, where the desired signal was included with SNR equal to 0 dB per branch, BPSK modulated. The jammer with SNR of 10 dB was included in 3 non-successive diversity components. An ULA array of 5 sensors was used. The DOAs of the desired and the jammer were 10 and 30 º respectively. In figure 2 it can be viewed the detail of the learning curve. The algorithm has been evaluated for other different scenarios showing not significant changes in neither convergence rate nor missadjustment.

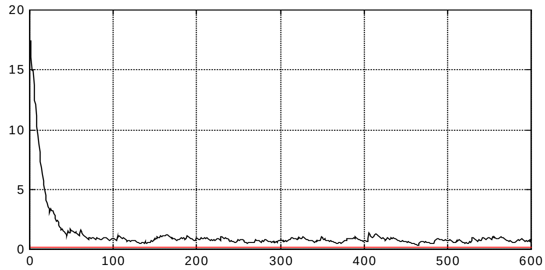

Fig. 2: MSE Evolution for the adaptive beamforming algorithm.

The spatial response of the beamformers can be viewed in figure 3 for a free band and for a hit band.

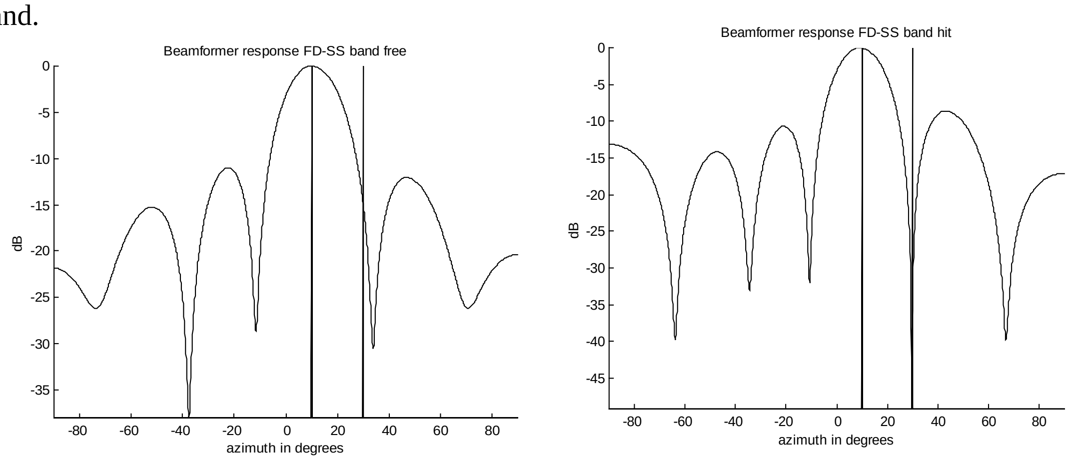

Fig. 3: Spatial response for a free (left) and hit (right) bands

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