Single and Multiple Antennas Alamouti Receivers for the Reception of Real-Valued Signals Corrupted by Interferences - The Alamouti SAIC/MAIC Concept

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Abstract - Several Interference Cancellation (IC) schemes have been developed during this last decade for wireless networks to mitigate the effect of intra-network interferences, when each user is equipped with multiple antennas and employs Space Time Block Code (STBC) at transmission. However, these IC techniques all require multiple antennas at reception, which remains a challenge at the handset level due to cost and size limitations. In this context, the purpose of this paper is to develop Single Antenna IC (SAIC) techniques, and their Multiple antennas extensions (MAIC), to mitigate the effect of both intra-network and external interferences for users using both real-valued constellations, such as ASK constellations, and the Alamouti's scheme at transmission.

I. INTRODUCTION

Orthogonal STBC, and the Alamouti scheme [1] in particular, are of particular interest in Multiple-Input Multiple-Output (MIMO) systems since they achieve full spatial diversity over fading channels and are decoded from linear processing at the receiver. Nevertheless, due to the expensive spectral resource, increasing network capacity requires the development of IC techniques allowing several users to share the same spectral resources without impacting the transmission quality. In this context several IC schemes have been developed during this last decade, where each user is equipped with multiple antennas and employs STBC at transmission [10], [11], [15], [14], [7], [8], [4]. However, these IC techniques all require multiple antennas at reception, which remains a challenge at the handset level due to cost and size limitations. For this reason, low complexity Single Antenna Interference Cancellation (SAIC) techniques [16], [3], [9] currently standardized and operational in GSM handsets [2], have been developed recently for single antenna and single carrier users using real-valued modulations or complex filtering of real-valued modulations, by using a widely linear (WL) filtering [13] at reception. Extension to multiple antennas at reception is called Multiple Antenna Interference Cancellation (MAIC) technique. As SAIC technology remains of great interest for 4G wireless networks, an extension of this technology to Orthogonal Frequency Division Multiplex transmission using one transmit antenna and the real-valued Amplitude Shift Keying (ASK) modulation has been presented very recently in [5]. Despite of the fact that ASK modulation is less power efficient than a corresponding complex QAM modulation, additional degrees of freedom are available and can be exploited for interference suppression at the receiver. Besides, it has been reported in [6] for DS-CDMA transmission and later in [5] for OFDM links, that transmission using real-valued data symbols with a WL receiver can lead to a higher spectral efficiency than using a complex symbol alphabet with linear receivers. In this context, the purpose of this paper is to extend the SAIC/MAIC technology to users using both real-valued constellations, such as ASK constellations, and the Alamouti's scheme at transmission. The ML and several WL MMSE receivers are developed, analyzed and compared in the presence of interferences. The best receivers are able to separate up to $2N$ Alamouti users from N antennas at reception, hence SAIC capability for $N = 1$.

II. HYPOTHESES, MODELS, STATISTICS AND WL FILTERING

A. Hypotheses

We consider a radio communication system that employs a real-valued constellation, the well-known Alamouti scheme with $M = 2$ transmit antennas [1] and N receive antennas. We denote by T the symbol period. Under these assumptions, assuming flat fading propagation channels which are invariant over at least two successive symbol periods, $(2n - 1)T$ and $2nT$ respectively, the observation vector over these two symbol periods can be written as

$$\boldsymbol{x}_1(n) = \mu_1 a_{2n-1} \boldsymbol{h}_1 + \mu_2 a_{2n} \boldsymbol{h}_2 + \boldsymbol{b}_1(n) \quad (1)$$

$$\boldsymbol{x}_2(n) = -\mu_1 a_{2n} \boldsymbol{h}_1 + \mu_2 a_{2n-1} \boldsymbol{h}_2 + \boldsymbol{b}_2(n) \quad (2)$$

where $\boldsymbol{x}_1(n)$ and $\boldsymbol{x}_2(n)$ are the (Nx1) observation vectors at symbol periods $(2n - 1)T$ and $2nT$ respectively, the quantities a_n are i.i.d real-valued random variables corresponding to the transmitted symbols, μ_i (i = 1, 2) is a real scalar which controls the power of the two transmitted signals received by the array of antennas; $\boldsymbol{h}_i$ (i = 1, 2), such that $\mathrm{E}[\boldsymbol{h}_i^{\mathrm{H}}\boldsymbol{h}_i] = N$, is the normalized propagation channel vector between transmit antenna i (i = 1, 2) and the receive array of antennas; H means transpose and conjugate; $\boldsymbol{b}_1(n)$ and $\boldsymbol{b}_2(n)$ are the sampled total noise vector at sample times $(2n - 1)T$ and $2nT$ respectively, potentially composed of intra-network interferences, external interferences (not generated by the network itself) and background noise. Note that for frequency selective propagation channels, model (1), (2) may also describe, after the Discrete Fourier Transform operation, observation vectors associated with a given sub-carrier, over two successive OFDM symbols, of a MIMO OFDM transmission using the Alamouti scheme.

B. Models

Defining the ($2N$ x 1) vectors, $\boldsymbol{x}(n)$, $\boldsymbol{b}(n)$, $\boldsymbol{f}_1$ and $\boldsymbol{f}_2$ by $\boldsymbol{x}(n) \triangleq [\boldsymbol{x}_1(n)^{\mathrm{T}}, \boldsymbol{x}_2(n)^{\mathrm{T}}]^{\mathrm{T}}$, $\boldsymbol{b}(n) \triangleq [\boldsymbol{b}_1(n)^{\mathrm{T}}, \boldsymbol{b}_2(n)^{\mathrm{T}}]^{\mathrm{T}}$, $\boldsymbol{f}_1 \triangleq [\mu_1\, \boldsymbol{h}_1^{\mathrm{T}}, \mu_2\, \boldsymbol{h}_2^{\mathrm{T}}]^{\mathrm{T}}$ and $\boldsymbol{f}_2 \triangleq [\mu_2\, \boldsymbol{h}_2^{\mathrm{T}}, -\mu_1\, \boldsymbol{h}_1^{\mathrm{T}}]^{\mathrm{T}}$, and defining the (2 x 1) vector $\boldsymbol{a}(n)$ and the ($2N$ x 2) matrix F by $\boldsymbol{a}(n) \triangleq [a_{2n-1}, a_{2n}]^{\mathrm{T}}$ and $F \triangleq [\boldsymbol{f}_1, \boldsymbol{f}_2]$ respectively, equations (3) and (4) can be written in a more compact form given by

$$\boldsymbol{x}(n) = a_{2n-1}\, \boldsymbol{f}_1 + a_{2n}\, \boldsymbol{f}_2 + \boldsymbol{b}(n) \triangleq F\, \boldsymbol{a}(n) + \boldsymbol{b}(n) \qquad (3)$$

Most of Alamouti receivers currently available for IC of intra-network interferences [10], [11], [14], [7] exploit the information contained in the ($2N$ x 1) vector $\bar{\boldsymbol{x}}(n)$, defined by $\bar{\boldsymbol{x}}(n) \triangleq [\boldsymbol{x}_1(n)^{\mathrm{T}}, \boldsymbol{x}_2(n)^{\mathrm{H}}]^{\mathrm{T}}$. Defining the ($2N$ x 1) vectors $\bar{\boldsymbol{b}}(n)$, $\boldsymbol{g}_1$ and $\boldsymbol{g}_2$ by $\bar{\boldsymbol{b}}(n) \triangleq [\boldsymbol{b}_1(n)^{\mathrm{T}}, \boldsymbol{b}_2(n)^{\mathrm{H}}]^{\mathrm{T}}$, $\boldsymbol{g}_1 \triangleq [\mu_1\, \boldsymbol{h}_1^{\mathrm{T}}, \mu_2\, \boldsymbol{h}_2^{\mathrm{H}}]^{\mathrm{T}}$ and $\boldsymbol{g}_2 \triangleq [\mu_2\, \boldsymbol{h}_2^{\mathrm{T}}, -\mu_1\, \boldsymbol{h}_1^{\mathrm{H}}]^{\mathrm{T}}$, defining the ($2N$ x 2) matrix G by $G \triangleq [\boldsymbol{g}_1, \boldsymbol{g}_2]$ and using (1) and (2), observation vector $\bar{\boldsymbol{x}}(n)$ can be written as

$$\bar{\boldsymbol{x}}(n) = a_{2n-1}\, \boldsymbol{g}_1 + a_{2n}\, \boldsymbol{g}_2 + \bar{\boldsymbol{b}}(n) \triangleq G\, \boldsymbol{a}(n) + \bar{\boldsymbol{b}}(n) \qquad (4)$$

To introduce WL filtering of $\boldsymbol{x}(n)$ in the following, we define the extended observation vectors $\tilde{\boldsymbol{f}}_1$, $\tilde{\boldsymbol{f}}_2$, $\tilde{\boldsymbol{b}}(n)$ and $\tilde{\boldsymbol{x}}(n)$ by the ($4N$ x 1) vectors $\tilde{\boldsymbol{f}}_1 \triangleq [\boldsymbol{f}_1^{\mathrm{T}}, \boldsymbol{f}_1^{\mathrm{H}}]^{\mathrm{T}}$, $\tilde{\boldsymbol{f}}_2 \triangleq [\boldsymbol{f}_2^{\mathrm{T}}, \boldsymbol{f}_2^{\mathrm{H}}]^{\mathrm{T}}$, $\tilde{\boldsymbol{b}}(n) \triangleq [\boldsymbol{b}(n)^{\mathrm{T}}, \boldsymbol{b}(n)^{\mathrm{H}}]^{\mathrm{T}}$ and $\tilde{\boldsymbol{x}}(n) \triangleq [\boldsymbol{x}(n)^{\mathrm{T}}, \boldsymbol{x}(n)^{\mathrm{H}}]^{\mathrm{T}}$ respectively. Then, defining the ($4N$ x 2) matrix $\tilde{F}$ by $\tilde{F} \triangleq [\tilde{\boldsymbol{f}}_1, \tilde{\boldsymbol{f}}_2]$ and using (3), observation vector $\tilde{\boldsymbol{x}}(n)$ can be written as

$$\tilde{\boldsymbol{x}}(n) = a_{2n-1}\, \tilde{\boldsymbol{f}}_1 + a_{2n}\, \tilde{\boldsymbol{f}}_2 + \tilde{\boldsymbol{b}}(n) \triangleq \tilde{F}\, \boldsymbol{a}(n) + \tilde{\boldsymbol{b}}(n) \qquad (5)$$

Models (3), (4) and (5) describe the equivalent reception at time nT_b, where $T_b = 2T$ is the duration of a block of two symbols, and by a virtual array of N_e antennas ($N_e = 2N$ for (3), (4) and $N_e = 4N$ for (5)) of two NB uncorrelated sources (a_{2n-1} and a_{2n}) associated with the linearly independent virtual channel vectors $\boldsymbol{f}_1$ and $\boldsymbol{f}_2$ ((3)), $\boldsymbol{g}_1$ and $\boldsymbol{g}_2$ ((4)) and $\tilde{\boldsymbol{f}}_1$ and $\tilde{\boldsymbol{f}}_2$ ((5)) respectively and corrupted by a total noise.

C. Second Order Statistics

The SO statistics of the data correspond to the six matrices $R_{x1}(n) \triangleq \mathrm{E}[\boldsymbol{x}_1(n)\boldsymbol{x}_1(n)^{\mathrm{H}}]$, $R_{x2}(n) \triangleq \mathrm{E}[\boldsymbol{x}_2(n)\boldsymbol{x}_2(n)^{\mathrm{H}}]$, $R_{x12}(n) \triangleq \mathrm{E}[\boldsymbol{x}_1(n)\boldsymbol{x}_2(n)^{\mathrm{H}}]$, $C_{x1}(n) \triangleq \mathrm{E}[\boldsymbol{x}_1(n)\boldsymbol{x}_1(n)^{\mathrm{T}}]$, $C_{x2}(n) \triangleq \mathrm{E}[\boldsymbol{x}_2(n)\boldsymbol{x}_2(n)^{\mathrm{T}}]$, $C_{x12}(n) \triangleq \mathrm{E}[\boldsymbol{x}_1(n)\boldsymbol{x}_2 n)^{\mathrm{T}}]$. In the following, the expected value operation, E[.], is considered on a burst scale for which the channel vectors are constant. Under this assumption, and using (3) and (4), these matrices can be written as

$$R_{x1}(n) = \pi_1\, \boldsymbol{h}_1\boldsymbol{h}_1^{\mathrm{H}} + \pi_2\, \boldsymbol{h}_2\boldsymbol{h}_2^{\mathrm{H}} + R_1(n) \triangleq R_{s1} + R_1(n) \qquad (6)$$

$$R_{x2}(n) = \pi_1\, \boldsymbol{h}_1\boldsymbol{h}_1^{\mathrm{H}} + \pi_2\, \boldsymbol{h}_2\boldsymbol{h}_2^{\mathrm{H}} + R_2(n) \triangleq R_{s2} + R_2(n) \qquad (7)$$

$$R_{x12}(n) = \sqrt{\pi_1}\sqrt{\pi_2}[\boldsymbol{h}_1\boldsymbol{h}_2^{\mathrm{H}} - \boldsymbol{h}_2\boldsymbol{h}_1^{\mathrm{H}}] + R_{12}(n) \triangleq R_{s12} + R_{12}(n) \qquad (8)$$

$$C_{x1}(n) = \pi_1\, \boldsymbol{h}_1\boldsymbol{h}_1^{\mathrm{T}} + \pi_2\, \boldsymbol{h}_2\boldsymbol{h}_2^{\mathrm{T}} + C_1(n) \triangleq C_{s1} + C_1(n) \qquad (9)$$

$$C_{x2}(n) = \pi_1\, \boldsymbol{h}_1\boldsymbol{h}_1^{\mathrm{T}} + \pi_2\, \boldsymbol{h}_2\boldsymbol{h}_2^{\mathrm{T}} + C_2(n) \triangleq C_{s2} + C_2(n) \qquad (10)$$

$$C_{x12}(n) = \sqrt{\pi_1}\sqrt{\pi_2}[\boldsymbol{h}_1\boldsymbol{h}_2^{\mathrm{T}} - \boldsymbol{h}_2\boldsymbol{h}_1^{\mathrm{T}}] + C_{12}(n) \triangleq C_{s12} + C_{12}(n) \qquad (11)$$

where $\pi_1 \triangleq \mu_1^2\pi_a$ and $\pi_2 \triangleq \mu_2^2\pi_a$ are the mean power of the transmitted signals 1 and 2 per receive antenna respectively, with $\pi_a \triangleq \mathrm{E}[|a_n|^2]$; $R_1(n) \triangleq \mathrm{E}[\boldsymbol{b}_1(n)\boldsymbol{b}_1(n)^{\mathrm{H}}]$, $R_2(n) \triangleq \mathrm{E}[\boldsymbol{b}_2(n)\boldsymbol{b}_2(n)^{\mathrm{H}}]$, $R_{12}(n) \triangleq \mathrm{E}[\boldsymbol{b}_1(n)\boldsymbol{b}_2(n)^{\mathrm{H}}]$, $C_1(n) \triangleq \mathrm{E}[\boldsymbol{b}_1(n)\boldsymbol{b}_1(n)^{\mathrm{T}}]$, $C_2(n) \triangleq \mathrm{E}[\boldsymbol{b}_2(n)\boldsymbol{b}_2(n)^{\mathrm{T}}]$, $C_{12}(n) \triangleq \mathrm{E}[\boldsymbol{b}_1(n)\boldsymbol{b}_2(n)^{\mathrm{T}}]$ and where R_{s1}, R_{s2}, R_{s12}, C_{s1}, C_{s2} and C_{s12} are the six matrices of SO statistics of the received useful signal. Expression (6) to (11) show that the Alamouti scheme generates non zero matrices R_{s1}, R_{s2}, R_{s12}, C_{s1}, C_{s2} and C_{s12}, except C_{s12} which becomes zero for $N = 1$. Thus, whatever the value of N is, the observation vector $\boldsymbol{x}(n)$ is SO noncircular.

D. Linear and Widely Linear Filtering

Time invariant (TI) and linear filtering of $\boldsymbol{x}(n)$, $\bar{\boldsymbol{x}}(n)$ and $\tilde{\boldsymbol{x}}(n)$ are respectively defined by the input-output relations $y(n) = \boldsymbol{w}^{\mathrm{H}}\boldsymbol{x}(n)$, $y(n) = \bar{\boldsymbol{w}}^{\mathrm{H}}\, \bar{\boldsymbol{x}}(n)$ and $y(n) = \tilde{\boldsymbol{w}}^{\mathrm{H}}\, \tilde{\boldsymbol{x}}(n)$, where $\boldsymbol{w}$, $\bar{\boldsymbol{w}}$ and $\tilde{\boldsymbol{w}}$ are ($2N$ x 1), ($2N$ x 1) and ($4N$ x 1) complex vectors respectively. These input-output relations describe what we call in the following a linear, a partially WL and a fully WL filtering of $\boldsymbol{x}(n)$ respectively.

III. MAXIMUM LIKELIHOOD RECEIVER

A. Presentation

We deduce from (9), (10) and (11) that in the presence of at least one synchronous intra-network interference the total noise vector $\boldsymbol{b}(n)$ is SO noncircular. Assuming a Gaussian and noncircular vector $\boldsymbol{b}(n)$, despite of the fact that intra-network interferences are not Gaussian, the probability density of the latter, i.e. the joint probability density of the real and imaginary part of $\boldsymbol{b}(n)$ becomes [17], [12]

$$\mathrm{p}[\tilde{\boldsymbol{b}}(n)] \triangleq \pi^{-2N} \det[R_{\tilde{b}}(n)]^{-1/2} \exp[-(1/2)\, \tilde{\boldsymbol{b}}(n)^{\mathrm{H}} R_{\tilde{b}}(n)^{-1} \tilde{\boldsymbol{b}}(n)] \qquad (12)$$

where $R_{\tilde{b}}(n) = \mathrm{E}[\tilde{\boldsymbol{b}}(n)\tilde{\boldsymbol{b}}(n)^{\mathrm{H}}]$. Under this assumption, we deduce from (5) that the ML receiver for the demodulation of vector $\boldsymbol{a}(n)$ in SO noncircular total noise is such that $\boldsymbol{a}(n)$ maximizes the ML criterion defined by

$$C_{nc\text{-}ml}[\boldsymbol{a}(n)] \triangleq \mathrm{p}[\tilde{\boldsymbol{b}}(n) = \tilde{\boldsymbol{x}}(n) - \tilde{F}\, \boldsymbol{a}(n)_{/}\, \boldsymbol{a}(n)] \qquad (13)$$

Using (12) into (13), we easily deduce that the maximization of (13) is equivalent to the minimization of $C_{nc}[\boldsymbol{a}(n)]$ defined by

$$C_{nc}[\boldsymbol{a}(n)] \triangleq |a_{2n-1}|^2\, \tilde{\boldsymbol{f}}_1^{\mathrm{H}} R_{\tilde{b}}(n)^{-1}\, \tilde{\boldsymbol{f}}_1 + |a_{2n}|^2\, \tilde{\boldsymbol{f}}_2^{\mathrm{H}} R_{\tilde{b}}(n)^{-1} \tilde{\boldsymbol{f}}_2 + 2\, a_{2n-1}\, a_{2n}\, \tilde{\boldsymbol{f}}_1^{\mathrm{H}} R_{\tilde{b}}(n)^{-1} \tilde{\boldsymbol{f}}_2 - 2[a_{2n-1}\, \tilde{\boldsymbol{f}}_1^{\mathrm{H}} R_{\tilde{b}}(n)^{-1} \tilde{\boldsymbol{x}}(n) + a_{2n}\, \tilde{\boldsymbol{f}}_2^{\mathrm{H}} R_{\tilde{b}}(n)^{-1} \tilde{\boldsymbol{x}}(n)] \quad (14)$$

where $\tilde{\boldsymbol{f}}_1^{\mathrm{H}} R_{\tilde{b}}(n)^{-1} \tilde{\boldsymbol{f}}_2$, $\tilde{\boldsymbol{f}}_1^{\mathrm{H}} R_{\tilde{b}}(n)^{-1} \tilde{\boldsymbol{x}}(n)$ and $\tilde{\boldsymbol{f}}_2^{\mathrm{H}} R_{\tilde{b}}(n)^{-1} \tilde{\boldsymbol{x}}(n)$ are real-valued quantities. The receiver which generates $\boldsymbol{a}(n)$ minimizing (14) is called the NC-ML receiver (ML receiver in potentially SO noncircular total noise). Its implementation requires the knowledge of $R_{\tilde{b}}(n)$, i.e. a Total Noise Alone Reference (TNAR), and vectors $\tilde{\boldsymbol{f}}_1$ and $\tilde{\boldsymbol{f}}_2$, i.e. $\mu_1\, \boldsymbol{h}_1$ and $\mu_2\, \boldsymbol{h}_2$. The NC-ML receiver exploits, in a SO optimal manner, all the information contained in $R_{\tilde{b}}(n)$, i.e. in $R_1(n)$, $R_2(n)$, $R_{12}(n)$, $C_1(n)$, $C_2(n)$ and $C_{12}(n)$. It is a coupled receiver in the general case of arbitrary matrix $R_{\tilde{b}}(n)$ and vectors $\mu_1\, \boldsymbol{h}_1$ and $\mu_2\, \boldsymbol{h}_2$, which means that it requires the joint estimation of a_{2n-1} and a_{2n}. This generates M^2 possibilities for vector $\boldsymbol{a}(n)$, where M is the number of states of the constellation.

B. Decoupling condition

We deduce from (14) that when $\tilde{\boldsymbol{f}}_1^{\mathrm{H}} R_{\tilde{b}}(n)^{-1} \tilde{\boldsymbol{f}}_2 = 0$, the minimization of $C_{nc}[\boldsymbol{a}(n)]$ reduces to the independent minimization of $C_{nc,1}[a_{2n-1}]$ and $C_{nc,2}[a_{2n}]$, with respect to a_{2n-1} and a_{2n} respectively, defined by

$$C_{nc,1}[a_{2n-1}] \triangleq |a_{2n-1}|^2 \tilde{\boldsymbol{f}}_1^{\mathrm{H}} R_{\tilde{b}}(n)^{-1} \tilde{\boldsymbol{f}}_1 - 2a_{2n-1} \tilde{\boldsymbol{f}}_1^{\mathrm{H}} R_{\tilde{b}}(n)^{-1} \tilde{\boldsymbol{x}}(n)$$
$$C_{nc,2}[a_{2n}] \triangleq |a_{2n}|^2 \tilde{\boldsymbol{f}}_2^{\mathrm{H}} R_{\tilde{b}}(n)^{-1} \tilde{\boldsymbol{f}}_2 - 2a_{2n}\, \tilde{\boldsymbol{f}}_2^{\mathrm{H}} R_{\tilde{b}}(n)^{-1} \tilde{\boldsymbol{x}}(n) \quad (15)$$

In this case, the NC-ML receiver becomes decoupled, which means that the estimations of a_{2n-1} and a_{2n} are done separately. This reduces the complexity of the search procedure to the test of $2M$ possibilities for $\boldsymbol{a}(n)$ instead of M^2. Note that the conventional Alamouti receiver [1] corresponds to (15) with $R_{\tilde{b}}(n) = \eta_2$ I, where η_2 is the mean power of the background noise per receive antenna.

IV. MMSE RECEIVERS

An alternative to the ML receiver corresponds to the MMSE-based receivers, which are decoupled receivers whose implementation does not require any TNAR, hence their great interest in practice. For this reason, we present in this section several MMSE-based receivers and we compare them to the ML receiver.

A. Presentation

A MMSE-based receiver for the demodulation of symbol a_{2n-1} implements a conventional ML receiver from the output of a MMSE filter for symbol a_{2n-1}. Two MMSE filters corresponding to the partially WL and the fully WL MMSE filters, are considered in the following. They give rise to two MMSE-based receivers called Partially WL MMSE (P-WL-MMSE) and Fully WL MMSE (F-WL-MMSE) receivers respectively. A MMSE filter for symbol a_{2n-1} minimizes the Mean Square Error (MSE), $\mathrm{E}[|a_{2n-1} - y_1(n)|^2]$, between its output $y_1(n)$ and symbol a_{2n-1}. It is then easy to show that the partially WL and the fully WL MMSE filters for symbol a_{2n-1} are respectively defined by

$$\bar{\boldsymbol{w}}_{1,mmse}(n) = R_{\bar{x}}(n)^{-1} \boldsymbol{r}_{\bar{x}a_{2n-1}}(n) \quad (16)$$
$$\tilde{\boldsymbol{w}}_{1,mmse}(n) = R_{\tilde{x}}(n)^{-1} \boldsymbol{r}_{\tilde{x}a_{2n-1}}(n) \quad (17)$$

where $\boldsymbol{r}_{\bar{x}a_{2n-1}}(n) \triangleq \mathrm{E}[\bar{\boldsymbol{x}}(n)\ a_{2n-1}] = \pi_a\ \boldsymbol{g}_1$, $\boldsymbol{r}_{\tilde{x}a_{2n-1}}(n) \triangleq \mathrm{E}[\tilde{\boldsymbol{x}}(n)\ a_{2n-1}] = \pi_a\ \tilde{\boldsymbol{f}}_1$. It is then easy to verify from (4) and (5) that the output $y_1(n)$ of each of these filters takes the form $y_1(n) = \alpha_1(n)\ a_{2n-1} + b_1(n)$, where a_{2n-1} is the symbol to demodulate, $\alpha_1(n)$ is a real quantity and $b_1(n)$ is the global noise for the symbol a_{2n-1}. Assuming a Gaussian global noise $b_1(n)$, a conventional ML receiver from $y_1(n)$ generates the symbol a_{2n-1} which minimizes $|\alpha_1(n)\ a_{2n-1} - y_1(n)|^2$, i.e. $\alpha_1(n)\ |a_{2n-1}|^2 - 2a_{2n-1}\mathrm{Re}[y_1(n)]$, which shows that $\mathrm{Re}[y_1(n)]$ is a sufficient statistic for the conventional ML estimation of a_{2n-1} from $y_1(n)$. We deduce from these results that the P-WL-MMSE receiver generates the symbols a_{2n-1} and a_{2n} minimizing respectively the criterions

$$\mathrm{C}_{pwl,1}(a_{2n-1}) = (\bar{\boldsymbol{w}}_{1,mmse}(n)^{\mathrm{H}} \boldsymbol{r}_{\bar{x}a_{2n-1}}(n) / \pi_a)\ |a_{2n-1}|^2 - 2\ a_{2n-1}\ \mathrm{Re}[\bar{\boldsymbol{w}}_{1,mmse}(n)^{\mathrm{H}}\ \bar{\boldsymbol{x}}(n)] \quad (18)$$
$$\mathrm{C}_{pwl,2}(a_{2n}) = (\bar{\boldsymbol{w}}_{2,mmse}(n)^{\mathrm{H}} \boldsymbol{r}_{\bar{x}a_{2n}}(n) / \pi_a)\ |a_{2n}|^2 - 2\ a_{2n}\ \mathrm{Re}[\bar{\boldsymbol{w}}_{2,mmse}(n)^{\mathrm{H}}\ \bar{\boldsymbol{x}}(n)] \quad (19)$$

where $\bar{\boldsymbol{w}}_{2,mmse} = R_{\bar{x}}(n)^{-1} \boldsymbol{r}_{\bar{x}a_{2n}}(n)$ and its implementation requires the knowledge of $R_{\bar{x}}(n)$, $\boldsymbol{r}_{\bar{x}a_{2n-1}}(n)$, $\boldsymbol{r}_{\bar{x}a_{2n}}(n)$ and π_a. In a same way, the F-WL-MMSE receiver generates the symbols a_{2n-1} and a_{2n} minimizing respectively the criterions

$$\mathrm{C}_{fwl,1}(a_{2n-1}) = (\tilde{\boldsymbol{w}}_{1,mmse}(n)^{\mathrm{H}} \boldsymbol{r}_{\tilde{x}a_{2n-1}}(n) / \pi_a)\ |a_{2n-1}|^2 - 2\ a_{2n-1}\ \mathrm{Re}[\tilde{\boldsymbol{w}}_{1,mmse}(n)^{\mathrm{H}}\ \tilde{\boldsymbol{x}}(n)] \quad (20)$$
$$\mathrm{C}_{fwl,2}(a_{2n}) = (\tilde{\boldsymbol{w}}_{2,mmse}(n)^{\mathrm{H}} \boldsymbol{r}_{\tilde{x}a_{2n}}(n) / \pi_a)\ |a_{2n}|^2 - 2\ a_{2n}\ \mathrm{Re}[\tilde{\boldsymbol{w}}_{2,mmse}(n)^{\mathrm{H}}\ \tilde{\boldsymbol{x}}(n)] \quad (21)$$

where $\tilde{\boldsymbol{w}}_{2,mmse}(n) = R_{\tilde{x}}(n)^{-1} \boldsymbol{r}_{\tilde{x}a_{2n}}(n)$ and its implementation requires the knowledge of $R_{\tilde{x}}(n)$, $\boldsymbol{r}_{\tilde{x}a_{2n-1}}(n)$, $\boldsymbol{r}_{\tilde{x}a_{2n}}(n)$ and π_a. Moreover, it is straightforward to show that $\bar{\boldsymbol{w}}_{1,mmse}(n)^{\mathrm{H}} \boldsymbol{r}_{\bar{x}a_{2n-1}}(n) / \pi_a \approx 1$ if $\pi_a\, \boldsymbol{g}_1^{\mathrm{H}} R_{\bar{G}}(n)^{-1} \boldsymbol{g}_1 >> 1$ and $\tilde{\boldsymbol{w}}_{1,mmse}(n)^{\mathrm{H}} \boldsymbol{r}_{\tilde{x}a_{2n-1}}(n) / \pi_a \approx 1$ if $\pi_a\ \tilde{\boldsymbol{f}}_1^{\mathrm{H}} R_{\tilde{G}}(n)^{-1} \tilde{\boldsymbol{f}}_1 >> 1$, where $R_{\bar{G}}(n) = R_{\bar{b}}(n) + \pi_a\ \boldsymbol{g}_2\ \boldsymbol{g}_2^{\mathrm{H}}$, $R_{\bar{b}}(n) = \mathrm{E}[\bar{\boldsymbol{b}}(n)\ \bar{\boldsymbol{b}}(n)^{\mathrm{H}}]$ and $R_{\tilde{G}}(n) = R_{\tilde{b}}(n) + \pi_a\ \tilde{\boldsymbol{f}}_2\ \tilde{\boldsymbol{f}}_2^{\mathrm{H}}$. The first and second conditions are verified when the symbol a_{2n-1} is received with a good Signal to Interference plus Noise Ratio at the output of the filters (16) and (17) respectively. A similar analysis can be done for the symbol a_{2n}. For these reasons, we define the Approximated P-WL-MMSE (AP-WL-MMSE) receiver by equations (18) and (19) in which the terms $\bar{\boldsymbol{w}}_{1,mmse}(n)^{\mathrm{H}} \boldsymbol{r}_{\bar{x}a_{2n-1}}(n) / \pi_a$ and $\bar{\boldsymbol{w}}_{2,mmse}(n)^{\mathrm{H}} \boldsymbol{r}_{\bar{x}a_{2n}}(n) / \pi_a$ have been replaced by 1. Similarly, we define the Approximated F-WL-MMSE (AF-WL-MMSE) receiver by equations (20) and (21) in which the terms $\tilde{\boldsymbol{w}}_{1,mmse}(n)^{\mathrm{H}}\boldsymbol{r}_{\tilde{x}a_{2n-1}}(n) / \pi_a$ and $\tilde{\boldsymbol{w}}_{2,mmse}(n)^{\mathrm{H}}\boldsymbol{r}_{\tilde{x}a_{2n}}(n) / \pi_a$ have been replaced by 1.

B. Properties

B1. *F-WL-MMSE receiver*

It is easy to show that when $\tilde{\boldsymbol{f}}_1^{\mathrm{H}} R_{\tilde{b}}(n)^{-1} \tilde{\boldsymbol{f}}_2 = 0$ (C1), vectors $R_{\tilde{x}}(n)^{-1}\tilde{\boldsymbol{f}}_1$ and $R_{\tilde{b}}(n)^{-1}\tilde{\boldsymbol{f}}_1$ are collinear, which means that $\mathrm{Re}[\tilde{\boldsymbol{f}}_1^{\mathrm{H}} R_{\tilde{x}}(n)^{-1}\tilde{\boldsymbol{x}}(n)]$ and $\mathrm{Re}[\tilde{\boldsymbol{f}}_1^{\mathrm{H}} R_{\tilde{b}}(n)^{-1}\tilde{\boldsymbol{x}}(n)]$ are proportional to each other. A similar result would be obtained for $\mathrm{Re}[\tilde{\boldsymbol{f}}_2^{\mathrm{H}} R_{\tilde{x}}(n)^{-1}\tilde{\boldsymbol{x}}(n)]$ and $\mathrm{Re}[\tilde{\boldsymbol{f}}_2^{\mathrm{H}} R_{\tilde{b}}(n)^{-1}\tilde{\boldsymbol{x}}(n)]$. We then deduce from (15), (20) and (21) that the F-WL-MMSE receiver corresponds in this case to the NC-ML receiver and becomes optimal. For a given total noise vector $\boldsymbol{b}(n)$, the condition C1 may be verified only for some particular channel vectors $\mu_1 \boldsymbol{h}_1$ and $\mu_2 \boldsymbol{h}_2$. Nevertheless, it is possible to show, after some elementary algebraic manipulations, that condition C1 is verified for all channel vectors $\mu_1 \boldsymbol{h}_1$ and $\mu_2 \boldsymbol{h}_2$, if and only if $\boldsymbol{b}(n)$ verifies condition C1' defined by

C1': $R_1(n) = R_2(n)$; $C_1(n) = C_2(n)$;
$R_{12}(n)^{\mathrm{H}} = -R_{12}(n)$; $C_{12}(n)^{\mathrm{T}} = -C_{12}(n)$

Condition C1' is verified in the absence of interference for a circular, temporally and spatially white noise vector $\boldsymbol{b}(n)$. It is also verified in the presence of one or several synchronous intra-network interferences as it can be deduced from (6) to (11). It is still verified in the presence of external interferences as long as $\boldsymbol{b}_1(n)$ and $\boldsymbol{b}_2(n)$ remain uncorrelated with the same SO statistics. Hence the optimality of the F-WL-MMSE receiver in numerous situations of practical interest.

B2. *P-WL-MMSE receiver*

It is easy to show that when $\boldsymbol{g}_1^{\mathrm{H}} R_{\bar{b}}(n)^{-1} \boldsymbol{g}_2 = 0$, vectors $R_{\bar{x}}(n)^{-1} \boldsymbol{g}_1$ and $R_{\bar{b}}(n)^{-1} \boldsymbol{g}_1$ are collinear. As $R_{\bar{b}}(n)^{-1} \boldsymbol{g}_1$ only exploits the information contained in matrices $R_1(n)$, $R_2(n)$ and $C_{12}(n)$, a necessary condition for $\mathrm{Re}[\boldsymbol{g}_1^{\mathrm{H}} R_{\bar{x}}(n)^{-1}\bar{\boldsymbol{x}}(n)]$ and $\mathrm{Re}[\boldsymbol{g}_1^{\mathrm{H}} R_{\bar{b}}(n)^{-1}\bar{\boldsymbol{x}}(n)]$ to be proportional to each other is that $C_1(n) = C_2(n) = R_{12}(n) = 0$. Conversely, assuming that $C_1(n) = C_2(n) = R_{12}(n) = 0$, it is straightforward to show that $\mathrm{Re}[\boldsymbol{g}_1^{\mathrm{H}} R_{\bar{x}}(n)^{-1}\bar{\boldsymbol{x}}(n)]$ and $\mathrm{Re}[\boldsymbol{g}_1^{\mathrm{H}} R_{\bar{b}}(n)^{-1}\bar{\boldsymbol{x}}(n)]$ are proportional and that $\tilde{\boldsymbol{f}}_1^{\mathrm{H}} R_{\tilde{b}}(n)^{-1}\tilde{\boldsymbol{f}}_2 = 2\mathrm{Re}[\boldsymbol{g}_1^{\mathrm{H}} R_{\bar{b}}(n)^{-1}\boldsymbol{g}_2]$. A similar result would be obtained for $\mathrm{Re}[\boldsymbol{g}_2^{\mathrm{H}} R_{\bar{x}}(n)^{-1}\bar{\boldsymbol{x}}(n)]$ and $\mathrm{Re}[\boldsymbol{g}_2^{\mathrm{H}} R_{\bar{b}}(n)^{-1}\bar{\boldsymbol{x}}(n)]$. We then deduce from these results and from (15), (18) and (19), that when C2 is verified, where C2 is defined by

C2 : $\boldsymbol{g}_1^{\mathrm{H}} R_{\bar{b}}(n)^{-1}\boldsymbol{g}_2 = 0$ and $C_1(n) = C_2(n) = R_{12}(n) = 0$

the P-WL-MMSE receiver corresponds to NC-ML receiver and becomes optimal. For a given total noise vector $\boldsymbol{b}(n)$, the condition $\boldsymbol{g}_1^{\mathrm{H}} R_{\bar{b}}(n)^{-1}\boldsymbol{g}_2 = 0$ may be verified only for some particular channel vectors $\mu_1\boldsymbol{h}_1$ and $\mu_2\boldsymbol{h}_2$. Nevertheless, it is possible to show, after some elementary algebraic manipulations, that the condition $\boldsymbol{g}_1^{\mathrm{H}} R_{\bar{b}}(n)^{-1}\boldsymbol{g}_2 = 0$ is verified for all channel vectors $\mu_1\boldsymbol{h}_1$ and $\mu_2\boldsymbol{h}_2$, if and only if $\boldsymbol{b}(n)$ is such that $R_1(n) = R_2(n)$ and $C_{12}(n)^{\mathrm{T}} = -C_{12}(n)$. Consequently, the P-WL-MMSE receiver is optimal for all channel vectors $\mu_1\boldsymbol{h}_1$ and $\mu_2\boldsymbol{h}_2$ if and only if $\boldsymbol{b}(n)$ verifies condition C2' defined by

C2': $R_1(n) = R_2(n)$; $C_1(n) = C_2(n) = R_{12}(n) = 0$;
$C_{12}(n)^{\mathrm{T}} = -C_{12}(n)$

In this case, the F-WL-MMSE reduces to a P-WL-MMSE. Condition C2' is verified in the absence of interference for a circular, temporally and spatially white noise vector $\boldsymbol{b}(n)$. It is also verified in the presence of SO circular external interferences as long as $\boldsymbol{b}_1(n)$ and $\boldsymbol{b}_2(n)$ remain uncorrelated with the same SO statistics. However, the P-WL-MMSE receiver, used in [10], [11], [14], [7], becomes sub-optimal in the presence of one or several intra-network interferences for which $C_1(n) = C_2(n) \neq 0$ and $R_{12}(n) \neq 0$. It remains sub-optimal in the presence of external interferences which are either SO noncircular or such that $\boldsymbol{b}_1(n)$ and $\boldsymbol{b}_2(n)$ are correlated, which is in particular the case for very NB external interferences.

C. Adaptive Implementation

In practice, $R_{\bar{x}}(n)$, $R_{\tilde{x}}(n)$, $\boldsymbol{r}_{\bar{x}a_{2n-1}}(n)$, $\boldsymbol{r}_{\bar{x}a_{2n}}(n)$, $\boldsymbol{r}_{\tilde{x}a_{2n-1}}(n)$ and $\boldsymbol{r}_{\tilde{x}a_{2n}}(n)$ are not known and have to be estimated from K training blocks of $2K$ symbols introduced in the Alamouti scheme. Assuming the constant value of the previous statistics over a burst containing both useful blocks and K training blocks, $R_{\tilde{x}}(n)$ and $\boldsymbol{r}_{\tilde{x}a_{2n-1}}(n)$ may be estimated by $\hat{R}_{\tilde{x}}(n)$ and $\hat{\boldsymbol{r}}_{\tilde{x}a_{2n-1}}(n)$ defined by

$$\hat{R}_{\tilde{x}}(n) \triangleq \frac{1}{K} \sum_{k=0}^{K-1} \tilde{\boldsymbol{x}}(k+k_0)\, \tilde{\boldsymbol{x}}(k+k_0)^{\mathrm{H}} \tag{22}$$

$$\hat{\boldsymbol{r}}_{\tilde{x}a_{2n-1}}(n) \triangleq \frac{1}{K} \sum_{k=0}^{K-1} \tilde{\boldsymbol{x}}(k+k_0)\, a_{2(k+k_0)-1} \tag{23}$$

where k_0 is the position of the first training block in the burst. Estimation of $\hat{\boldsymbol{r}}_{\tilde{x}a_{2n}}(n)$ is given by (23) with $a_{2(k+k_0)}$ instead of $a_{2(k+k_0)-1}$. Estimation of $R_{\bar{x}}(n)$, $\boldsymbol{r}_{\bar{x}a_{2n-1}}(n)$ and $\boldsymbol{r}_{\bar{x}a_{2n}}(n)$ are similar to that of $R_{\tilde{x}}(n)$, $\boldsymbol{r}_{\tilde{x}a_{2n-1}}(n)$ and $\boldsymbol{r}_{\tilde{x}a_{2n}}(n)$ with $\bar{\boldsymbol{x}}(n)$ instead of $\tilde{\boldsymbol{x}}(n)$.

V. Performance in presence of interferences

A. Hypotheses and processing capacity

We assume in this section that the total noise $\boldsymbol{b}(n)$ is composed of P_{int} synchronous intra-network (or internal) interferences, corresponding to other Alamouti users of the network with the same rectilinear modulation as the useful signal, P_{ext} external interferences, coming from other networks or jamming, and a background noise. We denote by $m_i(t)$ the complex envelope of the external interference i. An external interference i is said to be rectilinear if $m_i(t)^* = m_i(t)\, e^{j\phi i}$ and nonrectilinear otherwise, where * means complex conjugate . It is said to be coherent if $m_i((2n - 1)T) \approx m_i(2nT)\, e^{j\psi_i}$ and noncoherent otherwise. A coherent interference corresponds to a very NB signal compared to the useful signal. Under these assumptions, we assume that the P_{ext} external interferences are composed of P_{rc} rectilinear and coherent interferences, P_{rnc} rectilinear and noncoherent interferences, P_{nrc} nonrectilinear and coherent interferences and P_{nrnc} nonrectilinear and noncoherent interferences such that $P_{ext} = P_{rc} + P_{rnc} + P_{nrc} + P_{nrnc}$. Under these assumptions, it is easy to show that the maximal number of interferences which may be processed by the P-WL-MMSE and the F-WL-MMSE receivers are such that

P-WL: $2P_{int} + P_{rc} + 2P_{rnc} + 2P_{nrc} + 2P_{nrnc} \leq 2N - 2$

F-WL: $2P_{int} + P_{rc} + 2P_{rnc} + 2P_{nrc} + 4P_{nrnc} \leq 4N - 2$

This shows in particular the Single Antenna Intra-network Interference Cancellation Capability of the F-WL-MMSE receiver, contrary to the P-WL-MMSE receiver.

C. Performance illustration

To illustrate the performance of the previous receivers, we consider a mono-sensor reception ($N = 1$) and we assume that the useful ASK signal with $M = 4$ states (± 1, ± 3) is corrupted by one synchronous internal interference with an Interference to Signal Ratio equal to 10 dB. The useful signal is such that $\pi_1 = \pi_2$ and we note $\pi_s = \pi_1 + \pi_2 = 2\pi_1$. Same assumptions hold for the interference. The channel vectors of all the signals are assumed to be constant over a burst composed of 56 blocks of couples of information symbols and K blocks of couple of training symbols. The channels vectors are zero-mean i.i.d Gaussian from a burst to another with independent components. The number of bursts used for the simulations is 1 000 000.

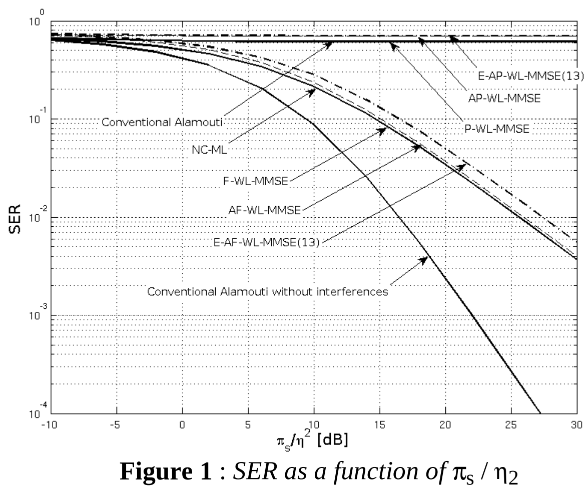

Figure 1 : *SER as a function of* π_s / η_2

Under these assumptions Fig 1 shows the variations of the symbol error rate (SER) at the output of several receivers as a function of π_s / η_2. An estimated receiver from K blocks is denoted by E-receiver(K). Note the poor performance of P-WL-MMSE and AP-WL-MMSE receivers and the optimal performance of the F-WL-MMSE and the quasi-optimal performance of the AF-WL-MMSE whose convergence is very quick.

VI. REFERENCES

[1] S.M. ALAMOUTI, "A simple transmit diversity technique for wireless communications", *IEEE Journal on Selected areas in Communications*, Vol 16, No8, pp. 1451-1458, Oct. 1998.

[2] M. AUSTIN, "SAIC and synchronised networks for increased GSM capacity", *3G Americas' SAIC Working group*, Sept 2003.

[3] P. CHEVALIER, F. PIPON, "New Insights into optimal widely linear array receivers for the demodulation of BPSK, MSK and GMSK signals corrupted by noncircular interferences – Application to SAIC", *IEEE Trans. Signal Processing*, Vol 54, No3, pp. 870-883, March 2006.

[4] J. KAZEMITABAR, H. JAFARKHANI, "Multiuser interference cancellation and detection for users with more than two transmit antennas", *IEEE Trans. On Communications*, Vol 56, N°4, pp. 574-583, April 2008.

[5] M. KONRAD, W. GERSTACKER, "Interference robust transmission for the downlink of an OFDM-Based mobile communications system", *EURASIP Journal on Wireless Communications and Networking*, Vol. 2008, Article ID 549371, 14 pages, doi: 10.1155/2008/549371.

[6] A. LAMPE, M. BREILING, "Asymptotic analysis of widely linear MMSE multiuser detection- complex versus real modulation", *Proc IEEE Info. Theory Workshop (ITW'2001)*, pp. 55-57, Sept 2001.

[7] H. LI, X. LU, G.B. GIANNAKIS, "Capon multiuser receiver for CDMA systems with space-time coding", *IEEE Trans. Signal Processing*, Vol 50, No5, pp. 1193-1204, May 2002.

[8] J. LI, K.B. LETAIEF, Z. CAO, "Co-channel interference cancellation for space-time coded OFDM systems", *IEEE Trans. On Wireless Communications*, Vol 2, No1, pp. 41-49, Jan. 2003.

[9] R. MEYER, W.H. GERSTACKER, R. SCHOBER, J.B. HUBER, "A single antenna interference cancellation algorithm for increased GSM capacity", *IEEE Trans. On Wireless Communications*, Vol 5, No7, pp. 1616-1621, July 2006.

[10] A.F. NAGUIB, N. SESHADRI, A.R. CALDERBANK, "Applications of space-time block codes and interference suppression for high capacity and high data rate wireless systems", *Proc. 32th Annual Asilomar Conference on Signals, Systems and Computer*, pp. 1803-1810, Pacific Grove, California, Nov. 1998.

[11] A. NAGUIB, N. SESHADRI, A.R. CALDERBANK, "Increasing data rate over wireless channels", *IEEE Signal Processing Magazine*, Vol 17, N°3, pp. 76-92, May 2000.

[12] B. PICINBONO, "Second order complex random vectors and normal distributions", *IEEE Trans. Signal Processing*, Vol 44, N°10, pp. 2637-2640, Oct. 1996.

[13] B. PICINBONO, P. CHEVALIER, "Widely linear estimation with complex data", *IEEE Trans. Signal Processing*, Vol 43, N°8, pp. 2030-2033, Aug. 1995.

[14] A. STAMOULIS, N. AL-DHAHIR, A.R. CALDERBANK, "Further results on interference cancellation and space-time block codes", *Proc. 35th Annual Asilomar Conference on Signals, Systems and Computer*, pp. 257-261, Pacific Grove, California, Nov. 2001.

[15] V. TAROKH, A. NAGUIB, N. SESHADRI, A.R. CALDERBANK, "Combined array processing and space-time coding", *IEEE Trans. Inform. Theory*, Vol 45, N°4, pp. 1121-1128, May 1999.

[16] H. TRIGUI, D.T.M. SLOCK, "Performance bounds for cochannel interference cancellation within the current GSM standard", *Signal Processing*, Elsevier, Vol. 80, pp. 1335-1346, 2000.

[17] A. VAN DEN BOS, "The multivariate complex normal distribution – A generalization", *IEEE Trans. Inform. Theory*, Vol 41, pp. 537-539, Mar. 1995.