Upper Bounds on the BER Performance of MTCM-STBC Schemes over Shadowed Rician Fading Channels

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Space-time block coding (STBC) provides substantial diversity advantages with a low decoding complexity. However, these codes are not designed to achieve coding gains. Outer codes should be concatenated with STBC to provide additional coding gain. In this paper, we analyze the performance of concatenated trellis-coded STBC schemes over shadowed Rician frequency-flat fading channels. We derive an exact pairwise error probability (PEP) expression that reveals the dominant factors affecting performance. Based on the derived PEP, in conjunction with the transfer function technique, we also present upper bounds on the bit error rate (BER), which are further shown to be tight through a Monte-Carlo simulation study.

Keywords and phrases: space-time block coding, trellis-coded modulation, Rician fading channels, shadowing, pairwise error probability.

1. INTRODUCTION

Space-time trellis coding was introduced in [1] as an effective transmit diversity technique to combat fading. These codes were designed to achieve maximum diversity gain. However, for a fixed number of transmit antennas, their decoding complexity increases exponentially with the transmission rate. Space-time block coding (STBC) [2] was proposed as an attractive alternative to its trellis counterpart with a much lower decoding complexity. The work in [2] was inspired by Alamouti’s early work [3], where a simple two-branch transmit diversity scheme was presented and shown to provide the same diversity order as maximal-ratio receiver combining with two receive antennas. Alamouti’s scheme is appealing in terms of its performance and simplicity. Assuming the channel is known at the receiver, it requires a simple maximum-likelihood decoding algorithm based only on linear processing at the receiver. STBC generalizes Alamouti’s scheme to an arbitrary number of transmit antennas and is able to provide the full diversity promised by the transmit and receive antennas. However, these codes are not designed to achieve a coding gain. Therefore, outer codes should be concatenated with STBC to achieve additional coding gains. A pioneering work towards this end is presented in [4] where concatenation of trellis-coded modulation (TCM) with STBC is considered. In [4], it is shown that the free distance of the trellis code dominates performance; therefore, the optimal trellis codes designed for additive white Gaussian noise (AWGN) are also optimum for concatenated TCM-STBC over quasistatic Rayleigh fading channels. We studied the same concatenated scheme combined with an interleaver in [5] over Rician fading channels. In this paper, we generalize our work to shadowed Rician channels. The shadowed Rician channel [6] is a generalization of the Rician model, where the line-of-sight (LOS) path is subjected to a lognormal transformation due to foliage attenuation or blockage, also referred to as shadowing. Specifically, we derive an exact pairwise error probability (PEP) for concatenated TCM-STBC schemes. Our exact evaluation of PEP is based on the moment-generating function technique [7, 8], which has been successfully applied to the analysis of digital communication systems over fading channels. Using the classical transfer function technique based on the exact PEP, we obtain upper bounds on bit error rate (BER) performance, which are further verified through simulation. Our analysis also reveals the selection criteria for trellis codes which should be used in conjunction with STBC.

The organization of the paper is as follows. In Section 2 we explain our system model, where the concatenated TCM-STBC is described and the channel model under consideration is introduced. In Section 3 an exact expression for PEP is derived for the TCM-STBC scheme using the MGF approach. Based on the derived PEP, we discuss the selection criteria
for trellis codes which should be used with space-time codes for optimal performance and compare them with the classical selection criteria for trellis codes over fading channels without transmitter diversity. In Section 4, using the transfer function technique in conjunction with the derived PEP expressions, we obtain upper bounds on the BER performance. Analytical performance results are presented for two example trellis codes, which are further confirmed through Monte-Carlo simulation.

2. SYSTEM MODEL

We consider a wireless communication scenario where the transmitter is equipped with $M$ antennas and the receiver is equipped with $N$ antennas. The binary data is first encoded by a trellis encoder. After trellis coded symbols are interleaved and mapped to constellation symbols, they are fed to the STBC encoder. After trellis coded symbols are interleaved and mapped to constellation symbols, they are fed to the STBC encoder. An STBC is defined [2] by an $L \times M$ code matrix, where $L$ represents the number of time intervals for transmitting $P$ symbols, resulting in a code rate of $P/L$. For Tarokh et al.’s orthogonal space-time block codes [2], the entries of the code matrix are chosen as linear combinations of the transmission symbols and their conjugates. For example, the code matrix for the well-known Alamouti’s scheme (i.e., STBC for 2 transmit antennas) is given by

$$
\begin{bmatrix}
    x_1 & x_2 \\
    -x_2^* & x_1^*
\end{bmatrix}
$$

with $M = P = L = 2$.

We assume that the transmission frame from each antenna consists of a total of $FL$ symbols (i.e., consecutive $F$ smaller inner-frames, each of them having duration $L$ symbols corresponding to the STBC length). The received signal at receiver antenna $n$ ($n = 1, 2, \ldots, N$) at time interval $f$ of the $f$th frame ($f = 1, 2, \ldots, F$) inner-frame is a superposition of $M$ transmitted signals:

$$
r_n^f(l) = \sum_{m=1}^{M} a_{m,n}^f x_m^f(l) + \eta_n^f(l),
$$

where $x_m^f(l)$ is the modulation symbol transmitted from the $m$th transmit antenna at time interval $l$ of the $f$th frame and $\eta_n^f(l)$ is additive noise, modeled as a complex Gaussian random variable with zero mean and variance $\sigma^2$ per dimension. $a_{m,n}^f$ represents the fading coefficient modeling the channel from the $m$th transmit to the $n$th receive antenna during the $f$th inner frame and are assumed to be independent and identically distributed (i.i.d.). The fading coefficient is assumed to remain constant over an inner-frame period (i.e., $L$ symbol intervals). This assumption is necessary to make use of the orthogonal structure of STBC to guarantee full spatial diversity. The assumption of quasistatic behavior of the channel over an inner-frame period can be justified using an $L$-symbol interleaver over a moderately slow varying channel. In our case, the fading amplitude is described by the shadowed Rician fading model. In this model, the LOS component is not constant but rather a lognormally distributed random variable. The fading coefficient can be expressed (dropping the subscripts and superscripts for notational convenience) as $\alpha = \mu + \xi + j\xi_1$, where $\xi_0$ and $\xi_1$ are independent Gaussian random variables with zero mean and variance $\sigma^2$. Here, the LOS component is given as $\mu = \exp(\xi_0)$ where $\xi_0$ is a Gaussian random variable with mean $m\mu$ and variance $\sigma^2_0$, and independent of $\xi_0$ and $\xi_1$. The conditional probability density function of the fading amplitude $|\alpha|$ is

$$
p_{|\alpha|/\mu}(|\alpha|/\mu) = \frac{1}{\sqrt{2\pi\sigma^2}} \exp\left(-\frac{|\alpha|^2}{2\sigma^2}\right),
$$

where $I_0(\cdot)$ is the zero-order modified Bessel function of the first kind, and the probability density function of the LOS component is given by

$$
p_\mu(\mu) = \frac{1}{\sqrt{2\pi m\mu}} \exp\left(-\frac{(\ln m - m\mu)^2}{2\sigma^2}\right).
$$

The parameters $\sigma$, $\sigma_0$, and $m\mu$ in (3) and (4) specify the degree of shadowing. Denoting by $\mathbb{C}^{m \times n}$ the vector space of $m$-by-$n$ complex matrices, and defining

$$
\begin{align*}
    r_n^f &= (r_n^f(1), r_n^f(2), \ldots, r_n^f(L))^T \in \mathbb{C}^{L \times 1}, \\
    a_n^f &= (a_{1,n}^f, a_{2,n}^f, \ldots, a_{M,n}^f)^T \in \mathbb{C}^{M \times 1}, \\
    \eta_n^f &= (\eta_n^f(1), \eta_n^f(2), \ldots, \eta_n^f(L))^T \in \mathbb{C}^{L \times 1},
\end{align*}
$$

the received signal can be written in matrix notation as

$$
r_n^f = X^f a_n^f + \eta_n^f, \quad n = 1, 2, \ldots, N, \quad f = 1, 2, \ldots, F,
$$

where $X^f \in \mathbb{C}^{L \times M}$ consists of space-time encoded symbols (which have been already trellis encoded) for the $f$th inner frame. At the receiver, first the received signal is passed through the space-time decoder, which is essentially based on linear processing for STBC from orthogonal designs [2]. After deinterleaving, the processed sequence is fed to the trellis decoder implemented by a Viterbi algorithm. If a multiple TCM (MTCM) scheme with $M$ symbols per branch is used (note that the number of transmit antennas is also given as $M$), the decoding steps can be combined in one step with a proper modification of the metric employed in the Viterbi algorithm. In this case, the received signal is just deinterleaved and fed directly to the Viterbi decoder without any further processing.

3. DERIVATION OF EXACT PEP

In this section, we analyze the PEP of the concatenated scheme over shadowed Rician fading channels assuming

1Throughout this paper, we use $(\cdot)^T$ and $(\cdot)^H$ for the transpose and transpose conjugate operations, respectively. Upper case bold face letters represent matrices and lower case bold face letters represent vectors.
perfect channel state information is available at the receiver. Assuming equal transmitted power at all transmit antennas, the conditional PEP of transmitting code matrix $X$ (which consists of $X^f, f = 1, 2, \ldots, F$) and erroneously deciding in favor of another code matrix $\hat{X}$ at the decoder is given by

$$P(X, \hat{X} | \alpha_{m,n}, \mu_{m,n}, m = 1, \ldots, M, n = 1, \ldots, N, f = 1, \ldots, F) = Q\left(\frac{1}{\sqrt{2\pi\sigma^2}} \int_0^{\pi/2} \Phi_1\left(-\frac{\mu}{\sin^2 \theta}\right) d\theta\right),$$

where $Q(\cdot)$ is the Gaussian Q-function.

To find the unconditional PEP, we still need to take an expectation of (13) with respect to $\mu$, whose distribution is given by (4). This expectation yields

$$P(X, \hat{X} | \mu) = \frac{1}{\pi} \int_0^{\pi/2} \prod_{f=1}^{F} \prod_{n=1}^{N} \left(1 + \Omega_f / \sin^2 \theta \right)^{2\sigma^2 \Omega_f / \sin^2 \theta} \left(\frac{1}{2\sigma^2 \Omega_f / \sin^2 \theta}ight)^{2\sigma^2 \Omega_f / \sin^2 \theta} d\theta.$$

Introducing the variable change $u = (\ln \mu - m_f) / \sqrt{2\sigma^2}$, (15) can be rewritten as

$$P(X, \hat{X}) = \frac{1}{\pi} \int_0^{\pi/2} \prod_{f=1}^{F} \prod_{n=1}^{N} \left(1 + \Omega_f / \sin^2 \theta \right)^{2\sigma^2 \Omega_f / \sin^2 \theta} \left(\frac{1}{2\sigma^2 \Omega_f / \sin^2 \theta}ight)^{2\sigma^2 \Omega_f / \sin^2 \theta} \exp\left(-\frac{u^2}{2\sigma^2 \Omega_f / \sin^2 \theta}ight) d\theta.$$

The inner integral has the form of $\int_0^{\infty} \exp(-u^2)f(u)du$, which can be expressed in terms of an infinite sum (see the appendix). This yields the final form of the exact PEP as

$$P(X, \hat{X}) = \frac{1}{\pi} \int_0^{\pi/2} \prod_{f=1}^{F} \left(1 + \Omega_f / \sin^2 \theta \right)^{2\sigma^2 \Omega_f / \sin^2 \theta} \left(\frac{1}{2\sigma^2 \Omega_f / \sin^2 \theta}ight)^{2\sigma^2 \Omega_f / \sin^2 \theta} \exp\left(-\Delta_f(\theta)\right) d\theta,$$

where

$$\Delta_f(\theta) = \sum_{k=2}^{\infty} \frac{(k-1)!!}{k!} (2\sigma^2)_{\text{even}}^{k}.$$
where
\[ \Delta_f(\theta) = \frac{1}{2\sigma^2} \frac{\Omega_f}{\sin^2 \theta} \exp \left( 2m_\mu \right) \] (18)
and \((k-1)!! = 1.3 \cdot \cdots \cdot k \) [11, page xlv]. The coefficients \( g_{k,d} \) in (17) can be computed by the recursive equation given in the appendix. It is worth noting that even considering only the first term in the infinite summation in (17) gives a very good approximation for practical values of shadowing. Setting \( k = 2 \) and noting that \( g_{2,1} = -1 \) and \( g_{2,2} = 1 \), we have

\[ P(X,X) \geq \frac{1}{\pi} \int_{\theta=0}^{\pi/2} \prod_{f=1}^{E_f/4N_0} \left( \frac{1 + K + (E_f/4N_0)(\beta/\sin^2 \theta) \sum_{p=1}^{P} x_{p} - \hat{x}_{p} \right)^2 \times \left[ 1 - 2\sigma^2 \Delta_f(\theta) + 2\sigma^2 (\Delta_f(\theta))^2 \right]^{MN} \] .

(19)

In our numerical results, taking more terms (i.e., \( k > 2 \)) did not result in a visible change in the plots.

It is also interesting to point out how (17) relates to the unshadowed case. Assuming there is no shadowing, \( \mu \) is no longer a log-normal random variable, but just given as a constant equal to its mean \( \mu = \exp(2m_\mu) \). Furthermore, inserting \( \sigma^2 = 0 \) in (17) and using the relationships \( \sigma^2 = 0.5/(1+K) \) and \( \mu = \sqrt{K/(1+K)} \) in terms of the well-known Rician parameter \( K \), we obtain

\[ P(X,X) \]

\[ \leq \left( \frac{E_f}{4N_0} \right)^{\frac{1}{2}} \prod_{f=1}^{P} \left( \frac{\beta}{M} \sum_{p=1}^{P} x_{p} - \hat{x}_{p} \right)^2 \times \left[ q(\sigma, \sigma_\mu, m_\mu) \right]^{\frac{1}{2}} \times \left( \frac{1 + 2\sigma^2 \Delta_f(\theta) + 2\sigma^2 (\Delta_f(\theta))^2}{1 - 2\sigma^2 \Delta_f(\theta) + 2\sigma^2 (\Delta_f(\theta))^2} \right)^{MN} \] .

(21)

where

\[ q(\sigma, \sigma_\mu, m_\mu) = \frac{1}{2\sigma^2} \exp \left( - \frac{1}{2\sigma^2} \exp (2m_\mu) \right) \] \times \left[ 1 + \sum_{k=2}^{K} \frac{(k-1)!!}{k!} (2\sigma_\mu)^k \sum_{d=1}^{D} g_{k,d} \left( \frac{1}{2\sigma^2} \exp (2m_\mu) \right)^d \right].

(22)

Here, \( \Psi \) is the set of inner frames (with a length of \( L \) symbols) at nonzero Euclidean distance summations and \(|\Psi|\) is the number of elements in this set. This can be compared to effective length (EL) in TCM schemes [12], which is defined as the smallest number of symbols at nonzero Euclidean distances. Contrary to the symbol-by-symbol count in the definition of EL, frame-by-frame count is considered here as a result of the multidimensional structure of STBC spanning an interval of \( L \) symbols. It should also be noted that symbol-by-symbol interleaving is considered for the single antenna case while an \( L \)-symbol interleaver is employed in our case. In (21), the slope of the performance curve, which yields the diversity order, is determined by \(|\Psi|NM\) and it can be defined as generalized effective length (GEL) for multiple antenna systems in analogy to the effective length for single antenna case.

The second term in (21) contributes to the coding gain, which corresponds to the horizontal shift in the performance curve. Recalling the definition of product distance (PD) for the single antenna case (which is given as the product of nonzero branch distances along the error event), we now define the generalized product distance (GPD)

\[ \prod_{f=1}^{P} \left( \frac{\beta}{M} \sum_{p=1}^{P} x_{p} - \hat{x}_{p} \right)^2 \times \left[ q(\sigma, \sigma_\mu, m_\mu) \right]^{\frac{1}{2}} \times \left[ \frac{1 + 2\sigma^2 \Delta_f(\theta) + 2\sigma^2 (\Delta_f(\theta))^2}{1 - 2\sigma^2 \Delta_f(\theta) + 2\sigma^2 (\Delta_f(\theta))^2} \right]^{MN} \] .

(23)

which involves the product of nonzero branch distance summations, where the summation is over \( P \) terms based on the STBC used.

The third term in (21) is completely characterized by channel parameters. Since maximization of diversity order is the primary design criterion, the first step in “good” code design is the maximization of \(|\Psi|\), since \( M \) and \( N \) are already fixed. Once diversity order is optimized, the third term becomes just a constant. This makes us conclude that the GEL and GPD are the appropriate performance criteria in the selection of trellis codes over shadowed Rician channels. This also shows that the trellis codes designed for optimum performance (based on classical effective code length and minimum product distance) over fading channels for the single transmit antenna case are not necessarily optimum for the multiple antenna case.

To derive the upper bound on bit error probability from the exact PEP, we follow the classical transfer function approach. The upper bound is given in terms of the transfer
function of the code \( T(D,I) \) by \([8,12]\)

\[
P_b \leq \frac{1}{\pi} \int_{0}^{\pi/2} \left( \frac{1}{n_b} \frac{\partial}{\partial I} T(D(\theta),I) \right)_{I=1} d\theta,
\]

where \( n_b \) is the number of input bits per transition and \( T(D(\theta),I) \) is the modified transfer function of the code, where \( D(\theta) \), is given in our case, by

\[
D(\theta) = \left( 1 + \frac{\Omega^2}{\sin^2 \theta} \right)^{-MN} \exp\left( -MN\Delta_f^2(\theta) \right) \times \\
\left[ 1 + \sum_{k=2}^{\infty} \frac{(k-1)!!}{k!} (2\sigma^2)^k \sum_{d=1}^{\infty} g_{k,d}(\Delta_f^2(\theta))^d \right]^{MN}
\]

based on the derived PEP in (17).

4. EXAMPLES

In this section, we consider two different TCM schemes as outer codes whose trellis diagrams are illustrated in Figure 1. These are 2-state 8-PSK-MTMC codes with 2 symbols per branch, which are optimized for best performance over AWGN and Rayleigh fading channels, respectively [12]. For convenience, we summarize the important parameters of these codes from [12]. The free distance of the code A2 is \( d_{\text{free}}^2 = 3.172 \). Its minimum EL is determined by the error event path of \( \{s_0,s_4\} \), which differs by one symbol from the correct path (the all-zeros path is assumed to be the correct path based on the uniform properties of the code) achieving

| Parameter | Light       | Average    | Heavy       |
|-----------|-------------|------------|-------------|
| \( \sigma^2 \) | 0.158       | 0.126      | 0.0631      |
| \( m_{\mu} \) | 0.115       | -0.115     | -3.91       |
| \( \sigma_{\mu} \) | 0.115       | 0.161      | 0.806       |

EL = 1. The corresponding PD is \( d_2^2 = 4 \). On the other hand, the code F2 has a free distance of \( d_{\text{free}}^2 = 2.343 \) and it achieves EL = 2 with a product distance of \( d_2^2 \times d_3^2 = 2 \), which is determined by the error event path of \( \{s_1,s_5\} \). Since EL is the primary factor affecting performance (PD as a secondary factor) over fading channels, F2 is expected to have better performance than A2.

As an example of the shadowed Rician model, we consider the Canadian mobile satellite channel [6]. Table 1 shows the values of shadowing parameters for this channel, which are determined by empirical fit to measured data within Canada. In this table, the terms light, average, and heavy are used to represent an increasing effect of the shadowing.

The upper bounds for both codes with the single transmit antenna are illustrated in Figure 2. No STBC is considered in this case. As expected for the single transmit antenna case, F2 performs better than A2, where the performance is determined by the choices of EL and PD. This observation holds for all considered degrees of shadowing.

In Figure 3, upper bounds for the concatenated scheme are illustrated. Here we use the STBC designed for 2-TX antenna (i.e., Alamouti’s code). Based on this code, we have \( P = L = M = 2 \) and \( \beta = 1 \). Our results demonstrate that
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Figure 2: Upper bounds for codes A2 and F2 with single transmit antenna over shadowed Rician channels (1-TX and 1-RX antenna).

the concatenated schemes using A2 and F2 as outer trellis codes achieve roughly the same performance. This is a result of the fact that the dominant factors for the single antenna case no longer determine performance. In the 2-TX antenna case, both schemes achieve GEL equal to 2 and GPD equal to 4, that is, $\left(\frac{d_1^2 + d_5^2}{2}\right)^2 = 4$ for F2 and $\left(\frac{d_0^2 + d_4^2}{2}\right)^2 = 4$ for A2, based on (23). Since both of them have equal GEL and GPD, their performances turn out to be almost identical. This observation holds to be true independent of considered degrees of shadowing.

Comparison between the one- and two-transmit-antenna cases also reveals interesting points on the performance. In both figures, code F2 gives a diversity order of 2 (i.e., slope of the curve), regardless of antenna numbers. Only an additional coding gain (i.e., horizontal shift in the curve) is observed with the use of two antennas. However, this result is somewhat a coincidence because of the particular choice of the parameters characterizing this specific example. For the single transmit antenna case, the code F2 has $EL = 2$ and the performance curve varies with $(Eb/N0)^{-2}$. On the other hand, for the 2-TX antenna case we have $|\Psi| = 1$, since an $L = 2$-symbol interleaver is used. However, the overall diversity is determined by GEL (i.e., $|\Psi|NM = 1 \cdot 1 \cdot 2 = 2$), resulting again in the same slope as in the single transmit antenna case.

To examine the tightness of upper bounds, we also evaluate the performance of codes A2 and F2 through computer simulation, assuming 2-TX antennas. Simulation results for the code F2 are illustrated in Figure 4 with the corresponding

Figure 3: Upper bounds for concatenated MTCM-STBC schemes with codes A2 and F2 as outer codes over shadowed Rician channels (2-TX and 1-RX antenna).

Figure 4: Upper bounds versus simulation results for code F2 (solid: upper bounds, dashed: simulation).
upper bounds (plotted as solid lines) computed by (24) and (25). The upper bounds are in very good agreement with simulation results, demonstrating the tightness of the new upper bounds based on the exact PEP. As expected (based on our previous discussion on upper bound expressions), code A2 yields nearly identical simulation results to those of code F2, which we do not include here for brevity.

5. CONCLUSION
We analyzed the performance of trellis-coded STBC schemes over shadowed Rician fading channels. Our analysis is based on the derivation of an exact PEP through the moment generating function approach. The derived expression provides insight into the selection criteria for trellis codes which should be used in conjunction with STBC over fading channels. Our results also show that the trellis codes designed for optimum performance over Rician channels with single transmit antenna are not necessarily optimum for the multiple transmit antenna case. Using transfer function techniques based on the new PEP, we present upper bounds on the bit error probability for the concatenated scheme. We also provide simulation results, which seem to be in good agreement with the derived upper bounds.

APPENDIX
This appendix evaluates the inner integral in (16) in terms of an infinite sum. Defining

\[ a = \frac{1}{2\sigma^2} \frac{\Omega_f}{1 + \Omega_f \sin^2 \theta}, \quad b = 2\sqrt{2a_f}, \quad c = 2m_{\mu}, \]

we can write the inner integral in (16) as

\[ \int_{-\infty}^{\infty} \exp \left( -u^2 \right) f(u) du = \sqrt{\pi} \sum_{k=0}^{\infty} \frac{f^k(0)}{k!} \frac{(k-1)!!}{2^{k/2}}. \]

Using the integral form given by [11, page 382, equation 3.462.1], it can easily be shown that the integral in (A.3) is zero for the odd values of \( k \). For even values of \( k \), we can use the result [11, page 382, equation 3.461.4] and express (A.3) as

\[ \int_{-\infty}^{\infty} \exp \left( -u^2 \right) f(u) du = \sqrt{\pi} \sum_{k=0}^{\infty} \frac{f^k(0)}{k!} \frac{(k-1)!!}{2^{k/2}}. \]

Replacing (A.2) by (A.6) with \( a, b, \) and \( c \) values given as in (A.1), one can obtain the final form for the inner integral of (16) leading to (17).

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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

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:

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Special Issue on Femtocell Networks

Call for Papers

Recently, there has been a growing interest in femtocell networks both in academia and industry. They offer significant advantages for next-generation broadband wireless communication systems. For example, they eliminate the deadspots in a macrocellular network. Moreover, due to short communication distances (on the order of tens of meters), they offer significantly better signal qualities compared to the current cellular networks. This makes high-quality voice communications and high data rate multimedia type of applications possible in indoor environments.

However, this new type of technology also comes with its own challenges, and there are significant technical problems that need to be addressed for successful deployment and operation of these networks. Standardization efforts related to femtocell networks in 3GPP (e.g., under TSG-RAN Working Group 4 and LTE-Advanced) and IEEE (e.g., under IEEE 802.16m) are already underway.

The goal of this special issue is to solicit high-quality unpublished research papers on design, evaluation, and performance analysis of femtocell networks. Suitable topics include but are not limited to the following:

- Downlink and uplink PHY/MAC design for femtocells in 3G systems, WiMAX systems, and LTE systems
- Interference analysis, avoidance, and mitigation
- Coexistence between a macrocellular network and femtocell network
- Resource allocation techniques
- Closed subscriber group (CSG) versus open-access femtocells
- Power control and power saving mechanisms (e.g., sleep/idle mode etc.)
- Mobility support and handover
- Time synchronization
- Multiple antenna techniques
- Tradeoffs between femtocells, picocells, relay networks, and antenna arrays
- Comparison with other fixed-mobile convergence (FMC) approaches such as UMA/GAN and dual-mode terminals
- Self-organizing networks and issues in self maintenance and self install
- Issues related to enterprise femtocells

Before submission, authors should carefully read over the journal’s Author Guidelines, which are located at http://www.hindawi.com/journals/wcn/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:

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Special Issue on
WiMAX, LTE, and WiFi Interworking

Call for Papers

Although WiMAX, LTE, and WiFi provide wireless broadband connectivity, they have been optimized for different usage models: WiFi for very high-speed local area network connectivity and WiMAX, LTE for high-speed wireless cellular connectivity. By combining WiMAX, LTE, and WiFi technologies, service providers can offer better usability of the networks infrastructure and support for seamless mobility and roaming. The unique similarities between WiMAX, LTE, and WiFi networks that make the proposed synergy promising is these technologies are fully packet switching uses IP-based technologies to provide connection services to the Internet. This standards- and IP-based network approach provides compelling benefits to service providers to collaborate between these technologies.

This special issue is intended to foster state-of-the-art research in the area of WiMAX, WiBro, LTE, and WiFi networking, and the corresponding technical advances in the design and deployment of feasible network architectures and protocols, and to present novel results and solutions to solve various problems and challenges foreseen in WiMAX, LTE, and WiFi interworking. The special issue will cover the following topical areas but are not limited to them:

- WiMAX, WiBro, MobileFi, LTE, and WiFi communications systems
- Single/dual radio handover
- Network architecture alternatives for interworking and integration
- Heterogeneous wireless networks
- Seamless vertical handover and session continuity
- Multiradio coexistence and power management
- Authentication, authorization, and accounting
- Security issues
- Common charging and billing
- Quality of services (QoS)
- Interworking using IMS, SIP, MIH, VCC, and UMA
- IEEE802.11u, IEEE802.16g/j/m/h
- Scenarios and usage cases
- WiMAX, WiBro, LTE, and WiFi Interworking Testbed
- Hybrid wireless mesh network
- Applications, VOIP, video streaming, and so forth

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| Deadline                  | Date         |
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