How to make quasi-rectilinear signals (MSK, GMSK, OQAM) almost equivalent to rectilinear ones (BPSK, ASK) for widely linear filtering in the presence of CCI

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How to make Quasi-rectilinear Signals (MSK, GMSK, OQAM) almost equivalent to Rectilinear ones (BPSK, ASK) for Widely Linear Filtering in the presence of CCI

Rémi Chauvat*†,Pascal Chevalier*†,Jean-Pierre Delmas‡
*CNAM, CEDRIC Laboratory, 292 rue Saint-Martin, 75003 Paris, France
†Thales-Communications-Security, HTE/AMS/TCP, 4 Av. des Louvresses, 92622 Gennevilliers, France
‡Telecom SudParis, UMR CNRS 5157, 9 rue Charles Fourier, 91011 Evry, France

***Abstract*—Widely linear (WL) receivers are able to fulfill single antenna interference cancellation (SAIC) of one rectilinear (R) or quasi-rectilinear (QR) co-channel interference (CCI), a function which is operational in GSM handsets in particular. However, SAIC requires enhancements for QR signals, for both VAMOS standard, an evolution of GSM/EDGE standard, and FBMC-OQAM networks, which are candidate for 5G mobile networks. In this context, the purpose of this paper is twofold. The first one is to show that, contrary to what is accepted as true in the literature, QR signals are less efficient than R ones for conventional WL filtering in the presence of CCI. From this result and for QR signals, the second purpose is to propose a SAIC/MAIC enhancement, based on WL frequency shifted (FRESH) filtering, making QR signals almost equivalent to R ones.**

I. Introduction

These two last decades, since the pioneer works on the subject [1–4], WL filtering has aroused a great interest for second-order (SO) non-circular signals [5] in many areas. Nevertheless, the subject which has attracted the greatest attention is CCI mitigation in radio communication networks using R or QR modulations. Let us recall that R modulations correspond to mono-dimensional modulations such as ASK or BPSK modulations, whereas QR modulations are complex modulations corresponding, after a simple derotation operation [6], to a complex filtering of a R modulation. Examples of QR modulations are MSK, GMSK or OQAM modulations. One remarkable property of WL filtering is its ability to fulfill SAIC of one R or QR multi-user (MU) CCI, allowing the separation of two users from only one receive antenna [7–9]. The powerfulness of this technology jointly with its low complexity are the reasons why it is operational in most of GSM handsets, generating significant network's capacity gains for the GSM system [9], [10]. Extension of the SAIC technology to a multi-antenna reception is called multiple antenna interference cancellation (MAIC). To further increase the spectral efficiency of speech services in emerging markets such as China, Eastern Europe, Africa or India, the Voice services over Adaptive Multi-user channels on One Slot (VAMOS) technology, has been recently standardized [11]. It allows the transmission of two GSM voice streams on the same TDMA slot at the same frequency through the Orthogonal Sub Channel (OSC) multiple access technique which aims at doubling the number of users served by a cell. The separation, at the handset level, of the two streams, potentially corrupted by co-channel OSC and/or GMSK interference, requires the implementation of enhanced SAIC receivers for QR signals, preliminary introduced in [12–14]. A similar need is also required to mitigate both inter-carrier interference (ICI) and CCI for networks which will use filter bank multi-carrier (FBMC) waveforms coupled with OQAM modulation, which are considered as promising candidates for the 5G mobile networks in particular [15]. First WL filtering based solutions are presented in [16–18].

In this context, the purpose of this paper is twofold. The first one is to show that, contrary to what is implicitly accepted as true in the literature [6–8], [12–14], [16–21], QR signals are less efficient than R ones for conventional WL filtering in the presence of CCI. Starting from this result, the second purpose of this paper is to propose an enhanced generic SAIC/MAIC technique for links using QR modulations, corrupted by QR CCI. This SAIC/MAIC enhancement is based on WL FRESH filtering of the data and makes, for SAIC/MAIC purpose, QR signals almost equivalent to R ones.

To compare QR and R signals for conventional WL filtering and to show the powerfulness of the proposed enhanced SAIC/MAIC technique for QR signals, we adopt a continuous-time (CT) approach, allowing us to remove both the filtering structure constraints imposed by a discrete-time (DT) approach and the potential influence of the sample rate. Besides, we choose a pseudo Maximum Likelihood Sequence Estimation (MLSE) approach much more easy to compute than a MLSE approach. Note that the scarce papers dealing with WL FRESH filtering for demodulation of QR signals correspond to [22–24]. While [22] concerns DS-CDMA systems, [24] considers a particular DT MMSE approach. Moreover, [23] mentions the proposed enhanced SAIC/MAIC technique for interference cancellation in the GSM context but through a DT approach at the symbol rate, which finally reduces to the conventional SAIC/MAIC approach.

II. Models and second order statistics

A. Observation model and SO statistics

We consider an array of N narrow-band antennas receiving the contribution of a signal of interest (SOI), which may be R or QR, and a total noise. The vector of complex amplitudes of the data at the output of these antennas can then be written as

$$\mathbf{x}(t) = \sum_k a_k \mathbf{g}(t - kT) + \mathbf{n}(t). \quad (1)$$

Here, $a_k = b_k$ for R signals whereas $a_k = j^k b_k$ for QR signals, where b_k are real-valued zero-mean i.i.d. r.v., corresponding to the SOI symbols for R signals and directly related to the SOI symbols for QR signals [8], [19], T is the symbol period for R, MSK and GMSK signals and half the symbol period for OQAM signals, $\mathbf{g}(t) = v(t) * \mathbf{h}(t)$ is the impulse response of the SOI global channel, $*$ is the convolution operation, $v(t)$ and $\mathbf{h}(t)$ are the impulse responses of the SOI pulse shaping filter and propagation channel respectively and $\mathbf{n}(t)$ is the zero-mean total noise vector.

The SO statistics of $\mathbf{x}(t)$ are characterized by the two correlation matrices $\mathbf{R}_x(t,\tau)$ and $\mathbf{C}_x(t,\tau)$, defined by

$$\mathbf{R}_x(t,\tau) \triangleq \mathrm{E}[\mathbf{x}(t+\tau/2)\mathbf{x}^H(t-\tau/2)] \quad (2)$$

$$\mathbf{C}_x(t,\tau) \triangleq \mathrm{E}[\mathbf{x}(t+\tau/2)\mathbf{x}^T(t-\tau/2)] \quad (3)$$

where $(.)^T$ and $(.)^H$ mean transpose and conjugate transpose respectively. Assuming that $\mathbf{n}(t)$ is composed of MU CCI, having the same nature (R or QR) and the same symbol period as the SOI, and stationary background noise, it is easy to verify that $\mathbf{R}_x(t,\tau)$ and $\mathbf{C}_x(t,\tau)$ are periodic functions of t with periods equal to T and T respectively for R signals and to T and $2T$ respectively for QR signals. Matrices $\mathbf{R}_x(t,\tau)$ and $\mathbf{C}_x(t,\tau)$ have then Fourier series expansions given by

$$\mathbf{R}_x(t,\tau) = \sum_{\alpha_i} \mathbf{R}_x^{\alpha_i}(\tau) e^{j2\pi\alpha_i t} \quad (4)$$

$$\mathbf{C}_x(t,\tau) = \sum_{\beta_i} \mathbf{C}_x^{\beta_i}(\tau) e^{j2\pi\beta_i t}. \quad (5)$$

Here α_i and β_i are the first and second SO cyclic frequencies of $\mathbf{x}(t)$ such that $\alpha_i = \beta_i = i/T$ ($i \in \mathbb{Z}$) for R signals and $\alpha_i = i/T$ and $\beta_i = (2i+1)/2T$ ($i \in \mathbb{Z}$) for QR signals [25], [26], $\mathbf{R}_x^{\alpha_i}(\tau)$ and $\mathbf{C}_x^{\beta_i}(\tau)$ are the first and second cyclic correlation matrices of $\mathbf{x}(t)$ for the cyclic frequencies α_i and β_i and the delay τ, defined by

$$\mathbf{R}_x^{\alpha_i}(\tau) \triangleq \langle \mathbf{R}_x(t,\tau) e^{-j2\pi\alpha_i t} \rangle_\infty \quad (6)$$

$$\mathbf{C}_x^{\beta_i}(\tau) \triangleq \langle \mathbf{C}_x(t,\tau) e^{-j2\pi\beta_i t} \rangle_\infty \quad (7)$$

where $\langle\cdot\rangle_\infty$ is the temporal mean operation in t over an infinite observation duration.

B. Extended models

For both R and QR signals, conventional linear processing of $\mathbf{x}(t)$ only exploits the information contained in the first ($\alpha = 0$) zero SO cyclic frequency of $\mathbf{x}(t)$.

For R signals, conventional WL processing of $\mathbf{x}(t)$ only exploits the information contained in the first and second $(\alpha, \beta) = (0, 0)$ zero SO cyclic frequencies of $\mathbf{x}(t)$ through the exploitation of the temporal mean of the first correlation matrix of the extended model

$$\widetilde{\mathbf{x}}(t) \triangleq [\mathbf{x}^T(t), \mathbf{x}^H(t)]^T = \sum_k b_k \widetilde{\mathbf{g}}(t - kT) + \widetilde{\mathbf{n}}(t) \quad (8)$$

where $\widetilde{\mathbf{g}}(t) \triangleq [\mathbf{g}^T(t), \mathbf{g}^H(t)]^T$ and $\widetilde{\mathbf{n}}(t) \triangleq [\mathbf{n}^T(t), \mathbf{n}^H(t)]^T$.

For QR signals, as no information is contained in $\beta = 0$, a derotation preprocessing of the data is required before WL filtering. Using (1) for QR signals, the derotated observation vector can be written as

$$\mathbf{x}_d(t) \triangleq j^{-t/T}\mathbf{x}(t) = \sum_k b_k \mathbf{g}_d(t - kT) + \mathbf{n}_d(t) \quad (9)$$

where $\mathbf{g}_d(t) \triangleq j^{-t/T}\mathbf{g}(t)$ and $\mathbf{n}_d(t) \triangleq j^{-t/T}\mathbf{n}(t)$. Expression (9) shows that the derotation operation makes a QR signal looks like a R signal, with a non-zero information for $\beta = 0$. Indeed, it is easy to verify that the two correlation matrices, $\mathbf{R}_{x_d}(t,\tau)$ and $\mathbf{C}_{x_d}(t,\tau)$ of $\mathbf{x}_d(t)$ are such that

$$\mathbf{R}_{x_d}(t,\tau) = j^{-\tau/T}\mathbf{R}_x(t,\tau) \quad (10)$$

$$\mathbf{C}_{x_d}(t,\tau) = j^{-2t/T}\mathbf{C}_x(t,\tau) \triangleq e^{-j2\pi t/2T}\mathbf{C}_x(t,\tau). \quad (11)$$

These expressions show that the first, α_{d_i}, and second, β_{d_i}, SO cyclic frequencies of $\mathbf{x}_d(t)$ are such that $\alpha_{d_i} = \alpha_i$ and $\beta_{d_i} = \beta_i - 1/2T = i/T$, which proves the presence of information at $\beta_{d_0} = 0$. Thus conventional WL processing of QR signals exploits the information contained in $(\alpha_{d_0}, \beta_{d_0}) = (0, 0)$ through the exploitation of the temporal mean of the first correlation matrix of the extended derotated model

$$\widetilde{\mathbf{x}}_d(t) \triangleq [\mathbf{x}_d^T(t), \mathbf{x}_d^H(t)]^T = \sum_k b_k \widetilde{\mathbf{g}}_d(t - kT) + \widetilde{\mathbf{n}}_d(t) \quad (12)$$

where $\widetilde{\mathbf{g}}_d(t) \triangleq [\mathbf{g}_d^T(t), \mathbf{g}_d^H(t)]^T$ and $\widetilde{\mathbf{n}}_d(t) \triangleq [\mathbf{n}_d^T(t), \mathbf{n}_d^H(t)]^T$. Comparing (8) and (12), we deduce that $\widetilde{\mathbf{x}}(t)$ and $\widetilde{\mathbf{x}}_d(t)$ have similar form. This means that similar conventional WL processing may be used for R and QR signals provided that the data vector $\mathbf{x}(t)$, used for R signals, is replaced by $\mathbf{x}_d(t)$ for QR signals.

III. Generic pseudo-MLSE receiver

To compare R and QR signals for conventional WL filtering in the presence of CCI, we need to introduce the chosen receiver, which corresponds to a pseudo-MLSE receiver.

A. Pseudo-MLSE approach

In order to only exploit the information contained in the SO statistics of the data, and for both R and QR signals, the CT MLSE receiver for the detection of the symbols b_k, would assume a Gaussian total noise despite the fact that the CCI are R or QR. Note that the Gaussian assumption would nevertheless be verified in practice for a high number of i.i.d. CCI. Moreover, to exploit the SO cyclostationarity and the SO non-circularity properties of the CCI, the total noise would be assumed to be cyclostationary and non-circular. However, under these assumptions, the CT MLSE receiver, which optimally exploits the CCI SO properties, is very challenging to derive, and even probably impossible to implement. Such a MLSE receiver would optimally exploit the information contained in all the (α_i, β_i) $i \in \mathbb{Z}$ through the probable implementation of an infinite number of time invariant (TI) filters acting on an infinite number of FRESH versions of $\mathbf{x}(t)$ and $\mathbf{x}^*(t)$, where $(.)^*$ means conjugate.

In this context, to overcome the difficulty to compute the CT MLSE receiver, a conventional WL approach consists in only exploiting the non-circularity of the data, i.e. of $\mathbf{x}(t)$ and $\mathbf{x}_d(t)$ for R and QR signals respectively, but not their cyclostationarity. In other words, it consists in computing the CT MLSE receiver from $\mathbf{x}(t)$ or $\mathbf{x}_d(t)$, for R and QR signals respectively, assuming a Gaussian non-circular but stationary total noise $\mathbf{n}(t)$ or $\mathbf{n}_d(t)$. It can be easily verified [27] that this approach is equivalent to compute the CT MLSE receiver from $\widetilde{\mathbf{x}}(t)$ (R signals) or $\widetilde{\mathbf{x}}_d(t)$ (QR signals) in Gaussian circular stationary extended total noise $\widetilde{\mathbf{n}}(t)$ or $\widetilde{\mathbf{n}}_d(t)$ respectively. To approximate the CT MLSE receiver in cyclostationary non-circular total noise, we adopt in the following the previous sub-optimal approach and we call it a CT two-inputs pseudo-MLSE approach. We will then compare in the following the output performance of the two inputs pseudo-MLSE receivers computed from (8) and (12) for R and QR signals corrupted by CCI of the same nature respectively.

B. Generic pseudo-MLSE receiver

We denote by $\widetilde{\mathbf{x}}_F(t)$ and $\widetilde{\mathbf{n}}_F(t)$ the generic extended observation and total noise vectors respectively. These vectors correspond respectively to $\widetilde{\mathbf{x}}(t)$ and $\widetilde{\mathbf{n}}(t)$, defined by (8), for R signals and to $\widetilde{\mathbf{x}}_d(t)$ and $\widetilde{\mathbf{n}}_d(t)$, defined by (12), for QR signals. Note that for conventional linear receivers, which assume Gaussian stationary circular total noise, $\widetilde{\mathbf{x}}_F(t)$ and $\widetilde{\mathbf{n}}_F(t)$ reduce respectively to $\mathbf{x}(t)$ and $\mathbf{n}(t)$, for R signals, and to $\mathbf{x}_d(t)$ and $\mathbf{n}_d(t)$, for QR signals, and the generic pseudo-MLSE receiver reduces to a one input pseudo-MLSE reveiver. Assuming a stationary, circular and Gaussian generic extended total noise $\widetilde{\mathbf{n}}_F(t)$, it is shown in [28] that the sequence $\widehat{\mathbf{b}} \triangleq (\widehat{b}_1, ..., \widehat{b}_K)$ which maximizes its likelihood from $\widetilde{\mathbf{x}}_F(t)$ is the one which minimizes the following criterion[1]:

$$C(\mathbf{b}) = \int [\widetilde{\mathbf{x}}_F(f) - \widetilde{\mathbf{s}}_F(f)]^H [\mathbf{R}^0_{\widetilde{n}_F}(f)]^{-1} [\widetilde{\mathbf{x}}_F(f) - \widetilde{\mathbf{s}}_F(f)] df. \tag{13}$$

Considering only terms that depend on the symbols b_k, the minimization of (13) is equivalent to that of the metric:

$$\Lambda(\mathbf{b}) = \sum_{k=1}^{K} \sum_{k'=1}^{K} b_k b_{k'} r_{k,k'} - 2 \sum_{k=1}^{K} b_k z_F(k) \tag{14}$$

where the real-valued sampled output $z_F(k)$ and $r_{k,k'}$ are defined by

$$z_F(k) = \int \widetilde{\mathbf{g}}^H_F(f) [\mathbf{R}^0_{\widetilde{n}_F}(f)]^{-1} \widetilde{\mathbf{x}}_F(f) e^{j2\pi fkT} df \tag{15}$$

$$r_{k,k'} = \int \widetilde{\mathbf{g}}^H_F(f) [\mathbf{R}^0_{\widetilde{n}_F}(f)]^{-1} \widetilde{\mathbf{g}}_F(f) e^{j2\pi f(k-k')T} df. \tag{16}$$

C. Interpretation of the generic pseudo-MLSE receiver

We deduce from (15) that $z_F(k)$ is the sampled version, at time $t = kT$, of the output of the TI filter whose frequency response is

$$\widetilde{\mathbf{w}}^H_F(f) \triangleq \left([\mathbf{R}^0_{\widetilde{n}_F}(f)]^{-1} \widetilde{\mathbf{g}}_F(f) \right)^H \tag{17}$$

and whose input is $\widetilde{\mathbf{x}}_F(t)$. The structure of the generic M inputs pseudo-MLSE receiver ($M = 1, 2$) is then depicted at Fig. 1. It is composed of the TI WL filter (17), which reduces to a linear filter for conventional receivers, followed by a sampling at the symbol rate and a decision box implementing the Viterbi algorithm, since $r^*_{k,k'} = r_{k',k}$.

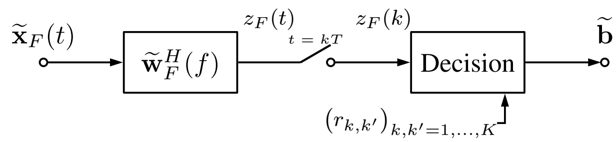

Fig. 1. Structure of the M inputs pseudo-MLSE receiver

D. SINR at the output of the generic pseudo-MLSE receiver

For real-valued symbols b_k, the symbol error rate (SER) at the output of the generic M inputs ($M = 1, 2$) pseudo-MLSE receiver is directly linked to the signal to interference plus noise ratio (SINR) on the current symbol before decision, i.e. at the output $z_F(n)$ [29, Sec 10.1.4], while the inter-symbol interference is processed by the decision box. For this reason, we compute the general expression of the output SINR hereafter and we will analyze its variations for both R and QR signals in particular situations in section IV. As $\widetilde{\mathbf{n}}_F(t)$ is cyclostationary and non-circular, the filter (17) does not maximizes the output SINR and can only be considered as a generic M inputs pseudo-matched filter. It is easy to verify from (1), (8), (9), (12), (15) and (16) that $z_F(n)$ can be written as

$$z_F(n) = b_n r_{n,n} + \sum_{k \neq n} b_k \mathrm{Re}[r_{n,k}] + z_{n,F}(n) \tag{18}$$

where the real-valued sample $z_{n,F}(n)$ is defined by (15) for $k = n$ with $\widetilde{\mathbf{n}}_F(f)$ instead of $\widetilde{\mathbf{x}}_F(f)$. The SINR on the current symbol is then defined by

$$\mathrm{SINR}_F \triangleq \pi_b r^2_{n,n} / \mathrm{E}\left[z^2_{n,F}(n)\right], \tag{19}$$

where $\pi_b \triangleq \mathrm{E}[b^2_n]$.

IV. SINR ANALYSIS FOR ONE CCI

A. Total noise model

To compare the extended models (8) and (12) for R and QR signals respectively, we assume that the total noise is composed of one MU CCI, having the same nature (R or QR) as the SOI, and a background noise. Under these assumptions, $\mathbf{n}(t)$ can be written as

$$\mathbf{n}(t) = \sum_k f_k \mathbf{g}_I(t - kT) + \mathbf{u}(t) \tag{20}$$

Here, $f_k = e_k$ for R signals whereas $f_k = j^k e_k$ for QR signals, where e_k are real-valued zero-mean i.i.d. r.v., corresponding to the interference symbols for R signals and directly related to the interference symbols for QR signals, $\mathbf{g}_I(t) = v(t) * \mathbf{h}_I(t)$, $\mathbf{h}_I(t)$ is the impulse response of the propagation channel of the CCI and $\mathbf{u}(t)$ is the background noise vector, assumed stationary, temporally and spatially white. To simplify the following analysis, we assume a raised cosine pulse shaping filter $v(t)$ with a roll-off γ and deterministic propagation channels with no delay spread such that

$$\mathbf{h}(t) = \mu\delta(t)\mathbf{h} \quad \text{and} \quad \mathbf{h}_I(t) = \mu_I \delta(t - \tau_I)\mathbf{h}_I \tag{21}$$

[1] All Fourier transforms of vectors $\mathbf{x}$ and matrices $\mathbf{X}$ use the same notation where t and τ is simply replaced by f, e.g., here $\mathbf{R}^0_{\widetilde{\mathbf{n}}_F}(f)$ is the Fourier transform of (6), where α_i and $\mathbf{x}(t)$ are replaced by 0 and $\widetilde{\mathbf{n}}_F(t)$ respectively, whereas $\widetilde{\mathbf{s}}_F(f) \triangleq \sum_{k=1}^{K} b_k \widetilde{\mathbf{g}}_F(f) e^{-j2\pi fkT}$, where $\widetilde{\mathbf{g}}_F(f)$ corresponds to $\widetilde{\mathbf{g}}(f)$ and $\widetilde{\mathbf{g}}_d(f)$ for R and QR signals respectively.

Here, μ and μ_I control the amplitude of the SOI and CCI, $\delta(t)$ is the Dirac pulse, τ_I is the delay of the CCI with respect to the SOI whereas $\mathbf{h}$ and $\mathbf{h}_I$, such that $\mathbf{h}^H\mathbf{h} = \mathbf{h}_I^H\mathbf{h}_I = N$, are the channel vectors of the SOI and CCI.

B. SINR computations and analysis for a zero roll-off

Under the previous assumptions, analytical interpretable expressions of the SINRs (19) are only possible for a zero roll-off. In this case, we denote by $\pi_s \triangleq \mu^2\pi_b$, $\pi_I \triangleq \mu_I^2\pi_e$ and η_2 the power of the SOI, the CCI and the background noise per antenna at the output of the pulse shaping matched filter, $\pi_e \triangleq \mathrm{E}[e_n^2]$, $\varepsilon_s \triangleq \pi_s\mathbf{h}^H\mathbf{h}/\eta_2$ and $\varepsilon_I \triangleq \pi_I\mathbf{h}_I^H\mathbf{h}_I/\eta_2$. Moreover, assuming $N = 1$ and a strong CCI ($\varepsilon_I \gg 1$) for models (8) and (12) and denoting by SINR_{R_M} and SINR_{QR_M} the SINR at the output of the M input pseudo-MLSE receiver for R and QR signals respectively, we obtain after tedious computations

$$\mathrm{SINR}_{R_1} = \frac{2\varepsilon_s}{1 + 2\varepsilon_I\cos^2(\phi_{Is})} \tag{22}$$

$$\mathrm{SINR}_{QR_1} = \frac{2\varepsilon_s}{1 + \varepsilon_I\left[1 - \cos\left(\frac{\pi\tau_I}{T}\right) + 2\cos\left(\frac{\pi\tau_I}{T}\right)\cos^2\left(\phi_{Is}\right)\right]} \tag{23}$$

$$\mathrm{SINR}_{R_2} \approx 2\varepsilon_s[1 - \cos^2(\phi_{Is})] \quad \phi_{Is} \neq k\pi \tag{24}$$

$$\mathrm{SINR}_{R_2} \approx \frac{2\varepsilon_s}{1 + 2\varepsilon_I} \quad \phi_{Is} = k\pi \tag{25}$$

$$\mathrm{SINR}_{QR_2} \approx 2\varepsilon_s\left[1 - \frac{1 + \cos^2\left(\phi_{Is} + \frac{\pi\tau_I}{2T}\right)}{2}\right]; \Psi_{Is} \neq k\pi \tag{26}$$

$$\mathrm{SINR}_{QR_2} \approx \frac{\varepsilon_s}{\varepsilon_I}\frac{9}{2[3 + 2\cos(4\phi_{Is})]}; \quad \Psi_{Is} = k\pi \tag{27}$$

where $\phi_{Is} \triangleq Arg(\mathbf{h}_I^H\mathbf{h})$ is the phase difference between the SOI and the CCI and $\Psi_{Is} \triangleq \phi_{Is} + \pi\tau_I/2T$.

A receiver performs SAIC as $\varepsilon_I \to \infty$, if the associated SINR does not converge toward zero. We deduce from (22) and (23) that the conventional receivers perform SAIC very scarcely, only when $\phi_{Is} = (2k+1)\pi/2$, for R signals, and when $(\tau_I/T, \phi_{Is}) = (2k_1, (2k_2-1)\pi/2)$ or $(2k_1+1, k_2\pi)$ for QR signals, where k, k_1 ans k_2 are positive or negative integer. However (24) and (26) show that the two-inputs pseudo-MLSE receivers perform SAIC as long as $\phi_{Is} \neq k\pi$, for R signals, and $\Psi_{Is} \neq k\pi$, for QR signals, enlightening the great interest of the WL filtering (17) in both cases. However, despite similar processing (17) and similar models (8) and (12) for R and QR signals, the output SINRs (24) and (26) correspond to different expressions. This proves the non equivalence of R and derotated QR signals for WL filtering in the presence of CCI, result which does not seem to be known by researchers. In particular, for $\gamma = 0$, while (24) only depends on $2\varepsilon_s$, the maximum output SINR obtained without interference, and ϕ_{Is}, (26) depends on $2\varepsilon_s$, ϕ_{Is} and τ_I/T.

To compare these results for $\gamma = 0$ and $\varepsilon_I \gg 1$ from a statistical perspective, we now assume that $\varepsilon_I \to \infty$ and ϕ_{Is} and $\pi\tau_I/2T$ are independent r.v. uniformly distributed on $[0; 2\pi]$. Under these assumptions, we deduce easily from (24) and (26) the expected value of the output SINRs respectively given by

$$\mathrm{E}[\mathrm{SINR}_{R_2}] \approx \varepsilon_s \tag{28}$$

$$\mathrm{E}[\mathrm{SINR}_{QR_2}] \approx \varepsilon_s/2 \tag{29}$$

We observe that $\mathrm{E}[\mathrm{SINR}_{QR_2}] \approx \mathrm{E}[\mathrm{SINR}_{R_2}]/2$, proving that QR signals are less efficient than R ones for WL filtering in the presence of CCI.

C. SINR computations and analysis for arbitrary roll-off

To complete the previous results for arbitrary values of γ, we still assume that ϕ_{Is} and $\pi\tau_I/2T$ are independent r.v. uniformly distributed on $[0; 2\pi]$. Under these assumptions, choosing $\varepsilon_s = 10$ dB and $\varepsilon_I = 20$ dB, Fig. 2 shows, for R and QR signals, for $M = 1, 2$ and $\gamma = 0, 0.5$, $p[(\mathrm{SINR}_F/2\varepsilon_s)_{dB} \geq x \text{ dB}] \triangleq p_F(x)$ as a function of x (dB).

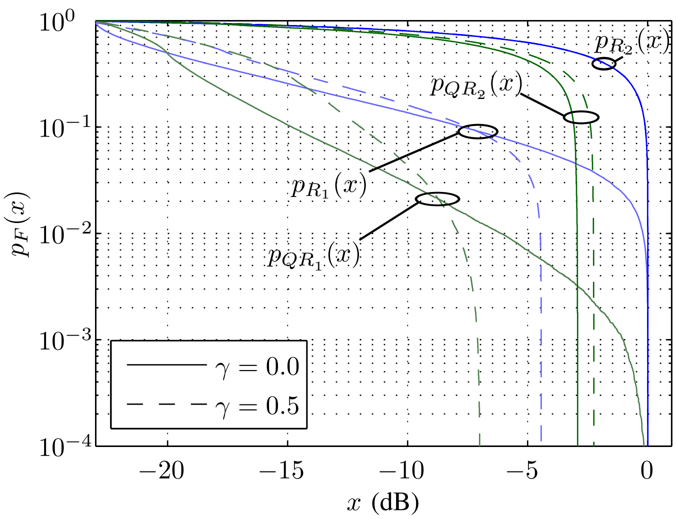

Fig. 2. $p_F(x)$ as a function of x, $N = 1$, $\varepsilon_s = 10$ dB, $\varepsilon_I = 20$ dB.

Note, for both R and QR signals, poor performance whatever γ for $M = 1$, i.e. for the conventional processing. Note, for $M = 2$ (SAIC processing), increasing and constant performance with γ for QR and R signals respectively, and the best performance of (8) with respect to (12) whatever γ. This confirms the lowest efficiency of QR signals with respect to R ones for SAIC for arbitrary values of γ. Note in particular, for $\gamma = 0.5$ and $x = -3$ dB, that $p_{QR_1}(x) = p_{R_1}(x) = 0\%$, $p_{QR_2}(x) = 26\%$ and $p_{R_2}(x) = 50\%$, proving the much better performance of (8) with respect to (12).

V. Enhanced SAIC/MAIC technique for QR signals

We describe in this section the reasons why QR signals are less efficient than R ones for WL filtering in the presence of CCI and we propose, for QR signals, a WL filtering enhancement to make them almost equivalent to R signals.

A. The lower efficiency of QR signals

The lower efficiency of QR signals with respect to R ones for WL filtering in the presence of CCI is directly related to the different SO non-circularity and cyclostationnary properties of QR and R signals. Indeed, while for R signals, the main information about their non-circularity is contained in the SO cyclic frequency $\beta = 0$ whatever the filter $v(t)$, for QR signals, it is symmetrically contained in $(\beta_0, \beta_{-1}) =$

$(1/2T, -1/2T)$, as illustrated in [25], [26], or equivalently in $(\beta_{d_0}, \beta_{d_{-1}}) = (0, -1/T)$ for derotated QR signals. As the pseudo-matched filter (17) applied to the model (12) only exploits the information contained in $(\alpha_{d_0}, \beta_{d_0}) = (0, 0)$, or $(\alpha_0, \beta_0) = (0, 1/2T)$, only a part of the main non-circularity information of QR signals is exploited by this receiver. It is not the case for R signals for which the pseudo-MLSE receiver applied to the model (8) exploits all the main non-circularity information of R signals.

B. Three inputs FRESH model

To overcome the previous limitation for QR signals, it is necessary to implement a WL filtering which takes into account all the main non-circularity information of QR signals. Such a result can be obtained by implementing the pseudo-MLSE receiver from the three inputs FRESH model defined by

$$\begin{aligned}\mathbf{x}_{F_3}(t) &\triangleq [\mathbf{x}^T(t), e^{j2\pi t/2T}\mathbf{x}^H(t), e^{-j2\pi t/2T}\mathbf{x}^H(t)]^T \\ &= j^{t/T}[\widetilde{\mathbf{x}}_d^T(t), e^{-j3\pi t/2T}\mathbf{x}_d^H(t)]^T \triangleq j^{t/T}\mathbf{x}_{d_{F_3}}(t) \\ &= \sum_k j^k b_k \mathbf{g}_{F_3}(t-kT) + \mathbf{n}_{F_3}(t) \end{aligned} \quad (30)$$

where $\mathbf{n}_{F_3}(t)$ corresponds to $\mathbf{x}_{F_3}(t)$ with $\mathbf{n}(t)$ instead of $\mathbf{x}(t)$ and $\mathbf{g}_{F_3}(t) \triangleq [\mathbf{g}^T(t), e^{j2\pi t/2T}\mathbf{g}^H(t), e^{-j2\pi t/2T}\mathbf{g}^H(t)]^T$. It is straightforward to verify that the temporal mean of the first correlation matrices of $\mathbf{x}_{F_3}(t)$ and $\mathbf{x}_{d_{F_3}}(t)$ exploits the information contained in $(\alpha_0, \alpha_{-1}, \alpha_1, \beta_0, \beta_{-1}) = (0, -1/T, 1/T, 1/2T, -1/2T)$, which allows us to exploit almost exhaustively both the cyclostationarity and the non-circularity of QR signals. Note that TI linear processing of $\mathbf{x}_{F_3}(t)$ or $\mathbf{x}_{d_{F_3}}(t)$ becomes now a time variant (TV) WL filtering of both $\mathbf{x}(t)$ and $\mathbf{x}_d(t)$, called here three inputs WL FRESH filtering of $\mathbf{x}(t)$ or $\mathbf{x}_d(t)$.

C. Three inputs pseudo-MLSE receiver

Following the developments of section III-B, the three inputs pseudo-MLSE receiver still generates the sequence $\widehat{\mathbf{b}} \triangleq (\widehat{b}_1, ..., \widehat{b}_K)$ which maximizes (14) but where $z_F(k)$, now denoted by $z_{F_3}(k)$, is defined by $z_{F_3}(k) \triangleq \mathrm{Re}[j^{-k}y_{F_3}(k)]$, where $y_{F_3}(k)$ is given by

$$y_{F_3}(k) = \int \mathbf{g}_{F_3}^H(f)[\mathbf{R}^0_{n_{F_3}}(f)]^{-1}\mathbf{x}_{F_3}(f)e^{j2\pi fkT}df \quad (31)$$

Here, $\mathbf{R}^0_{n_{F_3}}(f)$ is the Fourier transform of (6), where α_i and $\mathbf{x}(t)$ are replaced by 0 and $\mathbf{n}_{F_3}(t)$ respectively, while $r_{k,k'}$ is now defined by

$$r_{k,k'} = j^{k'-k}\int \mathbf{g}_{F_3}^H(f)[\mathbf{R}^0_{n_{F_3}}(f)]^{-1}\mathbf{g}_{F_3}(f)e^{j2\pi f(k-k')T}df \quad (32)$$

Noting that $y_{F_3}(k)$ is the sampled version, at time $t = kT$, of the output of the TI filter whose frequency response is

$$\mathbf{w}_{F_3}^H(f) \triangleq \left([\mathbf{R}^0_{n_{F_3}}(f)]^{-1}\mathbf{g}_{F_3}(f)\right)^H \quad (33)$$

and whose input is $\mathbf{x}_{F_3}(t)$, the structure of the 3 inputs pseudo-MLSE receiver is then depicted at Fig. 3. It is composed of the TI WL filter (33), followed by a sampling at the symbol rate, a derotation operation, a real part capture and a decision box implementing the Viterbi algorithm since $r^*_{k,k'} = r_{k',k}$.

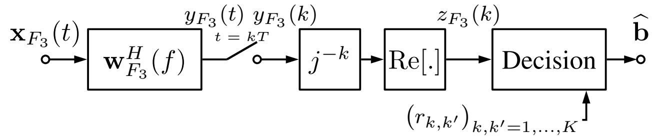

Fig. 3. Structure of the 3 inputs pseudo-MLSE receiver

D. SINR at the output of the three inputs pseudo-MLSE receiver

It is straighforward to show that the SINR in the output $z_{F_3}(n)$ is given by

$$\mathrm{SINR}_{QR_3} \triangleq \pi_b r^2_{n,n}/\mathrm{E}[(\mathrm{Re}[j^{-n}y_{n,F_3}(n)])^2] \quad (34)$$

where $y_{n,F_3}(n)$ is defined similarly as $y_{F_3}(n)$ but with $\mathbf{n}_{F_3}(f)$ instead of $\mathbf{x}_{F_3}(f)$.

To illustrate (34) and the behaviour of the three inputs pseudo-MLSE receiver, we consider again the total noise model (20) with (21) and we assume that ϕ_{Is} and $\pi\tau_I/2T$ are still independent r.v. uniformly distributed on $[0, 2\pi]$. Under these assumptions, choosing $\varepsilon_s = 10$ dB and $\varepsilon_I = 20$ dB, Fig. 4 shows, for R and QR signals, for $M = 2, 3$, and $\gamma = 0, 0.5, 1$, $p[(\mathrm{SINR}_F/2\varepsilon_s)_{dB} \geq x \text{ dB}] \triangleq p_F(x)$ as a function of x (dB), where F is used for R_2, QR_2 or QR_3. Note, for QR signals, increasing performance with γ for $M = 2, 3$ and the best performance of (30) with respect to (12) whatever γ. Note in particular, for $\gamma = 0.5$ and $x = -3$ dB, that $p_{QR_2}(x) = 26\%$, $p_{R_2}(x) = 50\%$ and $p_{QR_3}(x) = 63\%$, proving the much better performance of (30) with respect to (12) and the even better performance of (30) with respect to (8) for $x = -3$ dB. Note finally globally the almost comparable performance of the 2 and 3 inputs pseudo-MLSE receiver for R and QR signals respectively.

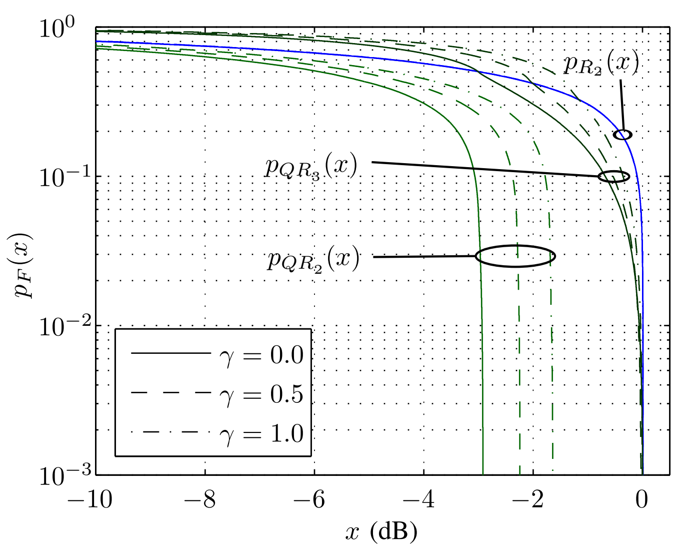

Fig. 4. $p_F(x)$ as a function of x, $N = 1$, $\varepsilon_s = 10$ dB, $\varepsilon_I = 20$ dB.

VI. Conclusion

We have shown in this paper, through a suboptimal but efficient CT pseudo-MLSE approach, that contrary to what is accepted as true in the literature, QR signals are less efficient than R signals for conventional WL filtering in the presence of CCI. This results is directly linked to the different non-circularity and cyclostationary properties of these signals. We have then proposed, for QR signals, a WL filtering enhancement to make them almost equivalent to R signals. Other approaches (discrete time, MMSE...) will be investigated elsewhere.

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