EURASIP

NEWS LETTER

ISSN 1687-1421, Volume 20, Number 1, March 2009

European Association for Signal Processing

Newsletter, Volume 20, Number 1, March 2009

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President's Message

This is the first EURASIP Newsletter of 2009. With this year we are starting a new two-year term of the elected EURASIP Administrative Committee (AdCom) with (alphabetical order) Fulvio Gini (Italy), Marc Moonen (Belgium), Ana Perez (Spain), Beatrice Pesquet-Popescu (France), Markus Rupp (Austria), Blent Sankur (Turkey), Abdelhak Zoubir (Germany), truly a European Society. I deeply thank Paulo Lobato Correia (Portugal) for taking over for so many years the treasurer post of EURASIP and Sergios Theodoridis (Greece) who leaves us as well after being president and past president of EURASIP.

With Marc Moonen, after a splendid two year term as EURASIPs president, taking on the position of the past president, I take on the position of current president for the coming two years. I served over the past four years as event chairman responsible for EURASIP sponsored and cooperated workshops and conferences. This activity allowed me to get to know many researchers throughout Europe which I enjoyed collaborating with very much and at the same time learned the ropes of EURASIP.

During the past year a lot of new movements started in EURASIP to improve further community services. After establishing reduced fees for our members when publishing in our open access journals and reduced conference fees for our cooperated workshops, we were offering free book downloads form Hindawi and discount offers from Elsevier. We relaunched the EURASIP PhD library, now with many hundreds of recent European PhDs. It is a free service and I invite everyone to download their PhD thesis up there for free downloads. As our open library with a collection of past EUSIPCO papers is growing we finally managed to scan all old proceedings that were only available on paper and we offer them electronically now. Please check under www.eurasip.org. As we made it a policy that our supported conferences either have their papers on our EURASIP Open Library or on IEEE Xplore, the amount of papers in our library keeps growing.

A second strong movement is the appointment of more than 50 European Liaisons that help distributing the EUSIPCO movement directly into the heart of our community. With more local people actively involved in our signal processing community we hope to address the needs of our members even better than in the past. Note that all work for EURASIP is a voluntary work and we intend to keep it this way since we can wholeheartedly claim that none of your membership fee goes into personal pockets but is returned at 100% to our community.

A new upcoming addition of our service is to establish the EURASIP JOP POST that allows members to offer job opportunities in signal processing. We believe that such service may become very useful in the future and in particular at those times when it is hard to find jobs for qualified people.

In spite of all changes—some already made and some expected—we intend to keep what we are, EURASIP, a friendly and open society for Signal Processing.

I wish all EURASIP members a happy and successful 2009.

Markus Rupp
President

EURASIP Message Liaisons

This message is to welcome the Local Liaisons and to announce the Local Liaison Program that has been created by EURASIP with the aim of coordinating and stimulating local activities, such as seminar days and short courses, which are funded by EURASIP. In addition to supporting local activities, a second goal of Local Liaison Officers is to contribute to promoting EURASIP at the national level, stimulating participation in EURASIP's conferences and workshops, raising awareness of the services offered by EURASIP, and helping membership development so that many people can associate each other with us. The third goal of the program is to increase the speed of response to problems. The Liaison understands local member's needs and workflows, and can frequently fix things.

Liaisons are appointed by EURASIP Adcom, and are chosen to best meet the profile of active people within the Signal Processing community. It is the responsibility of any Liaison who feels that insufficient time is being allocated to perform necessary Liaison work to notify the Adcom. For more information about the Local Liaison Program and to find out who your liaison is on your country, we encourage you to visit our website at www.eurasip.com/.

Ana Perez-Neira
Membership Development

Award Winning Papers by EURASIP

EURASIP is presently running ten scientific Journals (three with Elsevier and seven with Hindawi) and concomitantly we give awards to distinguish excellent papers. The award process is run by a separate committee for each Journal, appointed by the AdCom of EURASIP. Depending on the submission numbers, awards are given annually or biannually. EURASIP has witnessed over the years the truly excellent quality of papers submitted to our Journals, and we would like to take the opportunity to celebrate those award winning papers for our members.

In addition to the Journals papers award, each year during EURASIP's conference, EUSIPCO, student papers are awarded. We therefore present two of the award winning papers, namely, the best student award paper in EUSIPCO 2008 at Lausanne entitled "Total Variation Denoising using Posterior Expectation" by Cécile Louchet and Lionel Moisan and the award winning paper of EURASIPs Journal on Advances in Signal Processing entitled "Wideband Impulse Modulation and Receiver Algorithms for Multiuser Power Line Communications" by Andrea M. Tonello.

Markus Rupp
President

TOTAL VARIATION DENOISING USING POSTERIOR EXPECTATION

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ABSTRACT

Total Variation image denoising, generally formulated in a variational setting, can be seen as a Maximum A Posteriori (MAP) Bayesian estimate relying on a simple explicit image prior. In this formulation, the denoised image is the most likely image of the posterior distribution, which favors regularity and produces staircasing artifacts: in regions where smooth-varying intensities would be expected, constant zones appear separated by artificial boundaries. In this paper, we propose to use the Least Square Error (LSE) criterion instead of the MAP. This leads to a new denoising method called TV-LSE, that produces more realistic images by computing the expectation of the posterior distribution. We describe a Monte-Carlo Markov Chain algorithm based on Metropolis scheme, and provide an efficient convergence criterion. We also discuss the properties of TV-LSE, and show in particular that it does not suffer from the staircasing effect.

1. INTRODUCTION

Image denoising based on Total Variation (TV) was first proposed by Rudin, Osher and Fatemi (ROF) in 1992 [11]. Since then, the TV criterion was found to be very efficient in many other image processing tasks, including deblurring, interpolation, spectrum extrapolation, inpainting, decompression, etc. (see, e.g., [1, 8]). One reason for this is that the TV functional enforces a certain notion of regularity that is well suited to images: it puts a strong penalization on oscillations and random fluctuations, but allows discontinuities at the same time. This is an interesting property, because true images generally present discontinuities in the intensity map that are caused by occluding parts in the scene. However, because the ROF method is based on TV minimization, it tends to produce denoised images that present unnatural local configurations that permit to achieve a small overall TV. This is known as the *staircasing effect* [2, 10]: in ROF denoised images, one may often observe constant regions delimited by artificial discontinuities, as in Fig. 3 (middle row, left image).

As shown in [9], the staircasing effect of TV denoising is due to the fact that the total variation is not differentiable (and, as proven in [4], to the fact that noisy images are almost everywhere discontinuous). Smooth approximations and variants of the total variation functional [2, 5, 6] manage to avoid the staircasing effect, but they lose the nice geometrical properties of the total variation, in particular the co-area formula that connects the total variation measure with the image geometry via the level-set decomposition. In [3], a solution to the staircasing effect is proposed in the case of neighborhood filters, but it does not apply for variational formulations like TV denoising.

In this paper, we propose to use the TV criterion in a different framework, in order to avoid the staircasing effect while keeping the efficiency of the TV measure. In Section 2, we recall the Bayesian MAP interpretation of ROF denoising, and introduce a new denoising filter called TV-LSE, defined as the image estimate achieving the least square error Bayesian risk (like, e.g., the Wiener filter). A Monte-Carlo Markov Chain (MCMC) Metropolis sampler is then proposed in Section 3 to compute the posterior expectation required by this new filter, and a convergence criterion is given and analyzed. In Section 4, we discuss the properties of TV-LSE denoising and in particular its difference with TV-MAP (ROF) denoising. We show that unlike the latter, TV-LSE denoising does not suffer from the staircasing effect, and produce more realistic images while keeping good denoising efficiency.

2. BAYESIAN FORMULATION OF TV DENOISING

Let $u : \Omega \to \mathbb{R}$ be a discrete gray-level image defined on a rectangular domain $\Omega \subset \mathbb{Z}^2$. The discrete Total Variation of u is defined by

$$TV(u) = \sum_{(x,y)\in\Omega} |Du(x,y)|, \tag{1}$$

where $|Du(x,y)|$ is a discrete approximation of the gradient norm of u in (x,y). In this paper, we shall use the usual Euclidean norm in $\mathbb{R}^2$ and the simplest possible approximation of the gradient vector, given by

$$Du(x,y) = \begin{pmatrix} u(x+1,y) - u(x,y) \\ u(x,y+1) - u(x,y) \end{pmatrix}.$$

(with the convention that differences involving pixels outside Ω are zero). Given a (noisy) image u_0, the ROF method proposes to compute the unique image u that minimizes

$$E_\lambda(u) = \|u - u_0\|^2 + \lambda TV(u), \tag{2}$$

where $\|\cdot\|$ is the classical L^2 norm on images and λ is an hyperparameter that controls the level of denoising. This energy-minimization formulation can be translated into a Bayesian framework: let us consider, for $\beta > 0$, the prior density function

$$p_\beta(u) = \frac{1}{Z_\beta} e^{-\beta TV(u)}, \quad \text{where} \quad Z_\beta = \int_{\mathscr{E}_0} e^{-\beta TV(u)}\, du$$

$$\text{and} \quad \forall \mu \in \mathbb{R}, \quad \mathscr{E}_\mu = \left\{ u \in \mathbb{R}^\Omega,\ \sum_{\mathbf{x}\in\Omega} u(\mathbf{x}) = \mu|\Omega| \right\}.$$

The function p_β is a probability density function on each set $\mathscr{E}_\mu$, that can be used as a Bayesian prior to estimate the best

denoised image. If we assume that the noise is additive and Gaussian, i.e., that $u_0 = u + N$ where N is a Gaussian white noise with zero mean and variance σ^2, then from Bayes formula we can derive the posterior density

$$p(u|u_0) = \frac{p(u_0|u)p_\beta(u)}{p(u_0)} = \frac{1}{Z}\exp\left(-\frac{E_\lambda(u)}{2\sigma^2}\right), \quad (3)$$

where $\lambda = 2\beta\sigma^2$ and Z is a normalizing factor, depending on u_0 and λ, ensuring that $u \mapsto p(u|u_0)$ is a probability density function on $\mathbb{R}^\Omega$. Hence, the variational formulation ($\arg\min_u E_\lambda(u)$) is equivalent to the Bayesian Maximum A Posteriori (MAP) formulation

$$\hat{u}_{MAP} = \arg\max_u p(u|u_0), \quad (4)$$

which means that the ROF denoising filter simply selects the most likely image u according to the posterior distribution $p(u|u_0)$.

In a certain sense, the most complete denoising information consists in the whole posterior density function itself. However, one generally wants to build from this density a "best estimate" of the true image, according to some criterion. The MAP estimator is the one that minimizes Bayes risk when the cost function is a Dirac delta localized on the true solution. In a sense, it does not represent very well the posterior density function, because it only sees its maximum point (as we can see from (3), all posterior distributions that share the same value of λ yield the same MAP estimator, independently of their "spread" σ). Since $\hat{u}_{MAP}$ minimizes the energy $E_\lambda(u)$, it tends to present some exceptional structures that have a very small contribution to the energy, in particular "flat zones", that is, regions with uniform intensity, causing the well-known *staircasing effect*.

Instead of the hit-or-miss risk function leading to the MAP estimate, we propose to use the Least Square Error (LSE) criterion, that consists in finding the estimate $\hat{u}(u_0)$ that minimizes

$$\mathbb{E}_{u,u_0}\left(\|u-\hat{u}(u_0)\|^2\right) = \int_{\mathbb{R}^\Omega}\int_{\mathcal{E}_\mu}\|u-\hat{u}(u_0)\|^2 p(u,u_0)\,du\,du_0.$$

This minimum is attained by the posterior expectation (conditional mean), that is for

$$\hat{u}_{LSE} := \mathbb{E}(u|u_0) = \int_{u\in\mathbb{R}^\Omega} u p(u|u_0)\,du. \quad (5)$$

Thanks to (3), this can be rewritten

$$\hat{u}_{LSE} = \frac{\displaystyle\int_{\mathbb{R}^\Omega}\exp\left(-\frac{E_\lambda(u)}{2\sigma^2}\right)\cdot u\,du}{\displaystyle\int_{\mathbb{R}^\Omega}\exp\left(-\frac{E_\lambda(u)}{2\sigma^2}\right)du}. \quad (6)$$

3. MCMC ALGORITHM FOR TV-LSE DENOISING

3.1 Principle

TV-LSE denoising requires to evaluate the ratio of integrals arising in (6), each integral concerning thousands of variables (the dimension is the number of pixels). For such high-dimension integrals, only Monte-Carlo methods can be considered. Here we propose to use a Monte-Carlo Markov Chain (MCMC) following Metropolis scheme. Given a positive parameter $\alpha > 0$, let us consider a random chain of images $(\Upsilon_n)_{n\geq 0}$ consisting in an initial (random or deterministic) image Υ_0 and the transition defined by

$$\Upsilon_{n+1} = \begin{cases} \Upsilon_n + \alpha\Delta_n\delta_{\mathbf{X}_n} & \text{if} \quad R_n \geq Z_n, \\ \Upsilon_n & \text{else,} \end{cases}$$

where the random variables $(\Delta_n)_{n\geq 0}$, $(\mathbf{X}_n)_{n\geq 0}$ and $(Z_n)_{n\geq 0}$ are all independent, with $\Delta_n \sim U([-1,1])$, $\mathbf{X}_n \sim U(\Omega)$, $Z_n \sim U([0,1])$, and

$$R_n = \exp\left(-\frac{E_\lambda\left(\Upsilon_n + \alpha\Delta_n\delta_{\mathbf{X}_n}\right) - E_\lambda(\Upsilon_n)}{2\sigma^2}\right)$$

(note that if $E_\lambda\left(\Upsilon_n + \alpha\Delta_n\delta_{\mathbf{X}_n}\right) \leq E_\lambda(\Upsilon_n)$, then $R_n \geq 1$ so that $R_n \geq Z_n$ almost surely). Notice that two successive images Υ_n and Υ_{n+1} of the chain differ by one pixel at most, while Υ_n and $\Upsilon_{n+|\Omega|}$ are much less correlated. This is why we consider in the following the subsampled chain

$$U_n = \Upsilon_{|\Omega|n},$$

even if the results of this section remain true for (Υ_n). U_n is a Metropolis sampler for the posterior distribution, so it converges in law towards this distribution. This provides a way to estimate the TV-LSE denoising filter, as shown by the following Theorem.

Theorem 1 *For any $\alpha > 0$ and any distribution of U_0, we have almost surely*

$$\frac{1}{n}\sum_{k=1}^{n} U_k \quad \xrightarrow[n\to+\infty]{} \quad \hat{u}_{LSE}.$$

Proof — To simplify, the proof is given here in the case of a countable image space. Let l be a positive real number (quantization step), we assume that Δ_n follows the uniform distribution on $\{k/l,\ -l \leq k \leq l\}$, so that the discrete image space (state space) is $E = (\alpha\mathbb{Z}/l)^\Omega$. The classical ergodic Theorem on Markov chains [7] states that if the Markov chain is irreducible, has a stationary distribution π, and $h : E \to E$ satisfies $\int_E |h(u)|\,d\pi(u) < \infty$, then

$$\frac{1}{n}\sum_{k=1}^{n} h(U_k) \xrightarrow[n\to+\infty]{} \int_E h(u)d\pi(u) \text{ a.s. and in } L^1.$$

1) The chain (U_n) is irreducible because if u and u' are two images of E, then $\mathbb{P}(U_n = u'|U_0 = u) > 0$ for all $n \geq \|u'-u\|_\infty/\alpha$, so that u and u' communicate.

2) Let us write $\pi(u) = \frac{1}{Z}\exp(-\frac{E_\lambda(u)}{2\sigma^2})$ the posterior distribution. To prove that π is stationary for the subsampled chain (U_n), it is sufficient to prove that it is stationary for (Υ_n). The transition kernel P of Υ_n can be decomposed into $P_{u,u'} = q(u,u')e^{-\frac{(E_\lambda(u')-E_\lambda(u))_+}{2\sigma^2}}$, where $(x)_+ = \max(0,x)$ is the positive part of x, and

$$q(u,u') = \frac{1}{|\Omega|}\sum_{x\in\Omega}\frac{1}{2l+1}\mathbf{1}_{|u'(x)-u(x)|\leq\alpha, u'(y)=u(y)\forall y\neq x}(u,u')$$

is the instrumental distribution. If $\pi(u) \geq \pi(u')$, then $(E_\lambda(u') - E_\lambda(u))_+$ is null, and $\pi(u)P_{u,u'} = \pi(u)q(u,u')$ holds. But q is symmetric, so that $\pi(u)P_{u,u'} = \pi(u)q(u',u) = \pi(u')P_{u',u}$ since $\pi(u') \leq \pi(u)$. Consequently, π is reversible with respect to P, thus stationary for (Υ_n).

3) We conclude by applying the ergodic Theorem to the function $h = Id_E$, which is π-integrable as required. □

3.2 Convergence control

Theorem 1 is a theoretical result ensuring convergence when n tends to infinity, but in practice the real issue is: how large should n be to permit a reasonable approximation of $\hat{u}_{LSE}$? Here the speed of convergence depends on two factors: first, the number of iterations needed by the Markov Chain (U_n) to attain the stationary state; second, the number of iterations needed by the empirical average to estimate reasonably the true expectation. In MCMC simulations, it is common to introduce a "burn-in" phase, during which the random images are generated with the Markov chain but not taken into account in the expectation estimate. It amounts to consider, for $0 \leq b < n$, the partial average

$$S_n^b = \frac{1}{n-b} \sum_{k=b+1}^{n} U_k, \tag{7}$$

from which we compute

$$\mathbb{E}\|S_n^b - \hat{u}_{LSE}\|^2 = \mathbb{E}\|S_n^b - \mathbb{E}S_n^b\|^2 + \|\mathbb{E}S_n^b - \hat{u}_{LSE}\|^2.$$

The first term is the span (trace of the covariance matrix) of the estimator S_n^b (written $\mathrm{Span}(S_n^b)$), which converges to 0 like $A/(n-b)$ (for some constant A) when n tends to infinity [7]. The second term is the (squared) bias, due to the fact that the law of U_n is not exactly the posterior law, but (only) tends to it when $n \to +\infty$ by ergodic Theorem. If we assume that the state space is finite (which is always numerically the case), the convergence of U_n to the posterior distribution is geometric and

$$\forall n \geq 1, \ \ \|\mathbb{E}U_n - \hat{u}_{LSE}\| \leq B\gamma^n, \tag{8}$$

for some $B > 0$ and $0 < \gamma < 1$. From (8), we deduce that

$$\|\mathbb{E}S_n^b - \hat{u}_{LSE}\| \leq \frac{B}{n-b} \frac{\gamma^{b+1}(1-\gamma^{n-b})}{1-\gamma},$$

so that for $B' = \gamma^2 B^2/(1-\gamma)^2$ we have

$$\mathbb{E}\|S_n^b - \hat{u}_{LSE}\|^2 \leq \mathrm{Span}(S_n^b) + \frac{B'\gamma^{2b}}{(n-b)^2}. \tag{9}$$

Hence, S_n^b converges to $\hat{u}_{LSE}$ like $1/\sqrt{n}$ (recall that $\mathrm{Span}(S_n^b) \simeq A/(n-b)$), which is a rather slow convergence, that requires a good stopping criterion. Now let us consider another Markov chain $\tilde{U}_n$ defined like U_n (and independent from it). With obvious notations we have

$$\mathbb{E}\|\tilde{S}_n^b - S_n^b\|^2 = 2\ \mathrm{Span}(S_n^b),$$

and the empirical value $\|\tilde{S}_n^b - S_n^b\|^2$ is a very good approximation of its expectation since the dimension of the image space is very high. Thus, if we manage to choose b large enough to ensure that the bias term (rightmost term) is negligible in (9), then we expect to have

$$\|S_n^b - \hat{u}_{LSE}\| \simeq \frac{e_b}{\sqrt{2}} \quad \text{with} \quad e_b = \|\tilde{S}_n^b - S_n^b\|, \tag{10}$$

and we can use a test like $e_b \leq \varepsilon$ as a stopping criterion. Now how do we select the correct burn-in parameter b? Empirically, it can be observed that the function e_b decreases with b

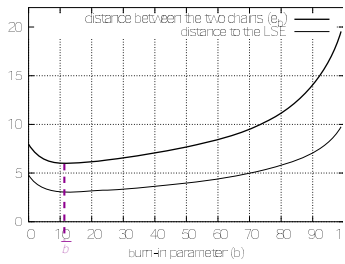

Figure 1: Selection of the burn-in parameter b. As a function of b, the distance $e_b = \|\tilde{S}_n^b - S_n^b\|$ between the two MC estimates (thick line) reaches a minimum value for $b = \bar{b}$. This value is a good choice for the burn-in parameter, because it is very close to the value of b for which the distance from $(S_n^b + \tilde{S}_n^b)/2$ to $\hat{u}_{LSE}$ is minimal (experiment made with $\lambda = 30$ and $\sigma = 10$ on a noisy image).

for small values of b, reaches a minimum, then increases with b (see Figure 1). This highlights a competition between the burn-in time b (that should not be too short because $\mathrm{Span}(U_n)$ decreases with n, as can be seen empirically) and the number of samples $(n-b)$ kept for the estimation. Intuitively, the value $\bar{b} = \arg\min_b e_b$ is an interesting compromise for b, since it is very close to the optimal value (see Figure 1).

3.3 Algorithm

The considerations above lead to the following algorithm.

Algorithm 1 TV-LSE algorithm

```
draw two random images U_0 and Ũ_0
n ← 0
repeat
    n ← n + 1
    draw U_n and Ũ_n from U_{n-1} and Ũ_{n-1}
    compute b̄ = arg min_b e_b
until e_b̄ ≤ 2ε
return (S_n^b̄ + S̃_n^b̄)/2
```

In practice the initial images U_0 and $\tilde{U}_0$ are drawn with i.i.d uniform intensity values in $[0, 256)$. Concerning the partial sums S_n^b, since all images $(U_k)_{1 \leq k \leq n}$ cannot be kept in memory at the same time, we constrain b to belong to a discrete set of values $E = \lfloor \lambda^{\mathbb{N}} \rfloor = \{\lfloor \lambda^p \rfloor, p \in \mathbb{N}\}$ (where $\lfloor \cdot \rfloor$ denotes the lower integer part, and $\lambda = 1.2$ in practice) and to be larger than a fraction of n ($n/6$ in practice). Hence, we simply have to maintain the partial sum S_n^b for $b \in E \cap [n/6, n)$ for all n, and there are at most 10 such values of b for any n (because $-\log(\frac{1}{6})/\log 1.2 \simeq 9.8$).

At the end of the algorithm, we estimate $\hat{u}_{LSE}$ with $(S_n^b + \tilde{S}_n^b)/2$ which is better than either S_n^b or $\tilde{S}_n^b$. Indeed, since S_n^b and $\tilde{S}_n^b$ are independent, we have, with a similar computation as before

$$\mathbb{E}\left\| \frac{S_n^b + \tilde{S}_n^b}{2} - \hat{u}_{LSE} \right\|^2 \leq \frac{1}{2}\mathrm{Span}(S_n^b) + \|\mathbb{E}S_n^b - \hat{u}_{LSE}\|^2.$$

In other terms, by averaging the two chains we maintain the same bias and divide the span by two. If the bias is negligible (it is the case in general when b is chosen large enough), then

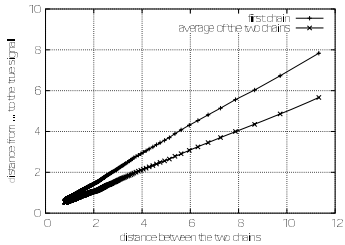

Figure 2: The curve $(e_b, \|S_n^b - \hat{u}_{LSE}\|)_n$ (thick line) and the curve $(e_b, \|(S_n^b + \tilde{S}_n^b)/2 - \hat{u}_{LSE}\|)_n$ (thin line) for a fixed value of b and $\lambda = 30$, $\sigma = 10$. Since the ratio of the two curves is approximately $\sqrt{2}$, the bias is negligible, which suggests that the MCMCs have reached the stationary regime.

we can expect to have $\|(S_n^b + \tilde{S}_n^b)/2 - \hat{u}_{LSE}\| \simeq e_b/2$, which is $\sqrt{2}$ times better than S_n^b or $\tilde{S}_n^b$ alone. This property, which can be checked numerically on Figure 1 (the ratio between the two functions is approximately 2), can be used to check the absence of bias, since averaging the two chains reduce the span but not the bias. Figure 2 corresponds to the same situation as Figure 1, that is $\lambda = 30$ and $\sigma = 10$, and also illustrates the negligibility of the bias in that case (the ratio of the two curves is approximately $\sqrt{2}$).

4. PROPERTIES OF TV-LSE DENOISING

4.1 LSE versus MAP

Whereas the classical TV denoising ($\hat{u}_{MAP}$) only depends on the parameter λ, the TV-LSE denoising depends on two parameters, λ and σ. In the Bayesian framework, σ^2 represents the variance of the noise, which indirectly controls the spread of the posterior distribution. Hence, even if it is natural to choose for σ^2 the (supposedly known) variance of the noise, we can also consider it as an abstract hyperparameter, as is λ in the variational TV denoising framework. First, we investigate the extreme values of σ from a theoretical point of view. We note $\hat{u}_{MAP}(\lambda)$ and $\hat{u}_{LSE}(\lambda, \sigma)$ the MAP and LSE results obtained from a given image u_0.

Theorem 2 *For all $\lambda > 0$, we have*

$$(i) \quad \hat{u}_{LSE}(\lambda, \sigma) \xrightarrow[\sigma \to 0]{} \hat{u}_{MAP}(\lambda),$$
$$(ii) \quad \hat{u}_{LSE}(\lambda, \sigma) \xrightarrow[\sigma \to +\infty]{} u_0.$$

Proof — When σ goes to 0, the unimodal probability distribution $\frac{1}{Z}\exp\left(-\frac{E_\lambda}{2\sigma^2}\right)$ converges to the Dirac distribution in $\hat{u}_{MAP}(\lambda) = \arg\min_u E_\lambda(u)$, whose expectation is $\hat{u}_{MAP}(\lambda)$, which proves (i). For (ii), consider the change of variable $u_0' = \frac{u_0}{\sigma}$ and $u' = \frac{u}{\sigma}$, then

$$\hat{u}_{LSE}(\lambda, \sigma) = \frac{\int_{\mathbb{R}^\Omega} \sigma u' e^{-\frac{1}{2}(\|u'-u_0'\|^2 + \frac{\lambda}{\sigma} TV(u'))} du'}{\int_{\mathbb{R}^\Omega} e^{-\frac{1}{2}(\|u'-u_0'\|^2 + \frac{\lambda}{\sigma} TV(u'))} du'}$$

so that, thanks to Lebesgue's dominated convergence theorem,

$$\hat{u}_{LSE}(\lambda, \sigma) \underset{\sigma \to \infty}{\sim} \sigma \frac{\int_{\mathbb{R}^\Omega} u' e^{-\frac{1}{2}\|u'-u_0'\|^2} du'}{\int_{\mathbb{R}^\Omega} e^{-\frac{1}{2}\|u'-u_0'\|^2} du'} \underset{\sigma \to \infty}{\sim} \sigma u_0' = u_0. \qquad \square$$

Thus, TV-MAP denoising can be seen as a special case of TV-LSE denoising, corresponding to $\sigma = 0$. When σ is very small, the posterior distribution is very concentrated around $\hat{u}_{MAP}$, so that starting the Markov chains with a random image causes a lot of bias, which makes our stopping criterion incapable of guaranteeing the precision ε on $\hat{u}_{LSE}$. We could improve a lot the previous algorithm for small values of σ by choosing to start the Markov chains with $\hat{u}_{MAP}$ (instead of random images), but this is not really worth it, since when σ is small, $\hat{u}_{LSE}$ is very close to $\hat{u}_{MAP}$, and hence not specially interesting. In practice, we never encountered convergence problems with the algorithm described in the previous section as soon as $\sigma \geq \lambda/10$.

A natural idea at this point would be to compare TV-MAP and TV-LSE denoising by keeping a fixed value of λ and making σ vary. This is not very interesting because, as one can see in the experiments, the method noise $\|\hat{u}_{LSE} - u_0\|$ decreases with σ, so that not only the denoising technique is different, but also the "amount of denoising". This is the reason why in the comparison we make, we always choose the denoising parameters (λ for the MAP, λ and σ for the LSE) so that the method noise is fixed. For TV-LSE, this leaves one degree of freedom that allows more or less departure from the TV-MAP model. A systematic exploration of this degree of freedom could be interesting, but for the sake of concision we shall only try to give an insight of TV-LSE denoising abilities by choosing one arbitrary (reasonable) value in the experiments.

Apart from computation time (typically 10 seconds for TV-MAP and 10 minutes for TV-LSE on a 512×512 image), the main difference between TV-MAP and TV-LSE denoising, illustrated on Figure 3, is the ability to TV-LSE to avoid two annoying artifacts of TV-MAP denoising: the staircasing effect and the creation of isolated pixels. These artifacts are even created by TV-MAP in a pure noise image (see Figure 4), which contradicts what could be considered as a basic requirement, that is, that a good denoising method should not create structures in noise (this requirement is very important for satellite image interpretation for example).

4.2 No staircasing effect for TV-LSE

We conclude with a theoretical result stating the absence of staircasing for TV-LSE denoised images.

Theorem 3 *Let u_0 be a random image such that the distribution of u_0 is absolutely continuous with respect to Lebesgue's measure. Let $k, k' \in \Omega$ be neighbor pixels (that is, such that $|k - k'| = 1$). Then the denoised image $\hat{u}_{LSE}$ satisfies*

$$\mathbb{P}_{u_0}\left(\hat{u}_{LSE}(k') = \hat{u}_{LSE}(k)\right) = 0.$$

Sketch of the proof — In this proof the gray value of any image u at pixel k will be denoted by u_k, and $v := u_0$ to simplify notations. Let $g : \mathbb{R} \to \mathbb{R}$ the function defined by

$$g(z) = \frac{\int_{\mathbb{R}^\Omega} (u_{k'} - u_k) \exp\left(-\frac{(u_k - z)^2 + \sum_{l \neq k}(u_l - v_l)^2 + \lambda TV(u)}{2\sigma^2}\right) du}{\int_{\mathbb{R}^\Omega} \exp\left(-\frac{(u_k - z)^2 + \sum_{l \neq k}(u_l - v_l)^2 + \lambda TV(u)}{2\sigma^2}\right) du}. \tag{11}$$

Then $\hat{u}_{LSE,k'} = \hat{u}_{LSE,k}$ is equivalent to $g(v_k) = 0$. With (11), the function g can be extended to a holomorphic $\mathbb{C} \to \mathbb{C}$ mapping, which proves that g is analytic. Now Assume that g is zero everywhere. The numerator of $g(z)$ in (11), written N, can be rewritten as the convolution product $N = G_\sigma * \varphi$, where G_σ is the cen-

Figure 3: Comparison between classical TV-MAP denoising and the proposed TV-LSE denoising on a part of the classical Lena image (top, left) corrupted by a Gaussian white noise with standard deviation $\sigma = 10$ (top, right). On the middle row, TV-MAP denoising has been applied for two levels of denoising (the level of denoising being measured by the "method noise", that is, the L^2 distance between the noisy image and the result). Left: method noise is 9.6 (to be compared to 10, the actual noise level), obtained with $\lambda = 19.25$; right: method noise is 7.68, obtained with $\lambda = 11.1$. On the bottom row, TV-LSE denoising has been applied for the same levels of denoising (same method noise). Left: method noise is 9.6, obtained with $(\lambda, \sigma) = (50, 20)$; right: method noise is 7.68, obtained with $(\lambda, \sigma) = (25, 15)$. On both TV-MAP images (middle row), the staircasing effect is visible: artificial boundaries are created between extremely flat zones. On the right TV-MAP image, another artifact appears: isolated pixels with extreme intensity values remain. These two artifacts do not arise with the TV-LSE images (bottom row), that look much more natural.

tered Gaussian function of bandwidth σ and φ is a real function defined by $\varphi(x) = \int (u_{k'} - u_k) \exp\left(-\frac{\sum_{l\neq k}(u_l - v_l)^2 + \lambda TV(u^{k,x})}{2\sigma^2}\right) d(u_l)_{l\neq k}$, where $u^{k,x}$ is the image defined by $u_l^{k,x} = u_l$ if $l \neq k$, and $u_k^{k,x} = x$. It can be proven that the discrete formulation of TV (Equation 1) ensures that φ is in L^1, so that by considering Fourier transforms (written $\hat{\cdot}$) we get $\widehat{G_\sigma}(\xi) \cdot \hat{\varphi}(\xi) = 0$ for all $\xi \in \mathbb{R}$. Since $\widehat{G_\sigma}(\xi)$ never vanishes, we deduce that $\hat{\varphi}$ is identically null, and so is φ. But as $\varphi(z)$ can be proved to be negative for large enough z, we have a contradiction, which proves that g cannot be zero everywhere.

Last, since g is analytic and non-identically null, the isolated zero Theorem states that $g^{-1}(\{0\})$ cannot contain any accumulation point. Thus, under the marginal distribution of v_k, the event $(g(v_k) \neq 0)$ a.s. occurs. Since v has been assumed to have a density with respect to Lebesgue measure, this yields

$$\begin{aligned}\mathbb{P}_v(g(v_k) = 0) &= \int_{(v_l)_{l\neq k}} \mathbb{P}_{v_k}(g(v_k) = 0) f(v)\, d(v_l)_{l\neq k} = 0 \\ &= \int_{(v_l)_{l\neq k}} 0 \cdot f(v)\, d(v_l)_{l\neq k} = 0\end{aligned}$$

which concludes the proof. □

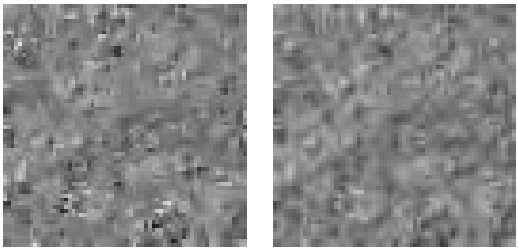

Figure 4: A pure noise image (Gaussian white noise with standard deviation 10) is denoised with the classical TV-MAP (left) and the proposed TV-LSE (right) method, with the same level of method noise (7.33), achieved with $\lambda = 9.37$ for MAP and $(\lambda, \sigma) = (40, 20)$ for LSE. As we can see, TV-MAP denoising creates severe structures in noise: artificial boundaries between artificially flat zones (the staircasing effect), and isolated pixels with extreme values. Like for the Lena image (Figure 3), these artifacts are avoided by the TV-LSE denoising method.

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Hindawi Publishing Corporation
EURASIP Journal on Advances in Signal Processing
Volume 2007, Article ID 96747, 14 pages
doi:10.1155/2007/96747

Research Article

Wideband Impulse Modulation and Receiver Algorithms for Multiuser Power Line Communications

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Received 8 November 2006; Accepted 23 March 2007

Recommended by Moisés Vidal Ribeiro

We consider a bit-interleaved coded wideband impulse-modulated system for power line communications. Impulse modulation is combined with direct-sequence code-division multiple access (DS-CDMA) to obtain a form of orthogonal modulation and to multiplex the users. We focus on the receiver signal processing algorithms and derive a maximum likelihood frequency-domain detector that takes into account the presence of impulse noise as well as the intercode interference (ICI) and the multiple-access interference (MAI) that are generated by the frequency-selective power line channel. To reduce complexity, we propose several simplified frequency-domain receiver algorithms with different complexity and performance. We address the problem of the practical estimation of the channel frequency response as well as the estimation of the correlation of the ICI-MAI-plus-noise that is needed in the detection metric. To improve the estimators performance, a simple hard feedback from the channel decoder is also used. Simulation results show that the scheme provides robust performance as a result of spreading the symbol energy both in frequency (through the wideband pulse) and in time (through the spreading code and the bit-interleaved convolutional code).

1. INTRODUCTION

The design of broadband communication modems for transmission over power lines (PL) is an interesting and open problem especially with reference to the development of reliable transmission and advanced signal processing techniques that are capable of coping with the harsh properties of the power line channel and noise [1]. In this paper, we deal with advanced signal processing algorithms for a wideband (beyond 20 MHz) impulse-modulated modem [2–4]. Up to date, impulse modulation has only been considered for application in ultra-wideband (UWB) wireless channels [5–7]. It has interesting properties in terms of simple baseband implementation and robustness against channel frequency selectivity and interference. Differently from the wireless context, PL channels have a narrower transmission bandwidth [8] and are characterized by several background disturbances as colored and impulse noise [9]. Nevertheless, wideband impulse modulation is an attractive scheme for application over this medium as experimental trials have shown [4]. The basic idea behind impulse modulation is to convey information by mapping an information symbol stream into a sequence of short-duration pulses. Pulses (referred to as monocycles) are followed by a guard time to cope with the channel time dispersion. The monocycle can be designed to shape the occupied spectrum and in particular to avoid the low frequencies where we typically experience higher levels of background noise. Since our system deploys a fractional bandwidth (ratio between signaling bandwidth and center carrier) larger than 20%, it can be classified as an ultra wideband system according to the FCC. We consider indoor applications such as local area networks, peripheral office connectivity, and home/industrial automation. Impulse modulation is an attractive transmission technique also for in-vehicle PLC systems and for PL pervasive sensor networks where the transmitting nodes need to use a simple modulation scheme. In general, we assume that a number of nodes (users) wish to communicate sharing the same PL grid. Communication is from one node to another node such that if other nodes simultaneously access the medium, they are seen as potential interferers. In order to allow for users' multiplexing, we deploy direct-sequence code-division multiple access (DS-CDMA) [6, 10–12]. The user's information is conveyed using a certain signature waveform that is a repetition of time-delayed and weighted monocycles that span a transmission frame.

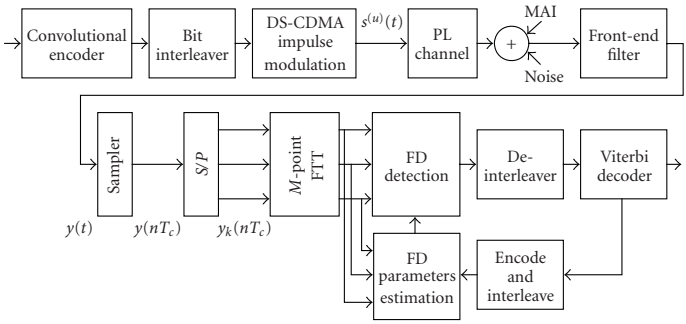


Figure 1: Impulse-modulated PL system with frequency-domain receiver processing and iterative decoding.

A key point in the proposed approach is that the symbol energy is spread over a wideband which makes the system robust to narrowband interference and capable of exploiting the channel frequency diversity. Furthermore, this modulation approach is simple at the transmitter side and requires a baseline correlation receiver that filters the received signal with a template waveform [2, 7]. The template waveform has to be matched to the equivalent impulse response that comprises the desired user's waveform and the channel impulse response. To achieve high performance, this receiver requires accurate estimation of the channel which can be complex if performed in the time domain [13, 14] because of the large time dispersion that is introduced by the wideband frequency-selective PL channel. Further, the channel-frequency selectivity introduces intercode interference (ICI) (interference among the codes that are assigned to the same user) and multiple-access interference (MAI) when multiple users access the network. This translates into performance losses and suggests some form of multiuser detection or interference cancellation. Therefore, in this paper we focus on the receiver side and we propose a novel frequency-domain (FD) detection approach which allows to obtain high performance and to keep the complexity at moderate levels. FD receivers have recently attracted considerable attention both for equalization in single carrier systems [15] and in multicarrier (OFDM) systems [16, 17]. We have investigated FD processing in a UWB wireless system in [10], and described preliminary results for the power line scenario in [11, 12]. The contribution of the present paper is about the derivation of a maximum likelihood joint detector that operates in the frequency domain in the presence of MAI and impulse noise (Section 3). The detection metric used in this receiver is conditional on the knowledge of the channel of the desired user and on the knowledge of the occurrence of the impulse noise.

From this receiver, with certain approximations, we describe in Section 4 several novel FD algorithms, in particular, a simplified FD joint detector, an FD iterative detector, and an FD interference decorrelator. They all include the capability of adapting to impulse noise and rejecting the ICI/MAI, but have different levels of performance and complexity.

We focus on the practical estimation of the parameters that are needed in the detection algorithms (Section 5). In particular, we address the FD channel estimation problem, the estimation of the correlation of the noise and the interference, and the estimation of the impulse noise occurrence. Frequency-domain channel estimation for the desired user is done with a recursive least-squares (RLS) algorithm [18]. Further, channel coding is also considered and it is based on bit-interleaved convolutional codes. In this case, we show that iterative processing [19] with simple hard feedback from the decoder allows to run the parameter estimators in a data decision-driven mode which betters the overall receiver performance.

Finally, we describe in Section 6 the key features of a PL impulse-modulated modem that has been used to assess performance and whose hardware prototype is described in [4]. To this respect, we propose the use of a wideband statistical channel model that allows to evaluate the system performance by capturing the ensemble of indoor PL grid topologies.

2. WIDEBAND SYSTEM MODEL

We consider a system where a number of nodes (users) communicate sharing the same PL network. Communication is from one node to another, such that if other nodes simultaneously access the medium, they are seen as potential interferers. The transmission scheme (Figure 1) uses wideband impulse modulation combined with DS data spreading [11]. Users' multiplexing is obtained in a CDMA fashion allocating the spreading codes among the users.

The signal transmitted by user u can be written as

$$s^{(u)}(t) = \sum_{k} \sum_{i \in C_u} b_k^{(u,i)} g^{(u,i)}(t - kT_f), \tag{1}$$

where $g^{(u,i)}(t)$ is the waveform (signature code) used to convey the ith information symbol $b_k^{(u,i)}$ of user u that is transmitted during the kth frame. Each symbol belongs to the pulse amplitude modulation (PAM) alphabet [18], and it

Figure 2: Frame format for user u and code i.

carries $\log_2 M_S$ information bits where M_S is the number of PAM levels, for example, with 2-PAM $b_k^{(u,i)}$ has alphabet $\{-1, 1\}$. T_f is the symbol period (frame duration) as shown in Figure 2. C_u denotes the set of code indices that are allocated to user u. Thus, user u can adapt its rate by transmitting $|C_u| = \text{size}\{C_u\}$ information symbols per frame.

The signature code (Figure 2) comprises the weighted repetition of $L \geq 1$ narrow pulses (monocycles):

$$g^{(u,i)}(t) = \sum_{m=0}^{L-1} c_m^{(u,i)} g_M(t - mT), \tag{2}$$

where $c_m^{(u,i)} \in \{-1, 1\}$ are the codeword elements (chips), and T is the chip period. The monocycle $g_M(t)$ can be appropriately designed to shape the spectrum occupied by the transmission system. In this paper we consider the second derivative of the Gaussian pulse (Figure 3(a)). An interesting property is that its spectrum does not occupy the low frequencies where we experience higher levels of man-made background noise (Figure 3(b)). Further, the symbol energy is spread over a wideband which makes the system robust to narrowband interference and capable of exploiting the channel frequency diversity. Since the attenuation in PL channels increases with frequency, we limit the transmission bandwidth to about 50 MHz using a pulse with $D = 126$ nanoseconds. In typical system design, we choose the chip period $T \geq D$ and we further insert a guard time T_g between frames to cope with the channel time dispersion (Figure 2). The frame duration has, therefore, duration $T_f = LT + T_g$.

2.1. User multiplexing

Users are multiplexed by assigning distinct codes to distinct users. In our design, the codes are defined as follows:

$$c_m^{(u,i)} = c_{1,m}^{(u)} c_{2,m}^{(i)}, \quad m = 0, \ldots, L-1, \; i = 0, \ldots, L-1, \tag{3}$$

where $\{c_{1,m}^{(u)}\}$ is a binary (± 1) pseudorandom sequence of length L allocated to user u, while $\{c_{2,m}^{(i)}\}$ is the ith binary (± 1) Walsh Hadamard sequence of length L [18]. With this choice, each node can use all L Walsh codes, which yields a peak data rate per user equal to $R = L/T_f$ symb/s. It approaches $\log_2 M_S/T$ bit/s with long codes. While the signals of a given user are orthogonal, the ones that belong to distinct transmitting nodes are not. The random code $\{c_{1,m}^{(u)}\}$ is used to introduce code diversity and to randomize the effect of the MAI.

2.2. Channel coding

We consider the use of bit-interleaved convolutional codes (Figure 1) [18]. A block of information bits is coded, interleaved, and then modulated. Interleaving spans a packet of N frames that we refer to as superframe. This coding approach yields good performance also in the presence of impulse noise as it will be shown in the following.

2.3. Received signal

The signals that are transmitted by distinct nodes (users) propagate through distinct channels with impulse response $h^{(u)}(t)$. At the receiver of the desired node, we deploy a bandpass front-end filter with impulse response $g_{\text{FE}}(t) = g_M(-t)$ that is matched to the transmit monocycle and that suppresses out-of-band noise and interference. Then, the output signal in the presence of N_I other users (interferers) reads

$$\begin{aligned} y(t) &= \sum_k \sum_{i \in C_0} b_k^{(0,i)} g_{\text{EQ}}^{(0,i)}(t - kT_f) + I(t) + \eta(t) \\ I(t) &= \sum_k \sum_{u=1}^{N_I} \sum_{i \in C_u} b_k^{(u,i)} g_{\text{EQ}}^{(u,i)}(t - kT_f - \Delta_u), \end{aligned} \tag{4}$$

where the equivalent impulse response for user u and symbol i (equivalent signature code) is denoted as $g_{\text{EQ}}^{(u,i)}(t) = g^{(u,i)} * h^{(u)} * g_{\text{FE}}(t)$. It comprises the convolution of the signature code of indices (u, i) with the channel impulse response of the corresponding user, and the front-end filter. The index $u = 0$ denotes the desired user. Δ_u denotes the time delay of user u with respect to the desired user's frame timing. $I(t)$ is the MAI term, while $\eta(t)$ denotes the additive noise. The users experience distinct channels that introduce identical maximum time dispersion.

2.4. Noise models

In this paper, we consider the presence of background colored and impulse noise [9]. Several impulse noise models have been proposed in the literature. For instance, the class A-B Middleton and the two-term Gaussian models [20, 21] have been used to characterize the probability density function (pdf) of the impulse noise. The temporal characteristics of asynchronous (to the main cycle) impulse noise have been modeled via Markov chains [9], or using a simple modification of the two-term mixture model which assumes that when a spike occurs, it lasts for a given amount of time [22]. In the receiver algorithms that we describe, differently from other approaches, we do not use optimal metrics that are based on the assumption of a stationary white noise process with a given pdf, for example, [23, 24]. In our approach

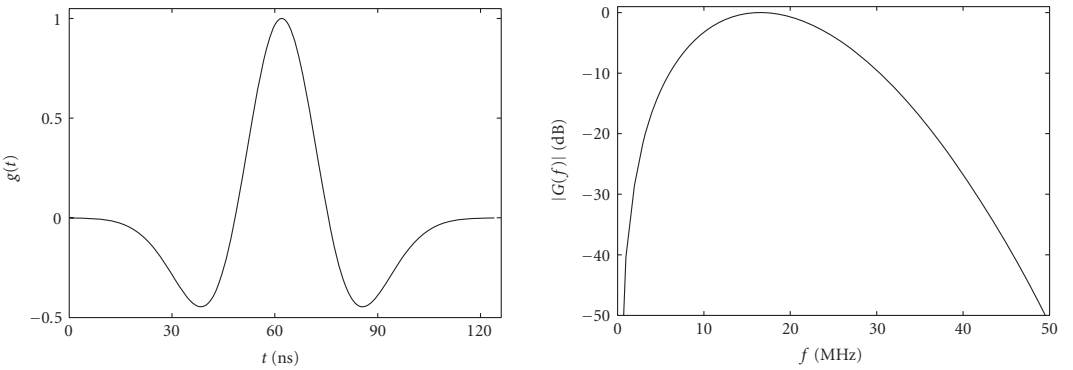


FIGURE 3: (a) Monocycle impulse response, $g_M(t) \sim (1 - \pi((t - D/2)/T_0)^2)\exp(-\pi/2((t - D/2)/T_0)^2)$, where $D \approx 5.23T_0$ is the monocycle duration. (b) Monocycle frequency response.

(see Section 3), the receiver adapts to the impulse noise occurrence and treats it as a nonstationary colored Gaussian process. To do so, as it will be explained, we need to estimate the impulse noise occurrence and its locally stationary correlation.

2.5. Statistical channel model

The frequency-selective PL channel is often modelled according to [8], that is, we synthesize the bandpass frequency response with N_P multipaths as

$$H_+(f) = \sum_{p=1}^{N_P} g_p e^{-j(2\pi d_p/v)f} e^{-(\alpha_0+\alpha_1 f^K)d_p}, \quad 0 \le B_1 \le f \le B_2, \tag{5}$$

where $|g_p| \le 1$ is the transmission/reflection factor for path p, d_p is the length of the path, $v = c/\sqrt{\varepsilon_r}$ with c speed of light, and ε_r, dielectric constant. The parameters α_0, α_1, K are chosen to adapt the model to a specific network. To assess the system performance, we may use this model once the reference parameters are chosen. Instead, we propose to evaluate performance with a statistical model that allows to capture the ensemble of PL grid topologies. It is obtained by considering the parameters in (5) as random variables. Then, we generate channel realizations through realization of the random parameters. We assume the reflectors (that generate the paths) to be placed over a finite distance interval. We fix the first reflector at distance d_1 and we assume the other reflectors to be located according to a Poisson arrival process with intensity $\Lambda[m^{-1}]$. The reflection factors g_p are assumed to be real, independent, and uniformly distributed in $[-1, 1]$. Finally, we appropriately choose α_0, α_1, K to a fixed value. If we further assume $K = 1$, the real impulse response can be obtained in closed form. This allows to easily generate a realization for user u (corresponding to a realization of the random parameters N_P, g_p, d_p) as follows:

$$h^{(u)}(t) = 2\,\mathrm{Re}\left\{\sum_{p=1}^{N_P}\left(g_p e^{-\alpha_0 d_p}\frac{\alpha_1 d_p + j2\pi(t - d_p/v)}{(\alpha_1 d_p)^2 + 4\pi^2(t - d_p/v)^2} \times \left(e^{j2\pi B_1(t-d_p/v)-\alpha_1 B_1 d_p} - e^{j2\pi B_2(t-d_p/v)-\alpha_1 B_2 d_p}\right)\right)\right\}. \tag{6}$$

We assume distinct users to experience independent channels, that is, the random parameters are independent for the channels of distinct users, which is appropriate in indoor PL channels due to the large number of path components. The impulse responses are assumed to be constant for a given amount of time and they change for a new (randomly picked) topology.

3. DETECTION ALGORITHMS FOR THE IMPULSE-MODULATED SYSTEM

In this section, we derive several detection algorithms that operate in the frequency domain (FD). Their performance is compared with the baseline correlation receiver as reported in Section 6.

3.1. Baseline receiver

The baseline receiver for the impulse-modulated system is the correlation receiver. Assuming binary data symbols, it computes the correlation between the received signal $y(t)$ and the real equivalent impulse response $g_{\mathrm{EQ}}^{(0,i)}(t)$. Thus, we obtain the decision metric $z_{\mathrm{DM}}^{(0,i)}(kT_f) = \int_{\mathbb{R}} y(t) g_{\mathrm{EQ}}^{(0,i)}(t - kT_f)dt$ for the ith symbol that is transmitted by user 0 in the kth frame. Then, a threshold decision is made, that is,

$\hat{b}_k^{(0,i)} = \text{sign}\{z_{\text{DM}}^{(0,i)}(kT_f)\}$. This baseline correlation receiver is optimal when the background noise is white Gaussian and there is perfect orthogonality among the received signature codes [2]. To implement the correlation receiver, we need to estimate the channel. Time-domain channel estimation [3, 13, 14] is complicated due to the large time dispersion of the PL channel that implies that $g_{\text{EQ}}^{(0,i)}(t)$ is an involved function of the channel and the transmitted waveform. Furthermore, the correlation receiver suffers from the presence of intercode interference (ICI) and multiple-access interference (MAI) that is generated by the dispersive PL channel in the presence of multiple users.

3.2. Maximum likelihood frequency-domain receiver

To improve the performance of the baseline receiver, we propose an FD signal processing approach. To derive the receiver algorithms, we treat the noise as the sum of two Gaussian distributed processes. Similarly, the receiver treats the MAI as Gaussian. Therefore, the overall impairment process is modeled by the receiver as

$$z(t) = \eta(t) + I(t) = w_T(t) + \alpha(t)w_{\text{IM}}(t) + I(t), \tag{7}$$

where $w_T(t)$ is the thermal noise, $w_{\text{IM}}(t)$ is the impulse noise, and $I(t)$ is the MAI. The multiplicative process $\alpha(t)$ accounts for the presence or absence of impulse noise. That is, at time instant t, the random variable $\alpha(t)$ is a Bernoulli random variable with parameter p and alphabet $\{0, 1\}$. We refer to it as Bernoulli process. All processes are treated as independent zero-mean Gaussian, not necessarily stationary, with correlation, respectively, as

$$\begin{aligned} \kappa_T(\tau_1, \tau_2) &= E[w_T(\tau_1)w_T(\tau_2)], \\ \kappa_{\text{IM}}(\tau_1, \tau_2) &= E[w_{\text{IM}}(\tau_1)w_{\text{IM}}(\tau_2)], \\ \kappa_I(\tau_1, \tau_2) &= E[I(\tau_1)I(\tau_2)]. \end{aligned} \tag{8}$$

Conditional on the Bernoulli process, the impairment is a Gaussian process with correlation

$$\begin{aligned} &\kappa_{z|\alpha}(\tau_1, \tau_2 \mid \alpha(t),\ t \in \mathbb{R}) \\ &\quad = \kappa_W(\tau_1, \tau_2) + \alpha(\tau_1)\alpha(\tau_2)\kappa_{\text{IM}}(\tau_1, \tau_2) + \kappa_I(\tau_1, \tau_2). \end{aligned} \tag{9}$$

The Gaussian approximation for the MAI improves as the number of interferers increases. The model used for the overall noise contribution allows to capture both stationary and nonstationary components of it. Further, it allows to describe impulse spikes of certain duration, power decay profile, and colored spectral components.

To proceed, we assume discrete-time processing (Figure 1) such that the received signal is sampled with period $T_c = T_f/M$, where M is the number of samples/frame, to obtain

$$y(nT_c) = \sum_k \sum_{i \in C_0} b_k^{(0,i)} g_{\text{EQ}}^{(0,i)}(nT_c - kT_f) + z(nT_c). \tag{10}$$

If we acquire frame synchronization with the desired user and we assume that the guard time is sufficiently long not to have interframe interference, that is, interference among the symbols of adjacent frames, we can write

$$\begin{aligned} y_k(nT_c) &= \sum_{i \in C_0} b_k^{(0,i)} g_{\text{EQ}}^{(0,i)}(nT_c - kT_f), \\ &\quad + z_k(nT_c) \quad n = 0, \ldots, M-1, \end{aligned} \tag{11}$$

with $y_k(nT_c) = y(kMT_c + nT_c)$, and $z_k(nT_c) = z(kMT_c + nT_c)$, $k \in \mathbb{Z}$.

Under the colored Gaussian impairment model in (7), and under the knowledge of both the channel and the Bernoulli process $\alpha(t)$ (meaning that we assume to know when the impulse noise occurs), the maximum likelihood receiver searches for the sequence of transmitted symbols $\mathbf{b}^{(0)} = \{b_k^{(0,i)}, k \in \mathbb{Z}, i \in C_0\}$ (belonging to the desired user) that maximizes the logarithm of the probability density function of the received signal $\mathbf{y} = \{\ldots, y(0), y(T_c), \ldots\}$ conditional on a given hypothetical transmitted symbol sequence, that is, $\log p(\mathbf{y} \mid \mathbf{b}^{(0)})$, [18, 25]. It follows that we have to search for the symbol sequence that minimizes the following log-likelihood function[1]

$$\begin{aligned} &\Lambda(\mathbf{b}^{(0)}) \\ &\quad = \sum_{l=-\infty}^{\infty} \sum_{m=-\infty}^{\infty} \Big(y(lT_c) - \sum_k \sum_{i \in C_0} b_k^{(0,i)} g_{\text{EQ}}^{(0,i)}(lT_c - kT_f) \Big) \\ &\qquad \times K^{-1}(lT_c, mT_c) \\ &\qquad \times \Big(y(mT_c) - \sum_k \sum_{i \in C_0} b_k^{(0,i)} g_{\text{EQ}}^{(0,i)}(mT_c - kT_f) \Big), \end{aligned} \tag{12}$$

where $K^{-1}(lT_c, mT_c)$ is the element of indices (l, m) of the matrix $\mathbf{K}^{-1}$, that is, the inverse of the correlation matrix of the impairment vector $\mathbf{z} = [\ldots, z(0), z(T_c), \ldots]$,

$$\mathbf{K} = E[\mathbf{z}\mathbf{z}^T]. \tag{13}$$

The elements of $\mathbf{K}$ are obtained by sampling (9) in the appropriate time instants, that is,

$$K(lT_c, mT_c) = \kappa_{z|\alpha}(lT_c, mT_c \mid \alpha(t),\ t \in \mathbb{R}). \tag{14}$$

As an example, if we suppose the absence of MAI, the diagonal elements of $\mathbf{K}$ represent the power of the thermal plus impulse noise, and they are typically large in the presence of impulse noise.

The likelihood (12) can be written as the scalar product $\Lambda(\mathbf{b}^{(0)}) = \mathbf{e}^\dagger \mathbf{K}^{-1} \mathbf{e} = \langle \mathbf{e}, \mathbf{K}^{-1}\mathbf{e} \rangle$ if we define the vector $\mathbf{e} = [\ldots, e(0), e(T_c), \ldots]^T$, with $e(lT_c) = y(lT_c) - \sum_k \sum_{i \in C_0} b_k^{(0,i)} \times g_{\text{EQ}}^{(0,i)}(lT_c - kT_f)$. Since the scalar product is irrelevant to an orthonormal transform (Parseval theorem), we have that

[1] $(\cdot)^T$ denotes the transpose operator. $(\cdot)^\dagger$ denotes the conjugate and transpose operator.

$\Lambda(\mathbf{b}^{(0)}) = \langle \widetilde{\mathbf{F}}\mathbf{e}, \widetilde{\mathbf{F}}\mathbf{K}^{-1}\mathbf{e}\rangle$ with $\widetilde{\mathbf{F}}$ being the block diagonal orthonormal matrix that has blocks all identical to the M-point discrete Fourier transform (DFT) matrix $\mathbf{F}$. If we assume the guard time to be sufficiently long such that $g_{\mathrm{EQ}}^{(0,i)}(nT_c)$ has support in $[0, MT_c)$, the vector $\mathbf{E} = \widetilde{\mathbf{F}}\mathbf{e}$ can be partitioned into nonoverlapping blocks equal to $\mathbf{E}_k = \mathbf{Y}_k - \sum_{i\in C_0} b_k^{(0,i)}\mathbf{G}_{\mathrm{EQ}}^{(0,i)}$, where

$$\begin{aligned}\mathbf{Y}_k &= \left[Y_k(f_0),\ldots,Y_k(f_{M-1})\right]^T = \mathrm{DFT}\{\mathbf{y}_k\},\\ \mathbf{G}_{\mathrm{EQ}}^{(0,i)} &= \left[G_{\mathrm{EQ}}^{(0,i)}(f_0),\ldots,G_{\mathrm{EQ}}^{(0,i)}(f_{M-1})\right]^T = \mathrm{DFT}\{\mathbf{g}_{\mathrm{EQ}}^{(0,i)}\}\end{aligned} \tag{15}$$

are the M-element vectors that are obtained by computing the M-point DFT at frequency $f_n = n/(MT_c)$, $n = 0,\ldots,M-1$, of the kth vector of samples $\mathbf{y}_k = [y_k(0),\ldots,y_k((M-1)T_c)]^T$, and of the ith equivalent signature code $\mathbf{g}_{\mathrm{EQ}}^{(0,i)} = [g_{\mathrm{EQ}}^{(0,i)}(0),\ldots,g_{\mathrm{EQ}}^{(0,i)}((M-1)T_c)]^T$.

It follows that

$$\Lambda(\mathbf{b}^{(0)}) = \langle \mathbf{E}, \widetilde{\mathbf{F}}\mathbf{K}^{-1}\widetilde{\mathbf{F}}^{\dagger}\mathbf{E}\rangle = \langle \mathbf{E}, \mathbf{R}^{-1}\mathbf{E}\rangle, \tag{16}$$

where we have used the identity $\widetilde{\mathbf{F}}^{-1} = \widetilde{\mathbf{F}}^{\dagger}$, and

$$\widetilde{\mathbf{F}}\mathbf{K}\widetilde{\mathbf{F}}^{\dagger} = E\left[\widetilde{\mathbf{F}}\mathbf{z}\mathbf{z}^T\widetilde{\mathbf{F}}^{\dagger}\right] = E\left[\mathbf{Z}\mathbf{Z}^{\dagger}\right] = \mathbf{R}. \tag{17}$$

Therefore, from (16), if we denote with $\mathbf{R}_{k,m}^{-1}$ the $M\times M$ block of indices (k,m) of $\mathbf{R}^{-1}$, the FD maximum likelihood receiver searches for the sequence of data symbols $\mathbf{b}^{(0)}$ (belonging to the desired user) that minimizes the log-likelihood function

$$\Lambda(\mathbf{b}^{(0)}) = \sum_{k=-\infty}^{\infty}\sum_{m=-\infty}^{\infty}\left[\mathbf{Y}_k - \sum_{i\in C_0} b_k^{(0,i)}\mathbf{G}_{\mathrm{EQ}}^{(0,i)}\right]^{\dagger} \times \mathbf{R}_{k,m}^{-1}\left[\mathbf{Y}_m - \sum_{n\in C_0} b_m^{(0,n)}\mathbf{G}_{\mathrm{EQ}}^{(0,n)}\right]. \tag{18}$$

Remarks 1. To compute the metric (18), we need to compute the DFT of each received frame (efficiently, via fast Fourier transform, FFT), and to estimate the channel frequency response, the impulse noise occurrence, and the correlation matrix of the impairment. This is treated in Section 5.

In (18), detection is jointly performed for the desired user's symbols, while all signals belonging to the other nodes are treated as interference whose FD correlation is included in the matrix $\mathbf{R}$ together with the correlation of the noise.

The metric can be easily extended to include a time-variant channel. The case, for instance, of a fast time-variant channel that is static only for a duration of frame can be captured in the metric (18) by changing $\mathbf{G}_{\mathrm{EQ}}^{(0,i)}$ into $\mathbf{G}_{\mathrm{EQ},k}^{(0,i)}$, that is, the frequency response of the channel for the kth frame.

The metric (18) provides a soft metric for the Viterbi channel decoder when convolutional codes are used. In the presence of impulse, noise some terms of (18) have negligible weight which corresponds to neglecting (puncturing) some of the trellis sections.

The DFT of the kth frame can be written as $\mathbf{Y}_k = \sum_{i\in C_0} b_k^{(0,i)}\mathbf{G}_{\mathrm{EQ}}^{(0,i)} + \mathbf{Z}_k$. The impairment multivariate process $\mathbf{Z}_k = [Z_k(f_0),\ldots,Z_k(f_{M-1})]^T$ has *time-frequency* correlation matrix equal to

$$\mathbf{R}_{k,m} = E\left[\mathbf{Z}_k\mathbf{Z}_m^{\dagger}\right] = \mathbf{F}\mathbf{K}_{k,m}\mathbf{F}^{\dagger}, \tag{19}$$

where $\mathbf{K}_{k,m}$ is the $M\times M$ matrix with entries $\kappa_{z|\alpha}((kM+n)T_c, (mM+l)T_c)$ for $n,l = 0,\ldots,M-1$, and $\mathbf{F}$ is the M-point DFT orthonormal matrix. In (18), $\mathbf{R}_{k,m}^{-1}$ denotes the $M\times M$ block of indices (k,m) of $\mathbf{R}^{-1}$, where $\mathbf{R}^{-1}$ is the inverse of the matrix $\mathbf{R}$ whose $M\times M$ block of indices (k,m) is $\mathbf{R}_{k,m}$. If $\mathbf{R}$ is block diagonal, for example, when we neglect the impairment correlation across frames, $\mathbf{R}_{k,k}^{-1}$ is equal to the inverse of the kth block, that is, equal to $(\mathbf{R}_{k,k})^{-1}$. As an example, if we consider independent noise samples, when the impulse noise hits a frame, $\mathbf{R}_{k,k}$ has diagonal elements that go to infinity. Then, $(\mathbf{R}_{k,k})^{-1}$ has diagonal elements that go to zero. Consequently, the corresponding additive terms in the metric (18) have zero weight.

4. SIMPLIFIED FD DETECTION ALGORITHMS

4.1. Simplified FD joint detector

To simplify the algorithm complexity, we neglect the temporal correlation of the impairment (MAI + noise) vector $\mathbf{Z}_k$, that is, we assume $\mathbf{R}_{k,m} = 0$ for $k \neq m$, and we denote $\mathbf{R}_{k,k}$ with $\mathbf{R}_k = E[\mathbf{Z}_k\mathbf{Z}_k^{\dagger}]$. Then, by dropping the terms that do not depend on the information symbols $\mathbf{b}_k^{(0)} = \{b_k^{(0,i)},\ i\in C_0\}$ that are transmitted in the kth frame by the desired user, the log-likelihood function simplifies to

$$\Lambda(\mathbf{b}_k^{(0)}) \sim -\mathrm{Re}\left\{\sum_{i\in C_0} b_k^{(0,i)}\mathbf{G}_{\mathrm{EQ}}^{(0,i)\dagger}\mathbf{R}_k^{-1}\left[\mathbf{Y}_k - \frac{1}{2}\sum_{n\in C_0} b_k^{(0,n)}\mathbf{G}_{\mathrm{EQ}}^{(0,n)}\right]\right\}. \tag{20}$$

We then make a decision on the transmitted symbols of frame k and user $u = 0$, as follows:

$$\hat{\mathbf{b}}_k^{(0)} = \arg\min_{\mathbf{b}_k^{(0)}}\{\Lambda(\mathbf{b}_k^{(0)})\}. \tag{21}$$

Therefore, according to (20) and (21), the FD receiver operates on a frame-by-frame basis and it exploits the frequency correlation of the impairment. We assume the correlation matrix to be full rank, otherwise pseudoinverse techniques can be used. Further, note that detection is jointly performed for all symbols that are simultaneously transmitted in a frame by the desired node. To obtain (20), we need to estimate $\mathbf{G}_{\mathrm{EQ}}^{(0,i)}$. The attractive feature with this approach is that the matched filter frequency response at a given frequency depends only on the channel response at that frequency. This greatly simplifies the channel estimation task. By exploiting the Hermitian symmetry of $\mathbf{G}_{\mathrm{EQ}}^{(0,i)}$, the estimation can be carried out only over $M/2$ frequency bins. A further simplification is obtained by observing that the Fourier transform of the equivalent channel of the desired user has significant energy only over a small fraction of the frequency bins, and only here channel estimation can be performed. Consequently, we can reduce the rank of the correlation matrix and combine only these frequency bins in the metric (20).

4.2. Iterative FD joint detector

The complexity of the *simplified FD joint detector* is still high because it increases exponentially with the number of symbols that are simultaneously transmitted by the desired user in a frame (equal to the number of assigned spreading codes). A possible way to simplify complexity is to search for the maximum of the metric in an iterative fashion. That is, we first detect symbol $\hat{b}_k^{(0,0)}$ by setting to zero all other symbols in $\Lambda(\mathbf{b}_k^{(0)})$. Then, we detect symbol $\hat{b}_k^{(0,1)}$ by setting $b_k^{(0,0)} = \hat{b}_k^{(0,0)}$ in $\Lambda(\mathbf{b}_k^{(0)})$. We detect new symbols using past decisions. Once all symbols are detected, we can rerun an iterative detection pass. This algorithm is similar in spirit to interference cancellation in CDMA systems [26] but it operates in the frequency domain.

4.3. FD full decorrelator

Another possibility is to perform detection of the symbols that belong to the desired node in a symbol-by-symbol fashion. That is, when we detect one symbol, we treat as interference both the signals of other users and the signals of the desired user that are associated to the other codes. Thus, the decision metric for the ith symbol of user 0 and frame k, can be derived similarly to (18) and (20), and it corresponds to

$$\Lambda(b_k^{(0,i)}) \sim -\operatorname{Re}\left\{ b_k^{(0,i)} \mathbf{G}_{\mathrm{EQ}}^{(0,i)\dagger} \left(\mathbf{R}_k^{(0,i)}\right)^{-1} \left[\mathbf{Y}_k - \frac{1}{2} b_k^{(0,i)} \mathbf{G}_{\mathrm{EQ}}^{(0,i)} \right] \right\}, \tag{22}$$

where $\mathbf{R}_k^{(0,i)}$ is the correlation matrix of the impairment (MAI + ICI + noise + other codes) that is seen by the symbol associated to the ith signature code of frame k:

$$\mathbf{R}_k^{(0,i)} = E\left[\mathbf{E}_k^{(0,i)} \mathbf{E}_k^{(0,i)\dagger}\right], \quad \mathbf{E}_k^{(0,i)} = \mathbf{Z}_k + \sum_{\substack{c \in C_0 \\ c \neq i}} b_k^{(0,c)} \mathbf{G}_{\mathrm{EQ}}^{(0,c)}. \tag{23}$$

This algorithm requires a matrix inversion for each code. When all codes are assigned, its complexity is lower than the FD joint detector when the channel and interference remain static for a long time, such that the inverse matrices can be computed once. A way to reduce further its complexity is to use a rank reduction approach, that is, we process only the frequency bins that exhibit sufficiently high energy. Finally, this algorithm becomes identical to the joint detector algorithm if the desired user deploys a single code.

5. PRACTICAL IMPLEMENTATION ALGORITHMS

The practical implementation of the above algorithms requires to estimate the frequency response of the desired user channel and the impairment correlation matrix. In this paper we propose to use a pilot channel (a Walsh code) as shown in Figure 4. We assume, instead, perfect frame synchronization with the desired user whose practical implementation is discussed in [27].

Assuming packet transmission of duration N frames, (super-frame), the pilot channel spans N frames, that is, it

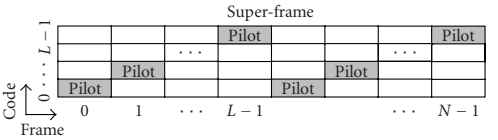


Figure 4: Super-frame format with pilot channel.

corresponds to a training sequence of length N symbols that we assume to have $\{-1, 1\}$ alphabet.

In order to better sound the channel, we propose to change the assigned Walsh code (pilot code) at each new frame (Figure 4). If we assume full-rate transmission, that is, a user is allocated to all $L-1$ Walsh codes, channel sounding is done in a cyclic manner as follows. The pilot channel uses the Walsh code 0 in the first frame of the super-frame, while the remaining $L-1$ codes are used for data transmission. Then, it uses code 1 in the second frame, and so on in a cyclic manner as Figure 4 shows. Distinct users deploy distinct pilot codes.

To improve the performance of the estimators, we consider the use of an iterative approach where we first take into account only the knowledge of the pilot symbols. Then, after detection/channel decoding, we rerun an estimation pass by exploiting the knowledge of all detected symbols.

We assume the user channel and the MAI vector to be stationary over the transmission of a super-frame. This holds true, for instance, assuming users with identical frame duration and spreading code length. However, we point out that during the detection stage the algorithms that we describe allow to perform adaptation to channel and MAI variations in a data decision-directed mode.

While the background noise is stationary, the impulse noise is in general not stationary such that the estimation of its correlation is not feasible. To solve this problem, we assume that conditional on its occurrence, the overall noise is locally stationary. This means that the correlation of the impulse noise can be estimated by averaging over the time windows where it is present. Clearly, the first thing to do is to locate the impulse noise.

5.1. Locating the impulse noise

To simplify the task, the estimation of the impulse noise occurrence is done on a frame-by-frame basis by making a comparison between the average received signal energy computed over a super-frame $E_{\mathrm{SF}} = \sum_{k=0}^{N-1} \mathbf{Y}_k^\dagger \mathbf{Y}_k / N/M$, and the energy computed over a frame $E_F(k) = \mathbf{Y}_k^\dagger \mathbf{Y}_k / M$.

To simplify further the algorithms, in the Viterbi decoding stage, we disregard the frames of index k for which $E_F(k)/E_{\mathrm{SF}} > E_{\mathrm{th}}$ for a given threshold E_{th}. This corresponds to puncturing the trellis sections that are associated with bits that are hit by impulse noise. This is because in correspondence to a noise spike the coded bit statistics are quite unreliable and it is better not to use them.

Finally, the adaptive estimations of the channel and the MAI-plus-background-noise correlation matrix are done neglecting the frames that are hit by impulse noise.

5.2. FD channel estimation

We implement FD channel estimation independently over the DFT output subchannels (frequency bins) using a one-tap recursive least-square (RLS) algorithm [18]. We approximate the equivalent channel frequency response for the ith code of the desired user (user 0) as follows:

$$\hat{G}_{\mathrm{EQ}}^{(0,i)}(f_n) \approx W^{(0,i)}(f_n)\hat{H}(f_n), \quad i=0,\ldots,L-1,\ n=0,\ldots,M-1, \tag{24}$$

where $W^{(0,i)}(f_n)$ denotes the M-point DFT (at frequency f_n) of the pilot signature code that comprises the front-end filter. The channel estimate $\hat{H}(f_n)$ is obtained via a one-tap RLS algorithm that uses the following error signal for the kth frame:

$$e_k(f_n) = Y_k(f_n) - \hat{H}_{k-1}(f_n)W^{(0,\,\mathrm{mod}\,(k,L))}(f_n)b_{\mathrm{TR},k}, \tag{25}$$

where $b_{\mathrm{TR},k}$, $k=0,\ldots,N-1$, is the known training symbol that is transmitted in the kth frame by the desired user, $\hat{H}_k(f_n)$ is the channel estimate for the kth iteration, and $\mathrm{mod}(\cdot,\cdot)$ denotes the remainder of the integer division (recall that the Walsh code that is associated to the pilot channel is cyclically updated frame after frame).

5.3. FD estimation of the MAI-plus-noise correlation matrix

Once we have obtained an estimate of the equivalent signature code frequency response $\hat{\mathbf{G}}_{\mathrm{EQ}}^{(0,i)}$, the MAI-plus-noise correlation matrix that is required in algorithm (20) can be estimated via time-averaging the error vector that is defined as $\hat{\mathbf{E}}_k = \mathbf{Y}_k - b_{\mathrm{TR},k}\hat{\mathbf{G}}_{\mathrm{EQ}}^{(0,\,\mathrm{mod}\,(k,L))}$:

$$\hat{\mathbf{R}} = \frac{1}{N}\sum_{k=0}^{N-1}\hat{\mathbf{E}}_k\hat{\mathbf{E}}_k^{\dagger}. \tag{26}$$

To introduce a tradeoff between the effects of noise and the effects of the MAI, we can perform diagonal loading of the estimated correlation matrix which also assures that the correlation matrix is full rank.

5.4. FD estimation of the ICI correlation matrix

Under the assumption of independent zero-mean symbols, and MAI uncorrelated from the desired user signal, the correlation of the interference that is seen by the ith signature code of the desired user can be written as

$$\hat{\mathbf{R}}^{(0,i)} = \hat{\mathbf{R}} + \hat{\mathbf{R}}_{\mathrm{ICI}}^{(0,i)}; \tag{27}$$

that is, as the sum of the correlation matrix of the MAI-plus-noise and the correlation matrix of the ICI experienced by the ith code of the desired user. After channel estimation, we can obtain an estimate of the ICI correlation matrix (assuming unit power data symbols) as follows:

$$\hat{\mathbf{R}}_{\mathrm{ICI}}^{(0,i)} = \sum_{c\in C_0,\,c\neq i}\hat{\mathbf{G}}_{\mathrm{EQ}}^{(0,c)}\hat{\mathbf{G}}_{\mathrm{EQ}}^{(0,c)\dagger}. \tag{28}$$

5.5. Data-aided iterative estimation with feedback from the channel decoder

The estimators can be improved by using a data decision-aided approach. That is, we can iteratively refine the estimation as data decisions are made. This turns out to be effective when the desired user transmits at high rate, and consequently the ICI is high. At the first pass, we estimate the channel and the correlation matrix assuming knowledge of only the pilot symbols. Then, in a second pass, we rerun estimation of the channel and the correlation matrix using the data decisions made at the first pass. In particular, if we assume to have detected all symbols in a super-frame of length N frames, we can rerun RLS channel estimation using the following error signal:

$$e_k(f_n) = Y_k(f_n) - \hat{H}_{k-1}(f_n)\sum_{c\in C_0}W^{(0,c)}(f_n)\hat{b}_k^{(0,c)}, \tag{29}$$

where $\{\hat{b}_k^{(0,c)},\ c\in C_0\}$ are all detected symbols plus the pilot symbol that is transmitted in the kth frame by the desired user. To re-estimate the correlation matrix of the MAI-plus-noise, we can implement (26) using the following error vector:

$$\hat{\mathbf{E}}_k = \mathbf{Y}_k - \sum_{c\in C_0}\hat{b}_k^{(0,c)}\hat{\mathbf{G}}_{\mathrm{EQ}}^{(0,c)}, \tag{30}$$

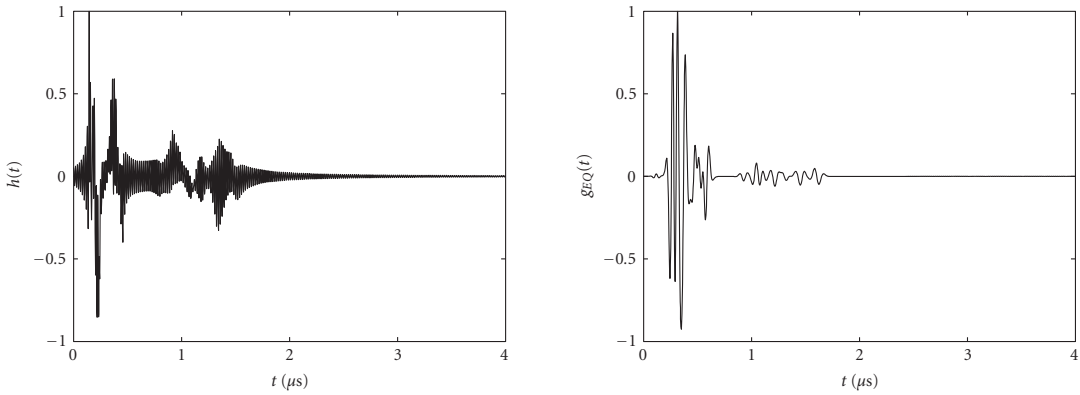
where $\hat{\mathbf{G}}_{\mathrm{EQ}}^{(0,c)}$ are the new channel estimates. Similarly, we can re-estimate the correlation matrix of the ICI-plus-noise according to (28) using, however, the new channel estimates.

The data decisions that are used in the above algorithms can be provided by the detector, or by the channel decoder. In the latter case, we just need to use a standard soft-input hard-output Viterbi decoder followed by re-encoding and interleaving, as Figure 1 shows. Further, to minimize the correlation with previous estimates, we can partition the super-frame into two parts so that we can obtain two estimates for the channel and the correlation matrix. The former estimates that are used for data detection in the first half of the super-frame are obtained running training with data decisions belonging to the second half of the super-frame, and vice versa.

6. PERFORMANCE RESULTS

6.1. System parameters

The performance of the system is assessed via simulations. We assume a frame duration $T_f = 4.096$ microseconds and a monocycle of duration $D \approx 126$ nanoseconds (Figure 3). The -20 dB bandwidth is equal to about 30 MHz. This choice has been made via experimental trials [4]. The guard time is



(a) Realization of channel response

(b) Realization of equivalent response

Figure 5: Examples of statistical channel realization (a) and equivalent impulse response (b).

$T_g = 2.048$ microseconds. The monocycle (at the transmitter and receiver front-end) and the channel are simulated with a sampling period of 2 nanoseconds (63 samples per monocycle). Then, the front-end filter output signal is downsampled to obtain a period T_c=16 nanoseconds. Thus, we collect $M = 256$ samples per frame and we use an FFT of size 256. The spreading codes have length $L = 16$ with a chip period $T = 128$ nanoseconds. The codes are obtained by the chip-by-chip product of the 16 Walsh-Hadamard codes and a random code for each user to be multiplexed. One code is reserved for training. We consider binary data symbols. Furthermore, a bit-interleaved convolutional code of rate 1/2 and memory 4 is used. The transmission rate can be adjusted according to the number of signature codes that are allocated to each user. The super-frame spans $N = 540$ frames (2.21 milliseconds). Consequently, the coded packet has length from a minimum of 540 bits with single code, to a maximum of 8100 coded bits with fulls-rate transmission (15 codes). A block interleaver that spans 540 frames is used. With these parameters, the uncoded transmission rate ranges from 244 kbit/s to 3.66 Mbit/s, while the net rate with coding is half of that. Clearly, it can be increased with higher level PAM or longer spreading codes, but we have made this choice to keep the simulation runtime within tolerable values.

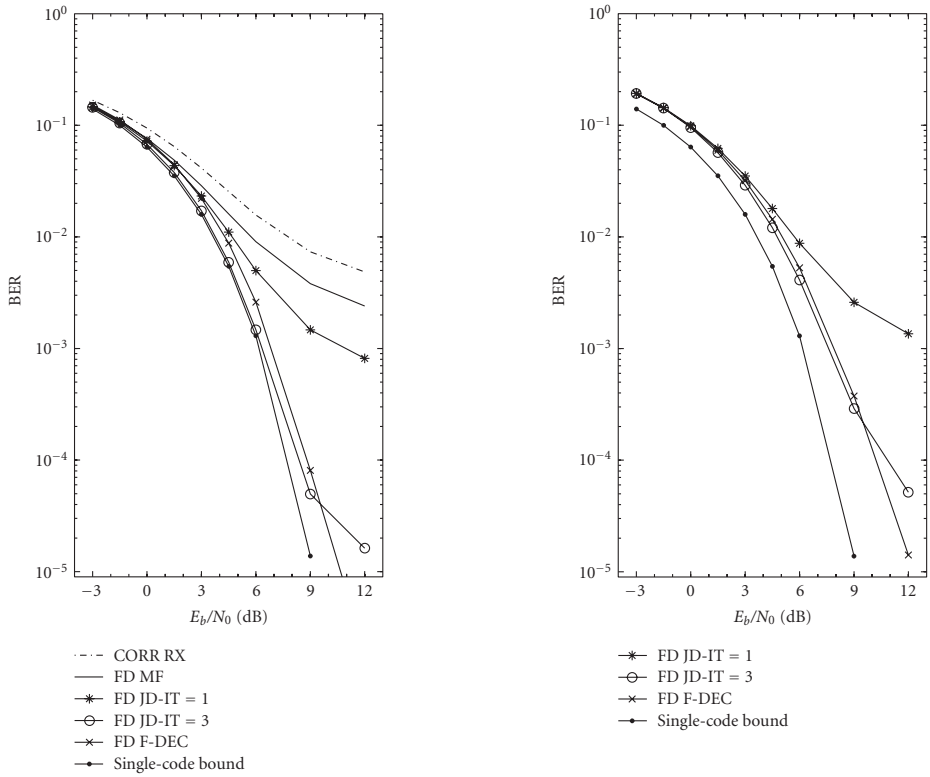
6.2. Channel parameters

Starting from the channel model in Section 2.3, we set B_1=0 and $B_2 = 55$ MHz. Having in mind an indoor environment where the number of paths is typically high, we fix for the underlying Poisson process an intensity $\Lambda = 1/15\,\mathrm{m}^{-1}$, that is, one reflector every 15 m in average. The first one is set at distance 30 m with $g_1 = 1$, while the maximum path distance is 300 m. Finally, we choose $K = 1$, $\alpha_0 = 10^{-5}\,\mathrm{m}^{-1}$, $\alpha_1 = 10^{-9}$ s/m. In Figure 5(a), we plot an example of channel realization while in Figure 5(b) we plot the equivalent channel response $g_{\mathrm{EQ}}(t) = g_M * h^{(u)} * g_{\mathrm{FE}}(t)$. The equivalent response is significantly compressed because the monocycle filters out the low-frequency components that are responsible for longer channel delays according to model (5). The channel is assumed to be static for the duration of a super-frame equal to 2.21 milliseconds, and then it randomly changes. In the simulations we truncate the channel impulse responses to 4 microseconds. However, we use a guard time of only 2.048 microseconds. The performance degradation that is due to the interframe interference that is generated by the tail of the channel is negligible.

6.3. Full-rate single-user performance

In Figure 6, we report the bit-error-rate (BER) performance before channel decoding averaged over at least 1500 PL grid topologies (channel realizations) as a function of E_b/N_0, that is, the energy per bit at the front-end output, over the noise spectral density. The additive background noise is white Gaussian. We point out that we normalize the channel such that the received bit energy is constant for all channel realizations. This choice removes the fading effect which is appropriate in the PL context differently, for instance, from the mobile wireless context [18]. A single full-rate user that deploys all available 16 Walsh codes is present.

In Figure 6(a), the performance with ideal channel knowledge is shown for the baseline correlation receiver (CORR RX), the FD-matched filter detector that takes into account only the colored noise (FD MF), the FD detector with single-code transmission (single code), the FD joint iterative detector (FD JD-IT) with up to 3 iterations, and finally the FD full decorrelator (FD F-DEC). All receivers significantly improve performance compared to the baseline correlation receiver. Since the front-end filter (matched to



(a) Uncoded—ideal channel estimate

(b) Uncoded—practical channel estimate

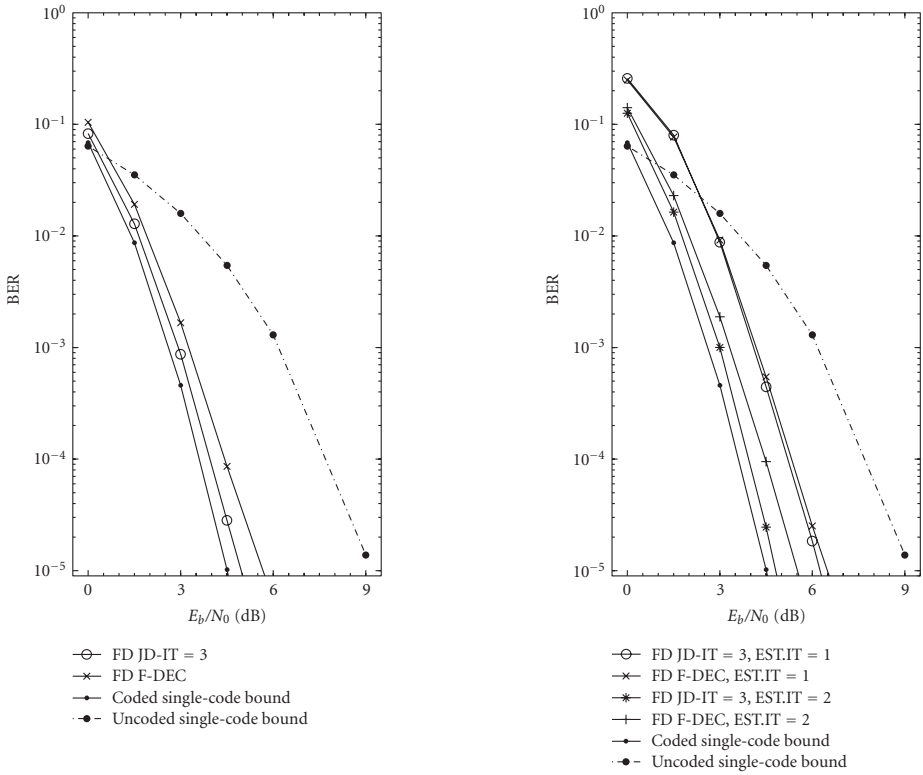
Figure 6: Average BER with one full-rate user without channel coding in AWGN.

the monocycle) colors the noise, the FD MF detector that takes it into account improves performance compared to the correlation receiver. However, the severely dispersive channel introduces intercode interference, thus an error floor is visible. If we use the FD full decorrelator, we get a significant performance gain. Here, to simplify complexity, we actually combine only the frequency bins that have energy above 1% of the maximum. Near ideal performance (single-code performance bound) is achieved with the FD iterative detector with only 3 iterations for E_b/N_o below 9 dB.

Figure 6(b) shows that with practical channel estimation (with the method in Section 5.2), the BER performance is within 1.5 dB from the ideal curves.

In Figure 7(a), we report BER at the output of the soft-input Viterbi decoder assuming ideal channel estimation, while in Figure 7(b) we assume practical channel estimation. With channel coding, the performance is improved. The curves with practical channel estimation are very close to the ideal curves. Here, curves labeled with EST.IT = 2 assume two channel estimation passes using hard feedback from the decoder (as explained in Section 5.5). With 3 iterative detection passes, we are within 0.5 dB from the single-code bound that corresponds to single code transmission and ideal channel estimation. The simplified F-DEC is within 0.5 dB from the iterative detector.

In Figure 8(a), we assume the presence of impulse noise and ideal channel estimation, while in Figure 8(b) we assume practical channel estimation. We report the BER both with channel coding (Cod) and without it (Uncod). In the simulation the impulse noise is generated according to the two-term Gaussian model [21, 22] whose probability density function can be defined as $p_\eta(a) = (1-\varepsilon)N(0,\sigma_1^2)+\varepsilon N(0,\sigma_2^2)$. The first term gives the zero-mean Gaussian background noise with variance σ_1^2. The second term represents the impulse component and it has variance $\sigma_2^2 = 100\sigma_1^2$. The occurrence probability is $\varepsilon = 0.01$. To stress the system performance, when an impulse occurs, we assume the Gaussian process with variance σ_2^2 to last for a period of time equal to 4 frames [22]. The spectrum of this noise can be shaped to increase its low-frequency components to reflect measured scenarios. However, if we do not do so, we get the worst-case scenario especially in our system where the transmission spectrum does



(a) Coded—ideal channel estimate

(b) Coded—practical channel estimate

Figure 7: Average BER with one full-rate user and with channel coding in AWGN.

not occupy the low frequencies. The position of the noise spikes within a super-frame is estimated. The results show that a performance degradation is introduced compared to the AWGN case. However, if we use the proposed modified Viterbi algorithm (curves labeled with Erasure), the performance comes close to that of the single code in AWGN. As Figure 8(b) shows a second channel estimation pass with feedback from the decoder yields near-ideal performance.

6.4. Multiuser performance with full-rate users

Users multiplexing can be done by partitioning the L Walsh codes among the users. To stress the system, we have assumed all users to be at full rate, that is, they deploy all 16 Walsh-Hadamard codes. As explained in Section 2.1, a random code is also used on top of the Walsh codes. In Figure 9(a), we assume the presence of one interferer with ideal channel/correlation estimation while in Figure 9(b) we assume the presence of three interferers with practical estimation. The overall interferers power equals the desired user power. The channels are independently drawn according to the statistical model, however, they are assumed to be static for the whole duration of a super-frame. The additive background noise is white Gaussian. Users are asynchronous with a random starting phase. Figure 9 shows that although there is some performance penalty compared to single-code single-user case due to the MAI, the FD detection algorithms allow to keep such a penalty small. This can be explained by the fact that the random codes and the multiple-access channel diversity introduce some degrees of freedom that can be exploited in the frequency domain by the interference cancellation algorithms. The iterative detector with 3 iterations performs better than the simplified full decorrelator for E_b/N_o smaller than 9 dB. Then, an error floor appears, though it may be reduced with further iterations. The simple bit-interleaved memory-4 convolutional code allows to significantly improve the BER performance.

With practical estimation (Figure 9(b)) of the channel, the BER performance exhibits an error floor at the first

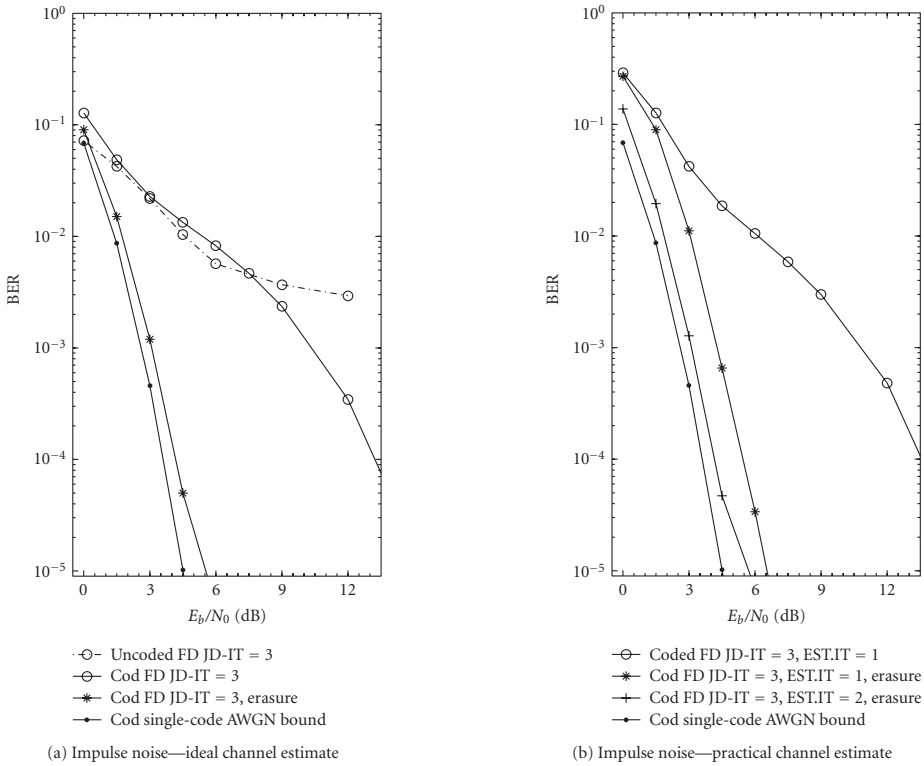


Figure 8: Average BER with one full-rate user and with impulse noise.

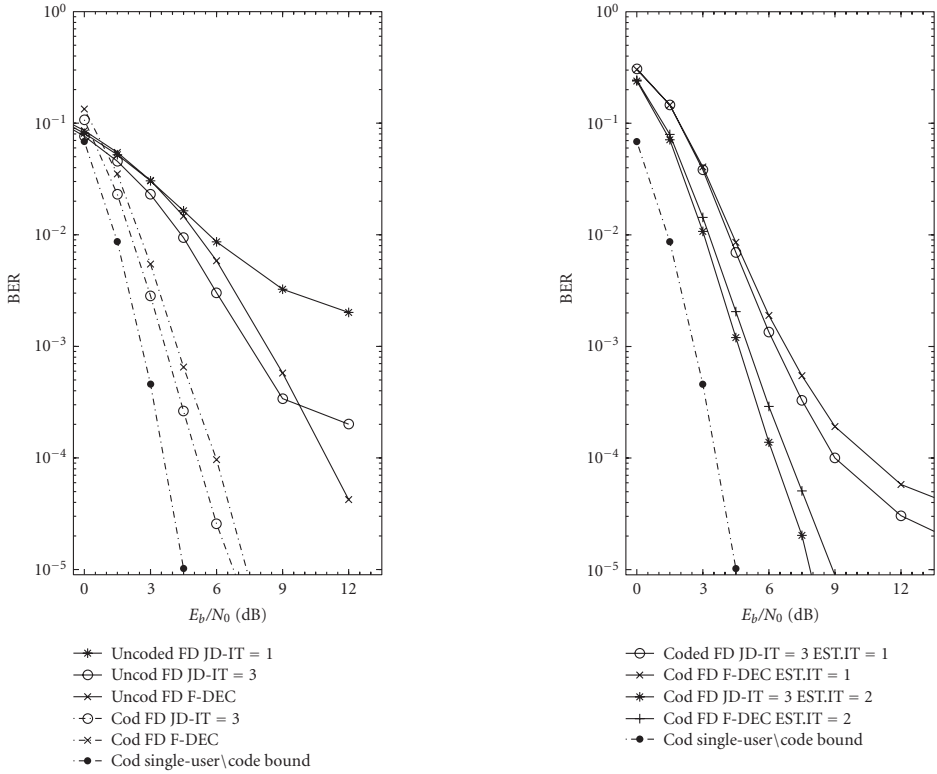
estimation pass (curves labeled with EST.IT = 1). Here, we assume to first run detection and channel decoding without performing MAI cancellation. For the JD-IT scheme, we run 3 iterations. Then, for the curves labeled with EST.IT = 2 we rerun a second channel estimation pass followed by practical estimation of the MAI correlation matrix using hard feedback from the convolutional decoder. Now, the practical curves are within about 1 dB from the curves with ideal channel/correlation estimation.

7. CONCLUSIONS

In this paper, we have investigated the application of wideband impulse modulation combined with CDMA for PL communications. This modulation approach requires a simple baseband time-domain implementation of the transmitter and the receiver. A key aspect is that the energy of each information symbol is spread over a wideband (yielding a low-spectral density signal) contrary to narrowband or multicarrier architectures that can be seen as a bank of narrowband systems. This allows to exploit the channel frequency diversity and to be robust to narrowband interference. Further, time diversity is exploited via the CDMA signature code together with the bit-interleaved convolutional code. This yields robustness to impulse noise.

Improved performance, relatively to the baseline correlation receiver, can be obtained with a maximum likelihood FD joint detector. This receiver adapts to channel time variations and to asynchronous impulse noise, and mitigates the detrimental effect of the ICI and MAI that are generated by the time-dispersive channel and that are significant in full-rate transmission. With certain simplifications we have derived a simplified FD joint detector, an FD iterative detector, and an FD interference decorrelator. They all include the capability of rejecting the ICI/MAI but have different levels of performance and implementation complexity. In particular, the FD full decorrelator receiver has the lowest complexity especially when we process a subset of the available frequency bins.

Algorithms for the FD estimation of the channel and of the correlation of the interference have also been described. Channel estimation can be performed independently over the frequency bins with one-tap RLS adaptive filters. To improve the performance of the estimators we have used a data



(a) 1 Interferer-ideal channel/correlation estimate

(b) 3 Interferers-practical channel/correlation estimate

Figure 9: Average BER with (a) one and three full-rate interferers (b) (worst-case scenario). (a) Ideal channel/correlation estimation. (b) Practical estimation of the channel/correlation.

aided approach with hard feedback from the Viterbi decoder. Few iterations have proved to be effective.

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Calendar of Events

Year	Date	Event	Location	EURASIP Involvement	Chairperson/Information
2009	June 3–5	7th International Workshop on Content-Based Multimedia Indexing (CBMI 2009)	Chania, Crete, Greece	Cooperation	Yannis Avrithis http://www.cbmi2009.org/
	June 11–12	2nd International Workshop on Cross-Layer Design (IWCLD 2009)	Mallorca, Spain	Cooperation	Lagunas, Nossek, Yuan http://www.iwcld2009.org
	June 17–19	16th International conference on Systems, Signals and Image Processing (IWSSIP '09)	Chalkida, Greece	Cooperation	Stamatis Voliotis http://www.teihal.gr/iwssip09/
	June 18–21	The 5th Conference on Speech Technology and Human Computer Dialogue	Constanta, Romania	Cooperation	Corneliu Burileanu www.sped2009.ro
	June 22–24	Crowncom 2009	Hannover, Germany	Cooperation	Thomas Kaiser http://www.crowncom2009.org/
	June 25–27	Nonlinear Speech processing (Nolisp '09)	Vic, Barcelona, Spain	Cooperation	Marcos Faundez Zanuy http://nolisp2009.uvic.cat/
	June 5–7	International conference on signal processing (DSP 2009)	Santorini, Greece	Cooperation	Tanos Skodras http://www.dsp2009.org/
	August 24–28	17th European Signal Processing Conference (EUSIPCO 2009)	Glasgow, UK	Sponsor	Bob Stewart http://www.eusipco2009.org
	September 16–18	6th International Symposium on Image and Signal Processing ans Analysis (ISPA 2009)	Salzburg, Austria	Cooperation	Peter Zinterhof, Sven Locaric http://www.isispa.org
2010	June	17th International conference on Systems, Signals and Image Processing (IWSSIP '10)	Rio de Janeiro, Brasil	Cooperation	
	June	7th international Symposium on Communication Systems, Networks and Digital Signal Processing (CSNDSP 2010)	Newcastle, UK	Cooperation	Fary Ghassemlooy http://www.csndsp.com/
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The 2009 European Signal Processing Conference (EUSIPCO-2009) is the seventeenth in a series of conferences organised by the European Association for Signal, Speech, and Image Processing (EURASIP, www.eurasip.org). The conference will be held in the Royal Concert Hall in the heart of the city of Glasgow, Scotland.
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- Multimedia signal processing, speech processing and coding
- Image processing and video coding
- Medical imaging and image analysis
- Signal detection and estimation
- Sensor array, multichannel and multidimensional signal processing
- Signal processing for communications and wireless networks
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Electronic submission of full papers	February 6, 2009
Notification of Acceptance	April 30, 2009
Submission of camera-ready papers	May 29, 2009

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- Communication systems and networks
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Cardiff, the capital of Wales, UK, in the proximity of where the first wireless communication by Marconi took place, has been the host for many international and social events, conferences, and workshops recently for its attractive natural resources and many constitutional developments. In 2009 delegates will be brought right up to date in an area of rebirth, Cardiff Bay, the heart of Europe's youngest capital. The history of Cardiff, however, goes back to the days of the Roman Empire; the remains of the Roman stone fortifications can be seen within the walls of Cardiff Castle. Cardiff's reputation as one of Britain's six cities of elegance arises from its remarkable civic centre, regarded as being world-ranking. The venue is located in the city centre with its museums, theatres, hotels, large shopping district and Cardiff Castle. It is also a short distance from Cardiff Bay, a major regeneration area. The workshop will contain a number of special sessions organised by internationally recognised experts. The programme will also include presentations of new results in lecture and poster sessions, together with plenary sessions delivered by eminent scientists.

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During the 50 years of activity ELMAR symposium became a significant scientific conference in the field of multimedia communications, image and video processing, navigation systems, speech and audio processing, telecommunications, wireless commununications, electronics in marine, naval architecture, sea ecology, and other advanced research areas. Besides, every year ELMAR symposium gathers specialists of various kinds (government representatives, navy, industry, universities and various business people from the region) to discuss the most recent issues and contribute to appropriate market development in Croatia.

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  - Title to be announced soon.
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Special Issue on Atypical Speech

CALL FOR PAPERS

Research in speech processing (e.g., speech coding, speech enhancement, speech recognition, speaker recognition, etc.) tends to concentrate on speech samples collected from normal adult talkers. Focusing only on these "typical speakers" limits the practical applications of automatic speech processing significantly. For instance, a spoken dialogue system should be able to understand any user, even if he or she is under stress or belongs to the elderly population. While there is some research effort in language and gender issues, there remains a critical need for exploring issues related to "atypical speech". We broadly define atypical speech as speech from speakers with disabilities, children's speech, speech from the elderly, speech with emotional content, speech in a musical context, and speech recorded through unique, nontraditional transducers. The focus of the issue is on voice quality issues rather than unusual talking styles.

In this call for papers, we aim to concentrate on issues related to processing of atypical speech, issues that are commonly ignored by the mainstream speech processing research. In particular, we solicit original, previously unpublished research on:

- Identification of vocal effort, stress, and emotion in speech
- Identification and classification of speech and voice disorders
- Effects of ill health on speech
- Enhancement of disordered speech
- Processing of children's speech
- Processing of speech from elderly speakers
- Song and singer identification
- Whispered, screamed, and masked speech
- Novel transduction mechanisms for speech processing
- Computer-based diagnostic and training systems for speech dysfunctions
- Practical applications

Authors should follow the EURASIP Journal on Audio, Speech, and Music Processing manuscript format described at the journal site http://www.hindawi.com/journals/asmp/. Prospective authors should submit an electronic copy of their complete manuscript through the journal Manuscript Tracking System at http://mts.hindawi.com/, according to the following timetable:

Manuscript Due	April 1, 2009
First Round of Reviews	July 1, 2009
Publication Date	October 1, 2009

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Special Issue on

Advances in Quality and Performance Assessment for Future Wireless Communication Services

CALL FOR PAPERS

Wireless communication services are evolving rapidly in tandem with developments and vast growth of heterogeneous wireless access and network infrastructures and their potential. Many new, next-generation, and advanced future services are being conceived. New ideas and innovation in performance and QoS, and their assessment, are vital to the success of these developments. These should be open and transparent, with not only network-provider-driven but also service-provider-driven and especially user-driven, options on management and control to facilitate always best connected and served (ABC&S), in whatever way this is perceived by the different stake holders. To wireless communication services suppliers and users, alike the complexity and integrability of the immense, diverse, heterogeneous wireless networks' infrastructure should add real benefits and always appear as an attractive user-friendly wireless services enabler, as a wireless services performance enhancer and as a stimulant to wireless services innovation. Effecting the integration of services over a converged IP platform supported by this diverse and heterogeneous wireless infrastructure presents immense QoS and traffic engineering challenges. Within this context, a special issue is planned to address questions, advances, and innovations in quality and performance assessment in heterogeneous wireless service delivery.

Topics of interest include, but are not limited to:

- Performance evaluation and traffic modelling
- Performance assessments and techniques at system/flow level, packet level, and link level
- Multimedia and heterogeneous service integration-performance issues, tradeoffs, user-perceived QoS, and quality of experience
- Network planning; capacity; scaling; and dimensioning
- Performance assessment, management, control, and solutions: user-driven; service-provider-driven; network-provider-driven; subscriber-centric and consumer-centric business model dependency issues
- Wireless services in support of performance assessment, management, and control of multimedia service delivery

- Performance management and assessment in user-driven live-access network change and network-driven internetwork call handovers
- Subscriber-centric and consumer-centric business model dependency issues for performance management, control, and solutions
- Simulations and testbeds

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Manuscript Due	August 1, 2009
First Round of Reviews	November 1, 2009
Publication Date	February 1, 2010

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Special Issue on
Image Processing and Analysis in Biomechanics

CALL FOR PAPERS

Computational methodologies of signal processing and analysis based on 1D-4D data are commonly used in different applications in society. In particular, image processing and analysis methodologies have enjoyed increased deployment in automated recognition, human-machine interfaces, computer-aided diagnostics, robotics surgery, and biomechanics analysis.

Image processing and analysis is fundamentally a multidisciplinary area, combining elements of informatics, mathematics, statistics, psychology, mechanics and physics, among others. One of the more important applications of image processing and analysis can be found in medical imagery, which continually promotes new research and development. Present trends include using statistical or physical procedures on medical images in order to have different objectives, such as organ segmentation, shape reconstruction, motion and deformation analysis, organ registration and comparison, virtual reality, computer-assisted therapy, or biomechanic analysis and simulation.

The research related with analysis and simulation of biomechanical structures has been a source of many challenging problems, involving geometric modeling, numerical modeling, biomechanics, material models for living tissues, experimental methodologies, and mechanobiology, as well as their application in clinical environments. A critical component for true realistic biomechanical analysis and simulations is to obtain accurately, from images, the geometric data and the behavior of the desired structures. For that, the use of automatic, efficient, and robust techniques of image processing and analysis is required.

The main objective of this Special Issue on *Image Processing and Analysis in Biomechanics* is to bring together recent advances in the field. Topics of interest include, but are not limited to:

- Signal processing in biomechanical applications
- Data interpolation, registration, acquisition and compression in biomechanics
- Segmentation of objects in images for biomechanical applications
- 3D reconstruction of objects from images for biomechanical applications
- 2D/3D tracking and object analysis in images for biomechanical applications
- 3D vision in biomechanics
- Biomechanical applications involving image processing and analysis algorithms

- Virtual reality in biomechanics
- Software development for image processing and analysis in biomechanics

Before submission authors should carefully read over the journal's Author Guidelines, which are located at http://www.hindawi.com/journals/asp/guidelines.html. Authors should follow the EURASIP Journal on Advances in Signal Processing manuscript format described at the journal site http://www.hindawi.com/journals/asp/. Prospective authors should submit an electronic copy of their complete manuscript through the journal Manuscript Tracking System at http://mts.hindawi.com/, according to the following timetable:

Manuscript Due	May 1, 2009
First Round of Reviews	August 1, 2009
Publication Date	November 1, 2009

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Special Issue on

Advanced Signal Processing for Cognitive Radio Networks

CALL FOR PAPERS

Cognitive radio is widely expected to usher in the next wave in wireless communications. In December 2003, the Federal Communications Commission (FCC) of the US government issued authorized cognitive radio techniques for spectrum sharing/reusing and approved the use of fixed and mobile services in TV bands. In October 2008, the FCC further approved the use of mobile white space devices in TV bands, and many governments worldwide have also moved to support this new spectrum usage model. This has been accompanied recently by a significant upsurge in academic research and application initiatives, such as the IEEE 802.22 standard on wireless regional area networks (WRANs) and the Wireless Innovation Alliance including Google and Microsoft as members, which advocates unlocking the potential in the "white space" of television bands.

However, cognitive radio networking is still in the early stages of research and development. To achieve full "cognition" and reliable communication over a wireless network, there are still tremendous technical, economical, and regulatory challenges. Signal processing plays a major role in cognitive radio networks. The aim of this special issue is to present a collection of high-quality research papers in advanced signal processing for cognitive radio including theoretical studies, algorithms, protocol design, as well as architectures, platforms, and prototypes which use advanced signal processing techniques. Topics of interest include, but are not limited to:

- Advanced spectrum sensing techniques and protocol support
- Cooperative spectrum sensing and communication
- Resource allocation for spectrum sharing
- Exploiting multiantennas for spectrum sharing
- Channel and environment learning techniques for cognitive radio
- Advanced coding and modulation for cognitive radio
- Information theory for cognitive radio
- Multiuser spectrum access techniques
- Security issues in cognitive radio networks
- Multimedia transmission over cognitive radio networks

- Optimization for bandwidth utilization
- Cognitive radio prototypes and test beds

Before submission authors should carefully read over the journal's Author Guidelines, which are located at http://www.hindawi.com/journals/asp/guidelines.html. Prospective authors should submit an electronic copy of their complete manuscript through the journal Manuscript Tracking System at http://mts.hindawi.com/, according to the following timetable:

Manuscript Due	May 1, 2009
First Round of Reviews	August 1, 2009
Publication Date	November 1, 2009

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Special Issue on

Filter Banks for Next-Generation Multicarrier Wireless Communications

CALL FOR PAPERS

Digital filter banks find various good applications in communications signal processing. In general, they can be used to obtain very sharp frequency selectivity to isolate different communications frequency channels from each other and from interfering spectral components. This can be done in a very flexible and dynamic manner. Thus, filter banks constitute a very powerful generic tool for software-defined radios and spectrally agile communication systems.

The theoretical capacity limits in communications can be approached by multicarrier techniques. With radio channels, multicarrier techniques can be combined with multi-antenna transmitters and receivers to provide efficiency. Existing or planned transmission systems rely on the OFDM technique to reach these goals. However, OFDM has a number of drawbacks, such as the use of the cyclic prefix to cope with the channel impulse response which results in a loss of capacity and the requirement of block processing to maintain orthogonality among all the subcarriers. Furthermore, the leakage among frequency subbands has a serious impact on the performance of FFT-based spectrum sensing and OFDM-based cognitive radio in general.

So far, some attempts have been made to introduce filter bank multicarrier (FBMC) in the radio communications arena, in particular, the isotropic orthogonal transform algorithm (IOTA). However, the full exploitation and optimization of FBMC techniques in the context of radio evolution have not been considered sufficiently. Consequently, advances in communication aspects of FBMC are still required to make it useful for future radio systems.

This has motivated advanced research in the European ICT project PHYDYAS, which supports this special issue. Topics of interest include, but are not limited to:

- Filter bank-based multicarrier transmission and prototype filter design
- Filter bank-based signal processing for other communication waveforms
- Filter bank applications in software-defined radio
- Data-aided and blind techniques for synchronization and channel estimation
- Preamble and pilot-pattern design
- Equalization and demodulation

- FBMC MIMO techniques and beamforming
- Radio scene spectrum analysis and cognitive radio
- Interference management
- Interlayer optimization and FBMC-specific scheduling
- Filter bank for channel coding
- Filter bank in AD and DA conversions

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Manuscript Due	June 1, 2009
First Round of Reviews	September 1, 2009
Publication Date	December 1, 2009

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Special Issue on

Advances in Multidimensional Synthetic Aperture Radar Signal Processing

CALL FOR PAPERS

Synthetic Aperture Radar (SAR) represents an established and mature high-resolution remote-sensing technology that has been successfully employed to observe, study, and characterize the Earth's surface. Over the last two decades, one-dimensional SAR systems have evolved more and more toward multidimensional configurations, enabling quantitative remote sensing with SAR sensors. The relevance of the multidimensional SAR technology is primarily supported by the recent launch of the Japanese ALOS, the German TERRASAR-X, and the Canadian RADARSAT-2 orbital systems, and also by the development of airborne and ground-based SAR systems.

Multidimensional SAR data can be generated by different sources of diversity: baseline, polarization, frequency, time, as well as their different combinations. This degree of freedom makes multidimensional SAR data sensitive to a wide range of geophysical and biophysical features of the Earth's surface. Consequently, the definition of novel multidimensional signal processing techniques is essential to benefit from this information richness, especially when the objective is the quantitative retrieval of new parameters, and also necessary to extend already existing one-dimensional SAR data tools.

This special issue is seeking for original contributions in the definition of novel signal processing techniques, and also for works on the assessment of new physical or statistical models to improve the understanding of multidimensional SAR data and the extraction of information, considering the important challenges and limitations imposed by the physics governing the imaging process. This issue is also open to contributions oriented toward the exploitation of multidimensional SAR data for novel and exciting applications.

Topics of interest include, but are not limited to:

- Multibaseline interferometry and differential interferometry
- 3D reconstruction and multidimensional SAR focusing techniques
- Polarimetry and polarimetric interferometry
- Multitemporal and multifrequency SAR
- Novel multidimensional SAR system configurations
- Multidimensional SAR data classification and change detection
- Information extraction from multidimensional SAR data

- Multidimensional SAR data statistical modeling, filtering, and estimation
- Definition and assessment of electromagnetic models
- Extraction and estimation of geophysical and biophysical parameters
- Space-time adaptive processing (STAP)

Before submission authors should carefully read over the journal's Author Guidelines, which are located at http://www.hindawi.com/journals/asp/guidelines.html. Prospective authors should submit an electronic copy of their complete manuscript through the journal Manuscript Tracking System at http://mts.hindawi.com/ according to the following timetable:

Manuscript Due	June 1, 2009
First Round of Reviews	September 1, 2009
Publication Date	December 1, 2009

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Special Issue on

Advances in Signal Processing for Maritime Applications

CALL FOR PAPERS

The maritime domain continues to be important for our society. Significant investments continue to be made to increase our knowledge about what "happens" underwater, whether at or near the sea surface, within the water column, or at the seabed. The latest geophysical, archaeological, and oceanographical surveys deliver more accurate global knowledge at increased resolutions. Surveillance applications allow dynamic systems, such as marine mammal populations, or underwater intruder scenarios, to be accurately characterized. Underwater exploration is fundamentally reliant on the effective processing of sensor signal data. The miniaturization and power efficiency of modern microprocessor technology have facilitated applications using sophisticated and complex algorithms, for example, synthetic aperture sonar, with some algorithms utilizing underwater and satellite communications. The distributed sensing and fusion of data have become technically feasible, and the teaming of multiple autonomous sensor platforms will, in the future, provide enhanced capabilities, for example, multipass classification techniques for objects on the sea bottom. For such multiplatform applications, signal processing will also be required to provide intelligent control procedures.

All maritime applications face the same difficult operating environment: fading channels, rapidly changing environmental conditions, high noise levels at sensors, sparse coverage of the measurement area, limited reliability of communication channels, and the need for robustness and low energy consumption, just to name a few. There are obvious technical similarities in the signal processing that have been applied to different measurement equipment, and this Special Issue aims to help foster cross-fertilization between these different application areas.

This Special Issue solicits submissions from researchers and engineers working on maritime applications and developing or applying advanced signal processing techniques. Topics of interest include, but are not limited to:

- Sonar applications for surveillance and reconnaissance
- Radar applications for measuring physical parameters of the sea surface and surface objects
- Nonacoustic data processing and sensor fusion for improved target tracking and situational awareness

- Underwater imaging for automatic classification
- Signal processing for distributed sensing and networking including underwater communication
- Signal processing to enable autonomy and intelligent control

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Manuscript Due	July 1, 2009
First Round of Reviews	October 1, 2009
Publication Date	January 1, 2010

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Special Issue on Dependable Semantic Inference

CALL FOR PAPERS

After many years of exciting research, the field of multimedia information retrieval (MIR) has become mature enough to enter a new development phase—the phase in which MIR technology is made ready to get adopted in practical solutions and realistic application scenarios. High users' expectations in such scenarios require high dependability of MIR systems. For example, in view of the paradigm "getting the content I like, anytime and anyplace" the service of consumer-oriented MIR solutions (e.g., a PVR, mobile video, music retrieval, web search) will need to be at least as dependable as turning a TV set on and off. Dependability plays even a more critical role in automated surveillance solutions relying on MIR technology to analyze recorded scenes and events and alert the authorities when necessary.

This special issue addresses the dependability of those critical parts of MIR systems dealing with semantic inference. Semantic inference stands for the theories and algorithms designed to relate multimedia data to semantic-level descriptors to allow content-based search, retrieval, and management of data. An increase in semantic inference dependability could be achieved in several ways. For instance, better understanding of the processes underlying semantic concept detection could help forecast, prevent, or correct possible semantic inference errors. Furthermore, the theory of using redundancy for building reliable structures from less reliable components could be applied to integrate "isolated" semantic inference algorithms into a network characterized by distributed and collaborative intelligence (e.g., a social/P2P network) and let them benefit from the processes taking place in such a network (e.g., tagging, collaborative filtering).

The goal of this special issue is to gather high-quality and original contributions that reach beyond conventional ideas and approaches and make substantial steps towards dependable, practically deployable semantic inference theories and algorithms.

Topics of interest include (but are not limited to):

- Theory and algorithms of robust, generic, and scalable semantic inference
- Self-learning and interactive learning for online adaptable semantic inference
- Exploration of applicability scope and theoretical performance limits of semantic inference algorithms
- Modeling of system confidence in its semantic inference performance

- Evaluation of semantic inference dependability using standard dependability criteria
- Matching user/context requirements to dependability criteria (e.g., mobile user, user at home, etc.)
- Modeling synergies between different semantic inference mechanisms (e.g., content analysis, indexing through user interaction, collaborative filtering)
- Synergetic integration of content analysis, user actions (e.g., tagging, interaction with content) and user/device collaboration (e.g., in social/P2P networks)

Authors should follow the EURASIP Journal on Image and Video Processing manuscript format described at http://www.hindawi.com/journals/ivp/. Prospective authors should submit an electronic copy of their complete manuscripts through the journal Manuscript Tracking System at http://mts.hindawi.com/, according to the following timetable:

Manuscript Due	December 1, 2009
First Round of Reviews	March 1, 2010
Publication Date	June 1, 2010

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Special Issue on

Simulators and Experimental Testbeds Design and Development for Wireless Networks

CALL FOR PAPERS

In the context of wireless networking, performance evaluation of protocols and distributed applications is generally conducted through simulation or experimentation campaigns. An efficient and accurate simulation of wireless networks raises various issues which generally need to be addressed from several research domains simultaneously. As examples, we can consider the wireless physical layer modeling and simulation, the support of large-scale networks, the simulation of complex RF systems such as MIMO ones, the emulation of wireless nodes or the interconnection of simulators, experimental testbeds, and so forth.

The aim of this Special Issue is to bring together academic and industry researchers and practitioners from both the wireless networking and the simulation communities to discuss current and future trends in simulation or experimentation techniques, models, and practices for the future communication system and to foster interdisciplinary collaborative research in this area. The guest editors seek high-quality papers on aspects of wireless network simulation, and value both theoretical and practical research contributions. Topics of interest include, but are not limited to:

- Radio medium modeling and cross-layer simulation
- Scalability, large-scale networks support
- Validation of simulators and simulation results
- Simulators benchmarking and comparisons
- Fluid-flow simulation for assessing QoS in large-scale networks
- Support of new emerging technologies (WiMax, 3.5G, Wireless Mesh Networks, 802.11x, etc.) in simulators
- Support of advanced RF systems (Multi-carrier schemes, MIMO, smart-antenna) in simulators
- Wireless node simulation or emulation
- Interoperability of simulators, emulators, and experiments
- Support of distributed physical layer schemes (distributed signal processing; cooperative schemes)

- Distributed simulation, and scalability of simulators
- Implementation of simulators
- Experimental testbeds for wireless networks
- Methodology for protocol and distributed application performance evaluation
- SDR techniques, cognitive radio approaches, dynamic spectrum access testbeds, and simulators as well as modeling
- Simulation and testbeds for cooperative communication protocols

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Manuscript Due	June 1, 2009
First Round of Reviews	September 1, 2009
Publication Date	December 1, 2009

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Simulators and Experimental Testbeds Design and Development for Wireless Networks

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Manuscript Due	June 1, 2009
First Round of Reviews	September 1, 2009
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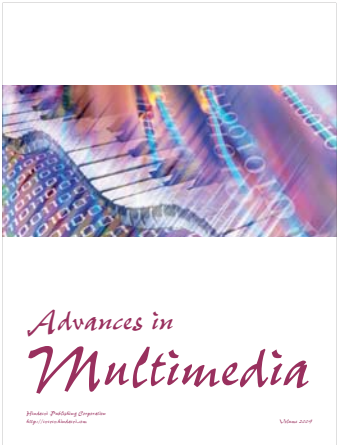
Advances in Multimedia

http://www.hindawi.com/journals/am/

Aims and Scope

Advances in Multimedia is aimed at presenting comprehensive coverage of the field of multimedia. The journal covers research and developments in multimedia technology and applications, including compression, storage, networking, communication, retrieval, algorithms, architectures, software design, circuits, multimedia signal processing, and multimodality devices and systems. Types of multimedia signals involved include audio, speech, video, image, graphics, geophysical, musical, sonar, radar, and medical signals.



Manuscript Submission

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We are committed to keeping the publication speed of the journal as fast as possible, while at the same time ensuring a thorough peer-review process. In order to ensure the fastest possible publication speed following the acceptance of a manuscript, once an article is accepted we make the author's version immediately available online. Then, in an average of 60 days, we publish the final edited version of the paper. Following an article-by-article schedule rather than an issue-by-issue schedule allows for a much faster publication speed, since articles are not delayed until an entire issue has been completed.

International Editorial Board

The journal has a distinguished Editorial Board with extensive academic qualifications, which ensures that the journal maintains high scientific standards and has broad international coverage.

Hindawi

Computational Intelligence & Neuroscience

http://www.hindawi.com/journals/cin/

Aims and Scope

Computational Intelligence and Neuroscience is a forum for the publication of research in the interdisciplinary field of neural computing, neural engineering, and artificial intelligence, where neuroscientists, cognitive scientists, engineers, psychologists, physicists, computer scientists, and artificial intelligence investigators among others can publish their work in one periodical that bridges the gap between neuroscience, artificial intelligence, and engineering. The journal provides research and review papers at an interdisciplinary level, with the field of intelligent systems for computational neuroscience as its focus.

Manuscript Submission

Manuscripts are invited and should be submitted by one of the authors of the manuscript through the online Manuscript Tracking System located at http://mts.hindawi.com.

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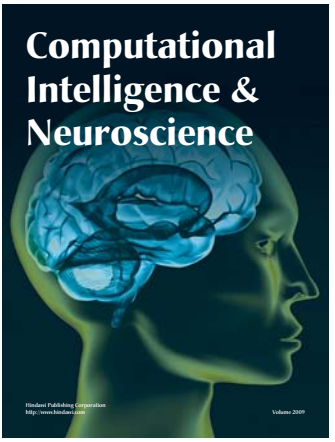

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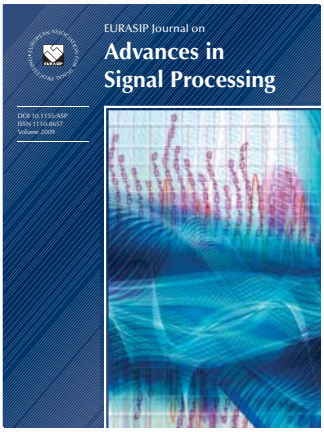
Hindawi

EURASIP Journal on
Advances in Signal Processing
http://www.hindawi.com/journals/asp/

Aims and Scope

The aim of the EURASIP Journal on Advances in Signal Processing is to highlight the theoretical and practical aspects of signal processing in new and emerging technologies. Application areas include (but are not limited to) communications, networking, sensors and actuators, radar and sonar, medical imaging, biomedical applications, remote sensing, consumer electronics, computer vision, pattern recognition, robotics, fiber optic sensing/transducers, industrial automation, transportation, stock market and financial analysis, seismography, and avionics.



Indexing/Abstracting

In order to provide the maximum exposure for all published articles, the EURASIP Journal on Advances in Signal Processing is covered by many leading abstracting and indexing databases.

Manuscript Submission

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Open Access

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Hindawi

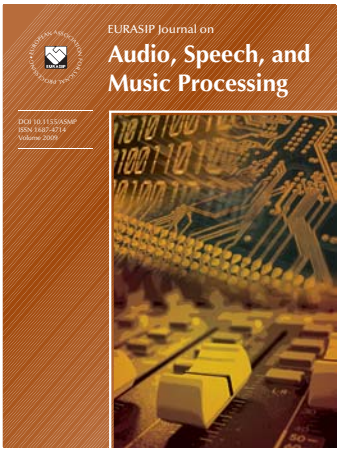
EURASIP Journal on

Audio, Speech, and Music Processing

http://www.hindawi.com/journals/asmp/

Aims and Scope

EURASIP Journal on Audio, Speech, and Music Processing is a peer-reviewed, open access journal, which aims at bringing together researchers, scientists, and engineers working on the theory and applications of the processing of various audio signals, with a specific focus on speech and music.



The journal is dedicated to original research work, but also allows tutorial and review articles. Articles deal with both theoretical and practical aspects of audio, speech, and music processing.

Manuscript Submission

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Hindawi

EURASIP Journal on

Bioinformatics and Systems Biology

http://www.hindawi.com/journals/bsb/

Aims and Scope

The overall aim of EURASIP Journal on Bioinformatics and Systems Biology is to publish research results related to signal processing and bioinformatics theories and techniques relevant to a wide area of applications into the core new disciplines of genomics, proteomics, and systems biology.

The journal is intended to offer a common platform for scientists from several areas including signal processing, bioinformatics, statistics, biology, and medicine, who are interested in the development of algorithmic, mathematical, statistical, modeling, simulation, data mining, and computational techniques, as demanded by various applications in genomics, proteomics, system biology, and more general in health and medicine.

Manuscript Submission

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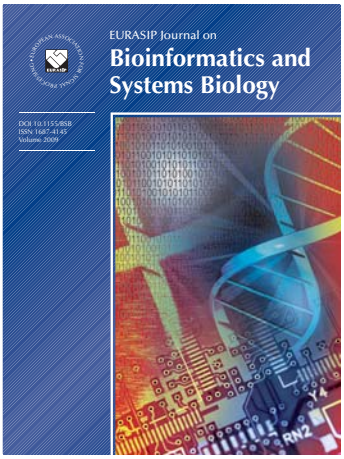

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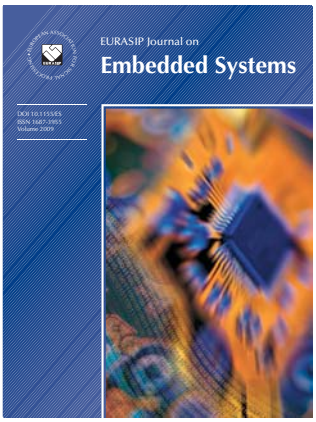
Hindawi

EURASIP Journal on Embedded Systems

http://www.hindawi.com/journals/es/

Aims and Scope

EURASIP Journal on Embedded Systems is a peer-reviewed open access journal that serves the large community of researchers and professional engineers who deal with the theory and practice of embedded systems, including complex homogeneous and heterogeneous embedded systems, specification languages and tools for embedded systems, modeling and verification techniques, hardware/software tradeoffs and codesign, new design flows, design methodologies and synthesis methods, platform-based design, component-based design, adaptation of signal processing algorithms to limited implementation resources, rapid prototyping, computing structures and architectures for complex embedded systems, real-time operating systems, methods and techniques for the design of low-power systems, interfacing with the real world, and novel application case studies and experiences.



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Hindawi

EURASIP Journal on Image and Video Processing

http://www.hindawi.com/journals/ivp/

Aims and Scope

EURASIP Journal on Image and Video Processing is a peer-reviewed, open access journal, intended for researchers from both academia and industry, who are active in the multidisciplinary field of image and video processing. The scope of the journal covers all theoretical and practical aspects of the domain, from basic research to the development of applications.

Contributed articles on image and video processing may be focused on specific techniques, on diverse functionalities and services, within the context of various activity sectors (e.g., multimedia, medical, aerial, robotics, security, communications, arts), or on employing diverse data formats.

Manuscript Submission

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Open Access

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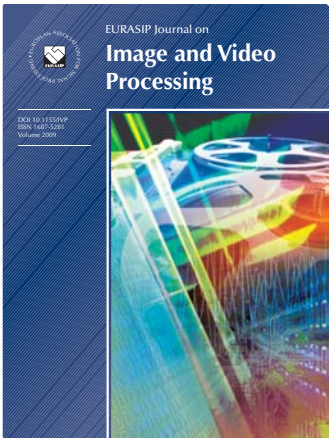

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Hindawi

EURASIP Journal on Information Security

http://www.hindawi.com/journals/is/

Aims and Scope

The overall goal of the EURASIP Journal on Information Security is to bring together researchers and practitioners dealing with the general field of information security with a particular emphasis on the use of signal processing tools to enable the security of digital contents. As such, it addresses any work whereby security primitives and multimedia signal processing are used together to ensure the secure access to the data. Enabling technologies include watermarking, data hiding, steganography and steganalysis, joint signal processing and encryption, perceptual hashing, identification, biometrics, fingerprinting, and digital forensics.

Manuscript Submission

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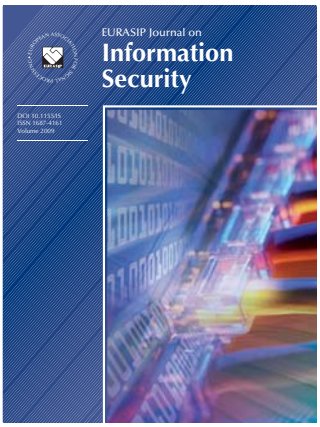

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Hindawi

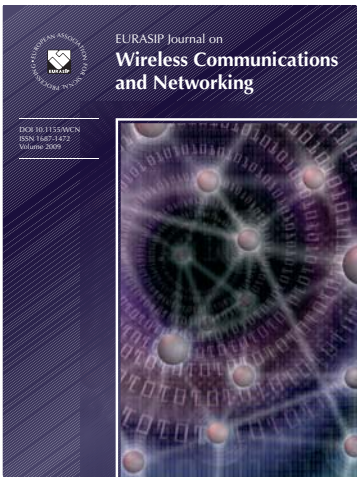
EURASIP Journal on

Wireless Communications and Networking

http://www.hindawi.com/journals/wcn/

Aims and Scope

The overall aim of the EURASIP Journal on Wireless Communications and Networking is to bring together science and applications of wireless communications and networking technologies, with emphasis on signal processing techniques and tools. Subject areas include antenna systems and design, channel modeling and propagation, coding for wireless systems, multiuser and multiple access schemes, optical wireless communications, resource allocation over wireless networks, security, authentication, and cryptography for wireless networks, signal processing techniques and tools, software and cognitive radio, wireless traffic and routing, ultra-wideband systems, vehicular networks, wireless multimedia communication, wireless sensor networks, and wireless system architectures and applications.



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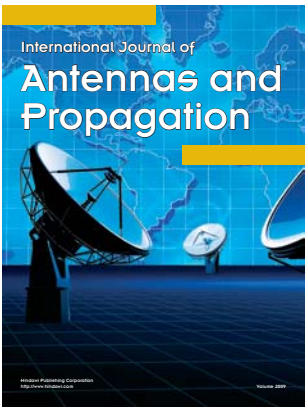
Hindawi

International Journal of Antennas and Propagation

http://www.hindawi.com/journals/ijap/



Aims and Scope

The overall aim of the International Journal of Antennas and Propagation is to explore emerging concepts and applications in antennas and propagation. The journal focuses on the physical link from antenna to antenna including antenna hardware and associated electronics, the nature and impact of propagation channels and measurement, prediction, and simulation methods for evaluating or designing antennas or the channel. The journal is directed at both practicing engineers and academic researchers and will highlight new ideas and challenges in antennas and propagation for both application development and basic research.

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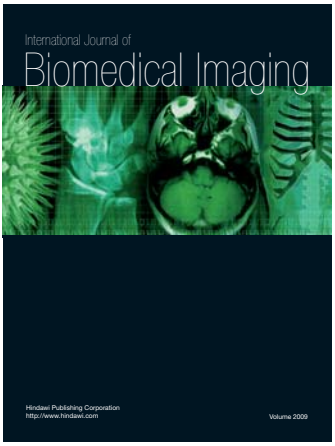
Hindawi

International Journal of
Biomedical Imaging
http://www.hindawi.com/journals/ijbi/

Aims and Scope

The overall goal of the International Journal of Biomedical Imaging is to promote the research and development of biomedical imaging by publishing high-quality research articles and reviews in this rapidly growing, interdisciplinary field. Generally speaking, the scope of the journal covers data acquisition, image reconstruction, and image analysis, involving theories, methods, systems, and applications.



Indexing/Abstracting

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Manuscript Submission

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Hindawi

International Journal of

Digital Multimedia Broadcasting

http://www.hindawi.com/journals/ijdmb/

Aims and Scope

International Journal of Digital Multimedia Broadcasting aims to provide a high-quality and timely forum for engineers, researchers, and educators whose interests are in digital multimedia broadcasting to learn recent developments, to share related challenges, to compare multistandards, and further to design new and improved systems.

Subject areas include (but are not limited to):

- Multimedia broadcasting overall system and standardization, multimedia signal compression, and coding for broadcasting
- Multimedia streaming and control, IPTV with broadcasting, multimedia content services, and digital rights management over broadcasting
- Modulation and demodulation
- Channel estimation and equalization
- VLSI design and system-on-chip implementation for multimedia broadcasting reception
- Cross-layer analysis and integration, single-chip solution, and power and spectral efficiency
- Antenna and propagation for multimedia transmission and reception
- Multistandards compatibility and multisystems interoperability
- Multibands frequency interface issues, spectrum management, and usage
- Filed-trials and testing analyses
- Quality of service and quality of experience in multimedia broadcasting

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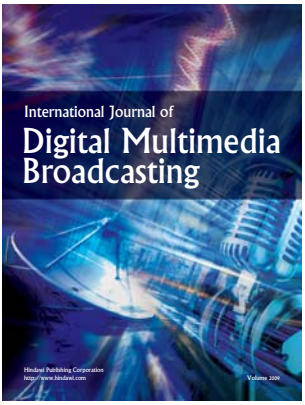

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Hindawi

International Journal of
Navigation and Observation

http://www.hindawi.com/journals/ijno/

Aims and Scope

The overall aim of the International Journal of Navigation and Observation is to explore emerging concepts and applications in navigation, positioning, Earth observation, and related fields. The journal is directed at both practicing engineers as well as academic researchers. It will highlight new ideas and challenges in both application development and basic research, thus seeking to bridge the gap between innovation and practical implementation. Authors of manuscripts with novel contributions to the theory and/or the practice of navigation, positioning, and Earth observation are encouraged to submit their contributions for consideration.

Manuscript Submission

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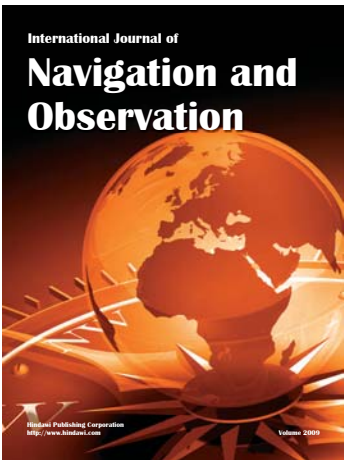

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Hindawi

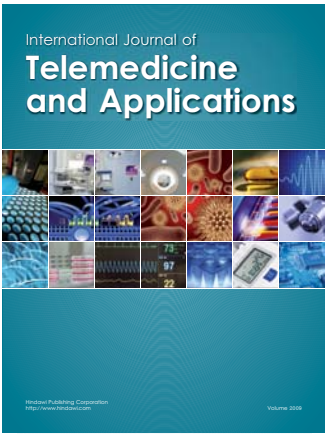
International Journal of

Telemedicine and Applications

http://www.hindawi.com/journals/ijta/

Aims and Scope

The overall aim of the International Journal of Telemedicine and Applications is to bring together science and applications of medical practice and medical care at a distance as well as their supporting technologies such as computing, communications, and networking technologies with emphasis on telemedicine techniques and telemedicine applications. Telemedicine is an information technology that enables doctors to perform medical consultations, diagnoses, and treatments, as well as medical education, away from patients. International Journal of Telemedicine and Applications will highlight the continued growth and new challenges in telemedicine, applications, and their supporting technologies, for both application development and basic research.



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Hindawi

Research Letters in Communications

http://www.hindawi.com/journals/rlc/

Aims and Scope

Research Letters in Communications is devoted to very fast publication of short, high-quality manuscripts in the broad field of communications. Average time from submission to publication will be around 60 days.

Manuscript Submission

Submitted manuscripts should not exceed 5 pages in their final published form. Manuscripts are invited and should be submitted by one of the authors of the manuscript through the online Manuscript Tracking System which is located at http://mts.hindawi.com.

Peer Review

All manuscripts are subject to peer review and are expected to meet standards of academic excellence. Submissions will be considered by an associate editor and—if not rejected right away—by peer reviewers, whose identities will remain anonymous to the authors.

Open Access

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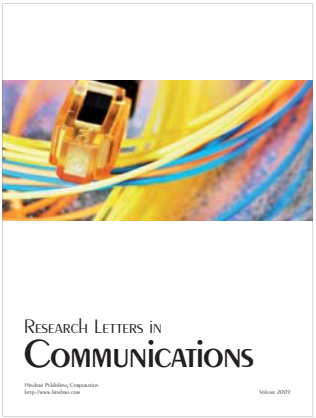

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Hindawi

VLSI Design

http://www.hindawi.com/journals/vlsi/

Aims and Scope

VLSI Design is a peer-reviewed, open access journal, which presents state-of-the-art papers in VLSI design, computer-aided design, design analysis, design implementation, simulation, and testing. Topics relating to both theory and applications are discussed. The journal's scope also includes papers that address technical trends, pressing issues, and educational aspects in VLSI Design.

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