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Contents

Front cover: Figure 6. The Helios NASA experimental heavier-than-air solar-powered HAP flight over Hawaii, at an altitude of around 21 km in 2001 [15].See paper by J. Gavan et al. pp. 20-31.

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The International Union of Radio Science (URSI) is a foundation Union (1919) of the International Council of Scientific Unions as direct and immediate successor of the Commission Internationale de Télégraphie Sans Fil which dates from 1913.

Editorial

Our Papers

In this issue, we have two *Reviews of Radio Science* from Commission C, an introduction and preview to an upcoming special issue, and a historical look at one of the URSI Commissions as it approaches its 20th birthday.

The efforts of Mario Luise and Phil Wilkinson in bringing us the two Commission C *Reviews* are gratefully acknowledged.

It is well known that wireless communications, particularly in a network environment, can often be improved by using various types of diversity. The invited *Review of Radio Science* by A. Abrardo and G. Ferrari explores an interesting type of diversity: collaborative diversity. This involves a network in which two or more nodes need to transmit to a common destination (for example, an access point). The various nodes receive a message transmitted from one node to the destination, and themselves retransmit that message to the destination. This provides additional information to the destination, and may enable improved throughput and reduced error rate for the network. The authors begin with a survey of the types of communication schemes that have been previously explored to exploit such additional information. They introduce a simple scenario that they use to compare various approaches. This involves two source nodes with some correlation between them, communicating with a single access point. They consider four approaches to coding signals within this scenario to take advantage of the source correlation: a non-cooperative system, cooperative source coding, cooperative source-channel coding, and joint source-channel coding. These approaches differ based on using the source-correlation information at the sources, at the access point, or at both the source(s) and the access point. The results of comparisons among these four approaches yield considerable insight into which method is likely to maximize throughput and reduce error rate under various real-world conditions. The results are relevant to a wide variety of applications, including cellular networks, wireless local-area networks, and wireless sensor networks.

Platforms in the sky for communications, broadcasting, remote sensing, surveillance, and a host of other civilian and military applications have proven to be very valuable. Satellites at various altitudes (e.g., in geosynchronous orbit, or low-Earth orbit) have become widely used, but there are many tradeoffs among cost, lifetime, serviceability, power availability, payload, and other factors. The paper by Jacob Gavan, Saad Tapuchi, and David Grace provides an introduction to and overview of an interesting alternative: high-altitude platforms (HAPs). They begin with a discussion of the capabilities and tradeoffs of satellites at various altitudes and HAPs. They then provide an interesting overview of the different basic concepts associated with HAPs, including factors that suggest optimum operating altitudes, various power sources, and operating ranges for use of such platforms for radio relay applications. Considerations for HAPs that are heavier than air and that are lighter than air are discussed. This is followed by an examination of the origin and development of HAPs. Three time periods are covered: prior to 1990, 1990 to the present, and what can be expected in the near future. In each case, examples are given of HAPs designs and experiments. The capabilities and tradeoffs of the various systems are explained. It is shown that there are a number of instances where both commercial and government organizations either have recently developed and deployed successful HAPs, or will do so in the very near future. A list of potential and planned HAPs applications, with references is given. The conclusions provide some additional insight into the likely future of HAPs. This paper is intended as an introduction and overview of a special issue of the *Radio Science Bulletin* that will be devoted to HAPs and various aspects of implementation and applications.

Galileo is the European global satellite navigation system. Because it has been developed more recently that most of the other satellite navigation systems in operation or planned, it can benefit from lessons learned from the operation of other systems, as well as from new technology. In their invited *Review of Radio Science*, Olivier Julien, Christophe Macabiau, and Jean-Luc Issler look at the signals to be used on Galileo. They begin with a brief introduction to the use of (particularly, the civil - or civilian) signals for navigation and positioning on satellite navigation systems. They then the signal structure used for GPS (Global Positioning System). This includes an explanation of how the signals were chosen (and why), how they are used, and some of the tradeoffs associated with the signal structure. The authors then provide an overview of signal processing for global navigation satellites. This includes models of the correlation operations, phase tracking, code-delay tracking, and the critical elements and overall performance of each of these. They then summarize the major constraints the satellites' signals impose on the performance of the systems. This is followed by a detailed look at the signal structure and operation chosen for Galileo, including the innovations incorporated and the reasons behind them. This includes the properties and lengths of the codes used, including secondary codes; the use of a data-less, or pilot, channel; and the use of binary offset carrier modulation. An overview of Galileo's services and frequency plan is presented. The signal structure, and the reasons behind the choices made, for each of Galileo's signals is explained. This is followed by an analysis of Galileo's signal performance, including phase tracking, code-delay tracking, and interoperability and compatibility with other systems. This paper is an interesting and readily understandable look at the performance and tradeoffs of the signals used in modern satellite navigation systems.

URSI's youngest Scientific Commission, Commission K, on Electromagnetics in Biology and Medicine, will celebrate its 20th birthday at the XXXth General Assembly and Scientific Symposium in Istanbul, Turkey, in 2011. The paper by James C. Lin, Paolo Bernardi, and Jørgen Bach Andersen, three of the people who were involved in Commission K from the beginning, provides a fascinating look at the formation and development of the Commission. It traces the origin of the Commission, and the factors that led to considering an URSI Commission on this topic. The way in which the Commission was organized at the XXIIIrd General Assembly in Prague, Czechoslovakia (now in the Czech Republic), in 1990 is documented. The development of the terms of reference, and how they have changed since the Commission was organized, are explained. The paper then traces the participation of the Commission in the General Assemblies of URSI over the first decade. The factors that developed in the field of the Commission, and their effects on the Commission's activities, are examined. The Commission's extensive participation in other international symposia, and its co-sponsorship of meetings with other organizations, is documented. References are provided to reports of these meetings, where available. The Commission's contributions to the *Review of Radio Science* books are given, along with a review of the URSI resolutions and recommendations to come out of the Commission.

This paper is a useful and interesting insight into the history of Commission K. I think it would be quite nice to have similar histories prepared for the other URSI Commissions, and I invite those among the readers of the *Radio Science Bulletin* who might be interested in preparing such a history to contact me.

Our Other Contributions

We also have two book reviews, under the editorship (and, in one case, authorship) of Kristen Schlegel. I urge you to read these. If you would like to serve as a reviewer, please contact Kristen (ks-ursi@email.de).

There are several announcements of meetings of interest to the URSI community, and of a fellowship competition in advanced electromagnetics. There is also a report on one of the workshops URSI co-sponsored.

Your Papers Are Welcome

The *Radio Science Bulletin* is always looking for good papers of general interest to radio scientists. Please send your contributions directly to me via e-mail at r.stone@ieee.org.

Ross

URSI Commission B
2010 International Symposium on Electromagnetic Theory – EMTS 2010

Berlin, Germany, August 16-19, 2010

Scope

Commission B of the International Union of Radio Science organizes a triennial series of International Symposia on Electromagnetic Theory. The 2010 Symposium will be held in Berlin, Germany, August 16-19, at the Steigenberger Hotel, Berlin-Charlottenburg. The scope of the Symposium covers all areas of electromagnetic theory and its applications. The working language of the Symposium is English. There will be a limited number of Young Scientist Awards (YSA) available for application. All accepted papers will be available through IEEE Xplore.

Topics

Contributions concerning all aspects of electromagnetic theory and its applications are welcome. Other organized sessions will be indicated on the paper submission page of the Web site. Novel and innovative contributions are particularly appreciated.

Basic Electromagnetic Theory
R1 Electromagnetic theory
R2 Mathematical modeling of EM problems
R3 Solutions to canonical problems
R4 Nonlinear phenomena

Scattering and Diffraction
R5 Scattering and Diffraction
R6 High-frequency methods
R7 Inverse scattering and imaging

Random, Inhomogeneous, Non-linear and Complex Media
R8 Propagation and scattering in layered structures
R9 Random media and rough surfaces
R10 Complex media
R11 Beam and pulse propagation and scattering in lossy and/or dispersive media

Computational Techniques
R12 Numerical methods: general aspects
R13 Numerical methods for integral and differential equations
R14 Hybrid methods

Transient Fields
R15 Time-domain methods
R16 Radiation, scattering, and reception of transient fields and/or wideband signals

Guided Waves
R17 Guided waves

EMC/EMI
R18 Interaction of EM waves with biological tissues R19 Modeling techniques for EMC/EMI Antennas
R20 Antennas: general aspects
R21 Antenna arrays
R22 Smart antennas
R23 UWB antennas

Systems
R24 EM theory and applications for radio systems
R25 Antennas and propagation for communication systems: Mobile, LAN etc.

Others
R26 Others (please specify your specific field)

Special Topics
S1 Antennas for imaging systems at terahertz frequencies
S2 Nonstandard boundary conditions in electro-magnetics
S3 Analytic identities and limitations in electromagnetic theory

Organization

The Chair of the Local Organizing Committee is L. Klinkenbusch. J. Detlefsen and T. Eibert are the co-Chairs. Karl J. Langenberg is Chair of the Technical Program Committee.

Deadlines and Contact

The deadline for Young Scientist Award submissions is **January 15, 2010**. The deadline for the receipt of all papers is **January 31, 2010**. Information for the Young Scientist Awards and general paper submission, along with information on registration and reservations, can be found at http://www.emts2010.de. Inquiries can be sent by e-mail to info@emts2010.de.

A Perspective on Distributed Radio Systems with Cooperative Coding

A. Abrardo
G. Ferrari

Abstract

In distributed radio systems, e.g., wireless sensor or ad-hoc networks, the system's performance may be significantly enhanced by means of cooperation among the nodes. In this paper, we present a perspective on *cooperative coding* in distributed radio systems. In particular, we consider a simple reference scenario where two source nodes need to transmit two correlated information sequences to a common access point (AP). This is a scenario of interest for wireless sensor networks, where the sensor may observe correlated phenomena. Four possible cooperative-coding situations are described: (i) a non-cooperative system (NCS), (ii) cooperative source coding (CSC), (iii) cooperative source-channel coding (CSCC), and (iv) joint source-channel coding (JSCC). While in the first scheme source correlation is not used at all, the other schemes differ according to the way source correlation is used: from cooperative source coding systems, where source correlation is used only at the sources, to joint source-channel coding systems, where source correlation is used only at the access point. Indeed, joint source-channel coding systems are attractive in scenarios (such as wireless scenarios) where communications between the sources might be problematic. As an illustrative example, we will present a practical joint source-channel coding system in the presence of block-faded channels, using low-density parity-check (LDPC) coding at the sources and a proper iterative decoder at the access point.

1. Introduction and Motivation

In this paper, we focus on distributed communication systems where two or more nodes need to transmit to a common remote destination. This model applies to many scenarios, such as cellular networks, wireless local-area networks with one access point (AP), ad-hoc wireless networks, wireless sensor networks, etc. In these scenarios, collaboration between the nodes might bring significant advantages in terms of *collaborative diversity* [1]. In a cooperative system, each user is assigned one or more partners. The partners overhear each other, process the received signals, and retransmit proper messages to the destination in order to provide extra information to the access point with respect to the signal sent by a single source. Even in the presence of noisy inter-partner channels, the virtual transmitting-antenna array formed by cooperating nodes provides additional diversity, and may improve the system's performance in terms of error rate and throughput.

In the literature, many schemes have been proposed to exploit collaborative diversity. These schemes differ, especially for the relaying technique used, i.e., on the basis of the information that is re-transmitted by cooperating nodes to guarantee the highest ratio between diversity degree and resource consumption. The simplest schemes are those where the nodes re-transmit all the received information in an orthogonal way (typically, with time-division multiplexing): the codes used are not very efficient, but the highest diversity is guaranteed [2-5]. In other schemes, only a concise version of the information received by a cooperating node is transmitted, e.g., a parity bit [6]. Finally, there are

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This is one of the invited *Reviews of Radio Science* from Commission C.

schemes where the nodes simultaneously access the shared radio medium – typically modeled as a Gaussian multiple-access channel (GMAC) – with Alamouti-like space-time coding [7, 8]. In this case, the first direct transmission corresponds to the first row of the Alamouti code matrix (which, nevertheless, corresponds to transmissions at different moments, since the nodes cannot transmit and receive at the same time), while the simultaneous transmissions are associated with the second row. In the latter case, the nodes have only to transmit, so that the transmissions can be simultaneous. A scheme of this type allows a much higher efficiency than the previous schemes, since the multiple-access interference is completely eliminated, owing to the orthogonality of the Alamouti matrix; obviously, perfect synchronization between the nodes is required. In classical cooperation scenarios, the idea is therefore that of making the nodes cooperate among themselves to implement a distributed channel-coding scheme, where different nodes retransmit the same information, in some sense.

However, in many application scenarios the information that resides in different nodes is *intrinsically correlated*. In other words, even without implementing any cooperation among the nodes, the same or, more generally, "similar" information is transmitted by the nodes. A significant application example where this situation typically appears is given by *wireless sensor networks* [9]. In this case, the design of efficient transmission of correlated signals, observed at different nodes, to one or more collectors is one of the main design challenges. In the presence of one collector node, this system model is often referred to as a reach-back channel [10-12]. In its simplest form, this problem can be summarized as follows. Two independent nodes have to transmit correlated sensed data to a collector node by using the minimum possible energy, i.e., by exploiting in some way implicit correlation among the data. In the case of orthogonal additive white Gaussian noise (AWGN) channels, the separation between source and channel coding is known to be optimal [13, 14]. This means that the theoretical limit can be achieved by first compressing each source up to the Slepian-Wolf (SW) limit, and then utilizing two independent capacity-achieving channel codes (one per source) [15]. In this case, no cooperation among the source nodes is required.

However, if the transmissions are not carried out through separate additive white Gaussian noise channels, then the Slepian-Wolf approach is no longer optimal. Alternative schemes, which encompass the presence of cooperating nodes, can bring significant advantages. In [16], a Gaussian multiple-access channel scheme was considered. First, the nodes exchange information through time-division-based transmission acts and taking into account the correlation, i.e., they transmit much less with respect to the entire information by relying on a distributed-source coding-based approach. Once each source node has the entire information (relative to both source nodes), it then compresses and retransmits it to the destination node (the access point), thus achieving a beamforming gain with respect to a scenario with no cooperation. This approach basically requires that the information is first compressed, thus exploiting the correlation between the sources, and then duplicated in order to obtain a coding and diversity, or beamforming, gain.

An alternative solution to exploit the correlation in this scenario is based on joint-source channel-coding (JSCC) schemes, where no cooperation among nodes is required, and the correlated sources are not source encoded but only channel encoded. The absence of direct cooperation between the source nodes is attractive in scenarios where the communication links between the source nodes may be noisy. If one compares a joint-source channel-coding system with a system based on source/channel-coding separation with the same information rate, the channel codes used in a joint-source channel-coding scheme must be less powerful (i.e., they have higher rates). This weakness can be compensated for by exploiting the correlation between the sources at the decoder, which jointly recovers the information signals by both source nodes, so that the final performance can approach the theoretical limits. For this reason, this approach is also referred to as joint channel decoding (JCD). This approach has attracted the attention of several researchers in the recent past, also because of its implementation simplicity [17-21]. Note that in the joint-source channel-coding approach, the sources are encoded independently of each other (i.e., for a given source, neither the realization from the other source nor the correlation model are available at the encoder), and transmitted through the channel. Correlation between the sources must be instead assumed to be known at the (common) receiver.

In this context, the scheme proposed in [16], with "intrinsic" cooperation, is of particular interest. The basic idea is (i) to let the two sources transmit simultaneously two correlated code words, and (ii) to let the decoder solve the bit erasures (which appear when the transmitted bits are different). In [16], it was shown that this scheme may achieve a 3 dB potential beamforming gain, but it requires perfect synchronization and perfect channel-state information (CSI) at the transmitters. A scheme of this type does not therefore seem feasible in the presence of a channel affected by multipath fading, where guaranteeing perfect channel-state information at the transmitters requires a supplementary signaling load, which cannot be sustained, in many cases. Hence, in the presence of multipath fading, the problem of designing suitable (and reliable) non-cooperative joint source-channel coding schemes so that an "intrinsic diversity gain" can be achieved at the decoder by exploiting the side information (i.e., the a priori correlation between the information sequences) is still an open issue.

In this work, we present a simple perspective on cooperative-coding strategies for distributed radio systems. For ease of derivation, we will introduce a simple reference scenario (two source nodes and a common access point), which will allow us to analytically evaluate the performance of various cooperative-transmission schemes in the presence or absence of explicit cooperation. In all cases, we consider the presence of block-faded channels and power control – under the constraint of maximum transmitted power. In particular, we first analyze various schemes where the source correlation is exploited at the source nodes and/or at the access point. Considering a joint-source channel-coding scheme in the case of orthogonal multiple access, we will introduce the concept of *correlation-induced* diversity gain, to be compared with *cooperation-induced* diversity gain. Our results show that in many cases, the presence of correlation between sources limits the necessity for explicit cooperation between them. Finally, we will present a "practical" example of a low-density parity-check (LDPC) coded joint-source channel-coding scheme.

2. Two-Source-Node Scenario

We consider the distributed radio system shown in Figure 1. The correlated information sequences at the source nodes are indicated by $\boldsymbol{x}$ and $\boldsymbol{y}$, respectively. This scenario may correspond to a scheme where two sensor nodes, denoted as SN_1 and SN_2, detect the two correlated signals $\boldsymbol{x}$ and $\boldsymbol{y}$. These signals are assumed to be independent and identically distributed (i.i.d.) correlated binary random variables, with

$$Pr\left(x_i = 0\right) = Pr\left(x_i = 1\right) = 0.5,$$

$$Pr\left(y_i = 0\right) = Pr\left(y_i = 1\right) = 0.5,$$

for $i = 0, \ldots, k-1$ and $\rho \triangleq Pr\left(x_i = y_i\right) > 0.5$.

The information signals, which are assumed to be detectable without error (i.e., ideal sensor nodes), must be delivered to the access point. To this aim, each sensor node establishes a direct link toward the access point and an indirect link toward the other sensor node, in order to exploit cooperative transmission. We assume that transmissions from the nodes to the access point and between nodes occur over orthogonal channels (e.g., by using time-division multiple access). Moreover, we assume that the communication links are all affected by multiplicative fading and additive white Gaussian noise. Referring to the equivalent low-pass signal representation, and considering digital transmission, we denote as $\boldsymbol{s}_x$ ($\boldsymbol{s}_y$) the complex transmitted sequence corresponding to the information signal $\boldsymbol{x}$ ($\boldsymbol{y}$), with α being the complex link-gain term, which encompasses both path loss and fading, and with $\boldsymbol{n}$ being the complex additive white Gaussian noise sequence.

Each source node transmits N symbols every k information symbols, so that $r \triangleq k/N$ corresponds to the effective transmission rate of each source node (this might encompass the presence of both source and channel coding, as will be described later). In more general terms, the problem at hand consists of transmitting $2k$ information symbols through $2N$ channel uses. Each source node can communicate to the access point and to the other source node. We assume that all transmissions are performed in an highly scattered environment, without line of sight (NLOS). Hence, the channel complex gains, denoted as α_x (direct link from SN_1 to the access point), α_y (direct link from SN_2 to the access point), α_{xy} (direct link from SN_1 to SN_2, and α_{yx} (direct link from SN_2 to SN_1) can be modeled as zero-mean Gaussian random variables (Rayleigh fading). The fading coefficients in the direct links, i.e., α_x and α_y, are supposed to be independent (this is reasonable if the two sensor nodes are more than a wavelength away), while the fading coefficients in the indirect (inter-sensor) links are supposed to be equal, i.e., $\alpha_{xy} = \alpha_{xy}$ (this is reasonable if only one carrier frequency is used for both transmissions and the source nodes are quasi-static). Slow fading and path loss are assumed to have the same statistical distribution for both direct links.

Note that in the presence of channel fluctuations, an optimal transmission scheme should encompass a joint power-control and link-adaptation mechanism to adapt both the transmission rates and the transmitted powers to the actual channel conditions, so that the ultimate capacity may be achieved. However, combining power control with link adaptation is a difficult task. Specifically, without knowing the transmission power beforehand, the SNR cannot be predicted, as it would be needed to choose the

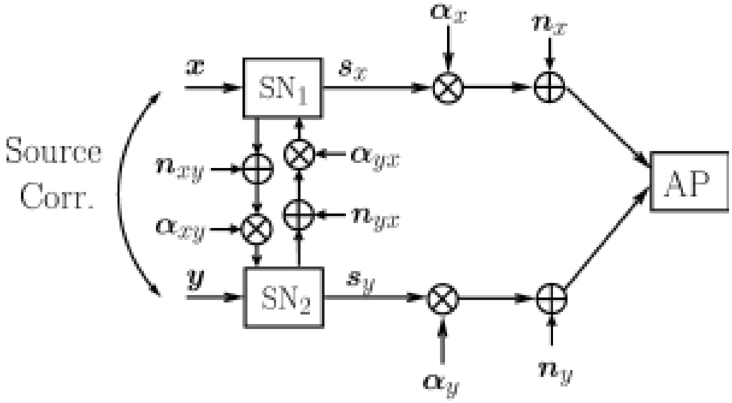

Figure 1. A distributed radio system where two source nodes need to transmit their information to a common access point and can communicate between themselves

Table 1. The main quantities used in the derivations

Quantity	**Definition**
$P_x = E\left(\lvert s_y\rvert^2\right)\left[P_y = \mathbb{E}\left(\lvert s_y\rvert^2\right)\right]$	power transmitted by SN_1 $[\mathrm{SN}_2]$ in the direct link
$P_{xy} = \mathbb{E}\left(\lvert s_{xy}\rvert^2\right)\left[P_{yx} = \mathbb{E}\left(\lvert s_{yx}\rvert^2\right)\right]$	power transmitted by SN_1 $[\mathrm{SN}_2]$ in the indirect link
R_x $[R_y]$	number of bits per detected sample transmitted by SN_1 $[\mathrm{SN}_2]$ in the direct link
R_{xy} $[R_{yx}]$	number of bits per detected sample transmitted by SN_1 $[\mathrm{SN}_2]$ in the indirect link
$G_x = \lvert\alpha_x\rvert^2$ $\left[G_y = \lvert\alpha_y\rvert^2\right]$	power gain term for the direct link of SN_1 $[\mathrm{SN}_2]$
$G_{xy} = \lvert\alpha_{xy}\rvert^2$ $\left[G_{yx} = \lvert\alpha_{yx}\rvert^2\right]$	power gain term for the indirect link of SN_1 $[\mathrm{SN}_2]$
$\sigma_d^2 = \mathbb{E}\left(\lvert n_x\rvert^2\right) = \mathbb{E}\left(\lvert n_y\rvert^2\right)$	AWGN variance in the direct links
$\sigma_i^2 = \mathbb{E}\left(\lvert n_{xy}\rvert^2\right) = \mathbb{E}\left(\lvert n_{yx}\rvert^2\right)$	AWGN variance in the indirect link
$\gamma_x = \frac{G_x}{\sigma_d^2}\left[\gamma_y = \frac{G_y}{\sigma_d^2}\right]$	SNR (normalized to the transmitted power) for the direct link of SN_1 $[\mathrm{SN}_2]$
$\gamma_{xy} = \frac{G_{xy}}{\sigma_i^2}\left[\gamma_{yx} = \frac{G_{yx}}{\sigma_i^2}\right]$	SNR (normalized to the transmitted power) for the indirect link of SN_1 $[\mathrm{SN}_2]$
$\Gamma_x = \Gamma_y = \mathbb{E}\left(G_x/\sigma_d^2\right)$	average SNR (normalized to the transmitted power) for the direct links
$\Gamma_{xy} = \Gamma_{yx} = \mathbb{E}\left(G_{xy}/\sigma_i^2\right)$	average SNR (normalized to the transmitted power) for the indirect links
$f_\gamma^{(y)}(u) = f_\gamma^{(x)}(u) = \frac{1}{\Gamma_x} e^{-\frac{u}{\Gamma_x}} U(u)$	common Rayleigh pdf of γ_x and γ_y
$f_\gamma^{(yx)}(u) = f_\gamma^{(xy)}(u) = \frac{1}{\Gamma_{xy}} e^{-\frac{u}{\Gamma_{xy}}} U(u)$	common Rayleigh pdf of γ_{xy} and γ_{yx}
$P_{max}^{(\mathrm{d})}$ $\left[P_{max}^{(\mathrm{i})}\right]$	maximum transmitted power in the direct [indirect] links, normalized to the inverse of the average SNR (i.e., $P_{max}^{(\mathrm{d})} = 1$ $\left[P_{max}^{(\mathrm{i})} = 1\right]$ means that the maximum power is equal to $1/\Gamma_x$ $\left[1/\Gamma_{xy}\right]$)

appropriate modulation/coding level. In turn, without knowing the modulation level, the transmitted power cannot be adjusted accordingly. Hence, in the following we assume that the transmission rates are fixed (i.e., no link adaptation), while the transmitted power can be adapted by means of an ideal closed power-control mechanism, which adjusts the transmitted power to a level sufficiently high to achieve desirable performance. In particular, the feedback power-adjustment messages sent by the access point are received without errors, and each sensor can set its transmitted power within a predefined range. The assumption of ideal power control in the presence of faded links may not be practical in the case of fast fading. In fact, the use of feedback power control requires the presence of very reliable return links (from the access point to the sources), to convey the channel-state information back to the transmitter, and a limited number of feedback commands. The impact of practical feedback-control strategies is beyond the scope of this paper (see [22] for more details). Instead, we take into account limitations in the transmitted power range by introducing $P_{max}^{(\mathrm{d})}$ and $[P_{max}^{(\mathrm{i})}]$ as the maximum transmitted power in the direct [indirect] links, normalized to the inverse of the average SNR (i.e., $P_{\max}^{(d)} = 1$ $[P_{\max}^{(i)} = 1]$ means that the maximum power is equal to $1/\Gamma_x$ $[1/\Gamma_{xy}]$).

For the sake of notational simplicity, in Table 1 we summarize the main quantities used in the remainder of this paper.

3. A Perspective on Possible Approaches

We now outline possible approaches to performing cooperative coding in distributed radio systems with correlated sources. These approaches, illustrated in Figure 2, can be summarized as follows:

- Non-cooperative system (NCS): the source correlation is *not* exploited at the sources and the access point;
- Cooperative source coding (CSC): the source correlation is exploited at the *sources*;
- Cooperative source-channel coding (CSCC): the source correlation is exploited at *both the source and the access point*;
- Joint source-channel coding (JSCC): the source correlation is exploited at the *access point*.

For each approach, the probability of incorrect delivery of the information signals from both sources, denoted as probability of error, will be derived. At the end, a comparative performance analysis is proposed.

3.1 First Approach: Non-Cooperative System

In this case, the two sensor nodes perform independent channel coding and the access point performs independent channel decoding, i.e., the source is not exploited at all. In this case, $R_x = R_y = 1$, i.e., the number of bits that must be transmitted by each sensor node every N channel uses is k. There is no multiple-access interference, i.e., the direct links are orthogonal (for example, time-division multiple access is considered). According to the Shannon channel-capacity formula, the maximum *common* transmit rate has to satisfy the following expressions:

$$r = \frac{1}{2}\log_2\left(1+\frac{P_x G_x}{\sigma_d^2}\right),$$

$$r = \frac{1}{2}\log_2\left(1+\frac{P_y G_y}{\sigma_d^2}\right),$$

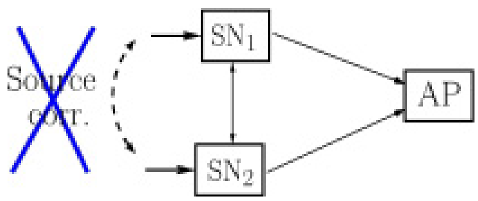

Figure 2a. A possible approach to exploiting the source correlation in a two-source-node distributed radio system: a non-cooperative system (NCS)

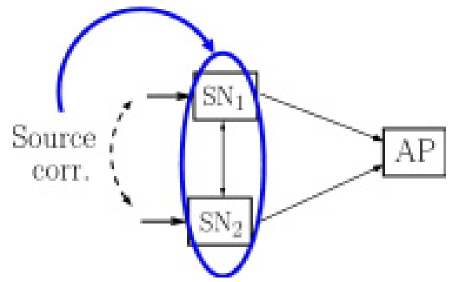

Figure 2b. A possible approach to exploiting the source correlation in a two-source-node distributed radio system: cooperative source coding (CSC).

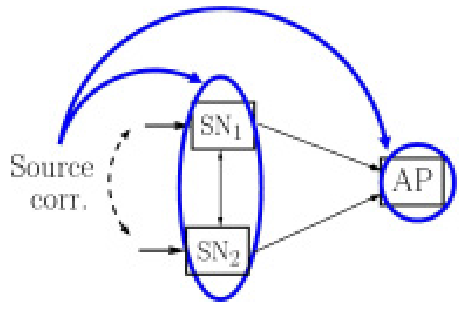

Figure 2c. A possible approach to exploiting the source correlation in a two-source-node distributed radio system: cooperative source-channel coding (CSCC).

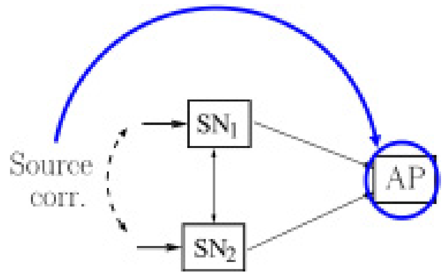

Figure 2d. A possible approach to exploiting the source correlation in a two-source-node distributed radio system: joint source-channel coding (JSCC).

from which the following expressions for the transmitted powers can be obtained:

$$P_x = \left(2^{2r} - 1\right)\frac{\sigma_d^2}{G_x} = \left(2^{2r} - 1\right)\frac{1}{\gamma_x},$$

$$P_y = \left(2^{2r} - 1\right)\frac{\sigma_d^2}{G_y} = \left(2^{2r} - 1\right)\frac{1}{\gamma_y}.$$

Given the distribution of the SNRs, γ_x and γ_y, and for the sake of notational simplicity defining the quantity $W_1 \triangleq 2^{2r} - 1$, the average transmitted powers become

$$\mathbb{E}(P_x) = \int_{\frac{W_1 \Gamma_x}{P_{max}^{(d)}}}^{\infty} \frac{W_1}{u}\frac{1}{\Gamma_x} e^{-\frac{u}{\Gamma_x}} \mathrm{d}u = \frac{W_1}{\Gamma_x}\int_{\frac{W_1}{P_{max}^{(d)}}}^{\infty} \frac{1}{u} e^{-u} \mathrm{d}u,$$

$$\mathbb{E}\left(P_y\right) = \mathbb{E}(P_x).$$

The total average energy (normalized to the symbol time) is

$$E_t = 2N\,\mathbb{E}(P_x).$$

Eventually, it is possible to evaluate the probability that a correct delivery of both information signals is not possible because of limited power resources. This probability, denoted as Pr_E, is the complement of the probability that both transmissions can be successfully carried out, i.e.,

$$Pr_E = 1 - \left(\int_{\Gamma_x \frac{W_1}{P_{max}^{(d)}}}^{\infty} \frac{1}{\Gamma_x} e^{-\frac{u}{\Gamma_x}} \mathrm{d}u\right)^2 = 1 - e^{-\frac{2W_1}{P_{max}^{(d)}}}.$$

3.2 Second Approach: Cooperative Source Coding

In this case, the two sensor nodes perform cooperative source coding and independent channel coding, whereas the access point performs independent channel decoding. It is worth to note that owing to the Slepian-Wolf theorem, ideal cooperative source coding can be achieved without any transmission between the two source nodes, i.e., without using the indirect links. Note that following the assumption of fixed transmitting rates, the transmitters are forced to select a compression rate that corresponds to one corner of the Slepian-Wolf region, leaving to closed-loop power control full responsibility for combating fading fluctuations. In this scenario, assuming that SN_1 transmits the information signal at full rate $R_x = 1$ (i.e., no compression is performed), SN_2 may transmit the "$\boldsymbol{y} \mid \boldsymbol{x}$" information at the following rate:

$$R_y = R_{y|x} = H\left(y \mid x\right)$$

$$= -\log_2\left(\rho\right)\rho - \log_2\left(1-\rho\right)\left(1-\rho\right).$$

As expected, the numbers of information bits that must be delivered by the two sensor nodes in this case are different. More precisely, $R_x > R_y$. Hence, the solution of assigning the same number of samples, *N*, to each source node is not the optimal solution. Denote by η ($0 \le \eta \le 1$) the percentage of time assigned to SN_1: the use of time-division multiple access (i.e., orthogonal channels) is assumed in this case, as well. In this case, taking into account that the access point performs independent decoding per each direct link, one obtains

$$\frac{r}{2\eta} = \frac{1}{2}\log_2\left(1 + P_x \gamma_x\right),$$

$$\frac{r R_{y|x}}{2(1-\eta)} = \frac{1}{2}\log_2\left(1 + P_y \gamma_y\right),$$

from which the following expressions for the transmitted powers can be derived:

$$P_x = \left(2^{\frac{r}{\eta}} - 1\right)\frac{1}{\gamma_x},$$

$$P_y = \left(2^{\frac{r R_{y|x}}{1-\eta}} - 1\right)\frac{1}{\gamma_y}.$$

Following the same considerations as the previous section, and for the sake of notational simplicity defining $W_2 \triangleq 2^{r/\eta} - 1$ and $W_3 \triangleq 2^{r R_{y|x}/(1-\eta)} - 1$, it is straightforward to derive the following expressions for the average transmitted powers:

$$\mathbb{E}(P_x) = \frac{W_2}{\Gamma_x}\int_{\frac{W_2}{P_{max}^{(d)}}}^{\infty} \frac{1}{u} e^{-u} \mathrm{d}u,$$

$$\mathbb{E}\left(P_y\right)=\frac{W_3}{\Gamma_x}\int_{\frac{W_3}{P_{max}^{(d)}}}^{\infty}\frac{1}{u}e^{-u}\,\mathrm{d}u\cdot$$

The total average energy (normalized to the symbol time) can be finally written as

$$E_t=2N\left[\eta\mathbb{E}\left(P_x\right)+\left(1-\eta\right)\mathbb{E}\left(P_y\right)\right]\cdot$$

The error probability, Pr_E, in this case is

$$Pr_E=1-\exp\left\{-\frac{W_2+W_3}{P_{max}^{(d)}}\right\}\cdot$$

3.3 Third Approach: Cooperative Source-Channel Coding

In this case, the two sensor nodes cooperate to exploit the potential benefit brought about by the use of space-time coding for multiple-input multiple-output (MIMO) systems. According to this approach, originally proposed in [16], two source nodes need to transmit the same information bits[1]. Hence, as a first step, SN_1 transmits its information signal, $\boldsymbol{x}$, to SN_2. SN_2 then needs to transmit its information signal, $\boldsymbol{y}$, to SN_1. Of course, by invoking the Slepian-Wolf theorem, these transmissions can be performed at the reduced rates $R_{x|y}$ and $R_{y|x}$, respectively. Note that due to the symmetry of the problem, $R_{x|y}=R_{y|x}$.

Assume now that the two indirect inter-source transmission steps take a percentage of time δ (with respect to the overall time spent to transmit to the access point), i.e., $\delta/2$ for the transmission from SN_1 to SN_2, and $\delta/2$ for the other transmission. In this case, one can write:

$$\frac{rR_{x|y}}{\delta}=\frac{1}{2}\log_2\left(1+P_{xy}\gamma_{xy}\right),$$

$$\frac{rR_{y|x}}{\delta}=\frac{1}{2}\log_2\left(1+P_{yx}\gamma_{yx}\right),$$

which yields

$$P_{xy}=\left(2^{\frac{2rR_{x|y}}{\delta}}-1\right)\frac{1}{\gamma_{xy}},$$

$$P_{yx}=\left(2^{\frac{2rR_{y|x}}{\delta}}-1\right)\frac{1}{\gamma_{yx}}\cdot$$

Following the same considerations as the previous sections, for the sake of notational simplicity defining $W_4\triangleq 2^{2rR_{x|y}/\delta}-1$ and $\vartheta=\Gamma_{xy}/\Gamma_x=\Gamma_{yx}/\Gamma_y$, and noting that $2^{rR_{x|y}/\delta}-1=2^{rR_{y|x}/\delta}-1$, the average transmitted powers can be straightforwardly expressed as

$$\mathbb{E}\left(P_{xy}\right)=\frac{W_4}{\Gamma_{xy}}\int_{\frac{\vartheta W_4}{P_{max}^{(d)}}}^{\infty}\frac{1}{u}e^{-u}\,\mathrm{d}u,$$

$$\mathbb{E}\left(P_{yx}\right)=\mathbb{E}\left(P_{xy}\right)\cdot$$

Following the same derivation as in the previous section, the probability of correct transmission in the indirect links, denoted as $Pr_C^{(\mathrm{i})}$, can be expressed as

$$Pr_C^{(\mathrm{i})}=\exp\left\{-\frac{2\vartheta W_4}{P_{max}^{(d)}}\right\}\cdot$$

After this first transmission step, the two nodes can send the same information signal, $(\boldsymbol{x},\boldsymbol{y})$ at the rate $1+R_{x|y}$ by using a MIMO transmission scheme, thus achieving the ideal MIMO capacity; obviously, the two source nodes have to be synchronous. We remark that the multiple-access interference can be completely eliminated due to the structure of the Alamouti matrix. The MIMO capacity formula for the 2×1 transmission scheme depends on the correlation properties of the channel vector (α_x,α_y). On the basis of the independence assumption introduced in Section 2, it can immediately be concluded that the capacity is the same as the two-degree diversity case. Hence, observing that the remaining transmissions in the direct links can be performed simultaneously by the two nodes in the remaining $1-\delta$ fraction of time, and that the two nodes use the same transmission power, $P_t/2$ (water-filling transmission schemes are not considered), one finally obtains

$$\frac{r\left(R_x+R_{x|y}\right)}{2\left(1-\delta\right)}=\frac{1}{2}\log_2\left[1+\frac{P_t\left(\gamma_x+\gamma_y\right)}{2}\right],$$

from which it follows that

$$P_t=\left[2^{\frac{r\left(R_x+R_{x|y}\right)}{2(1-\delta)}}-1\right]\frac{2}{\gamma_x+\gamma_y}\cdot$$

In order to evaluate the average transmission power, it is now necessary to derive the probability density function (pdf) of the sum of the SNRs, $\gamma_x + \gamma_y$. By the independence assumption, it can be observed that $\gamma_x + \gamma_y$ has a central chi-square distribution with four degrees of freedom and a mean of $\Gamma_x + \Gamma_x$. Hence, its probability density function is

$$f_{\gamma_x+\gamma_y}(u) = \frac{u}{\left(\Gamma_x + \Gamma_y\right)^2} e^{-\frac{u}{\Gamma_x+\Gamma_y}} = \frac{u}{4\Gamma_x^2} e^{-\frac{u}{2\Gamma_x}} .$$

Observing that the constraint on the maximum power in this case is $P_t \leq 2P_{max}^{(d)} / \Gamma_x$, and defining the quantity $W_5 \triangleq 2^{r\left(R_x+R_{x|y}\right)/\left[2(1-\delta)\right]} - 1$, the average transmitted power, P_t, can be expressed in closed form as follows:

$$\mathbb{E}\left(P_t\right) = \int_{\frac{\Gamma_x W_5}{P_{max}^{(d)}}}^{\infty} \frac{2W_5}{u} \frac{u}{4\Gamma_x^2} e^{-\frac{u}{2\Gamma_x}} \,\mathrm{d}u$$

$$= \frac{W_5}{\Gamma_x} \int_{\frac{W_5}{2P_{max}^{(d)}}}^{\infty} e^{-u} \,\mathrm{d}u$$

$$= \frac{W_5}{\Gamma_x} e^{-\frac{W_5}{2P_{max}^{(d)}}} .$$

The probability of correct transmission in the direct link can then be computed as follows:

$$Pr_C^{(\mathrm{d})} = Pr\left\{P_t \leq 2\frac{P_{max}^{(d)}}{\Gamma_x}\right\}$$

$$= \int_{\frac{\Gamma_x W_5}{P_{max}^{(d)}}}^{\infty} \frac{u}{4\Gamma_x^2} e^{-\frac{u}{2\Gamma_x}} \,\mathrm{d}u$$

$$= e^{-\frac{W_5}{2P_{max}^{(d)}}} \left(1 + \frac{W_5}{2P_{max}^{(d)}}\right) .$$

Taking into account all (direct and indirect) transmissions, the total average energy is

$$E_t = 2N\left[\delta\mathbb{E}\left(P_{xy}\right) + (1-\delta)\mathbb{E}\left(P_t\right)\right],$$

and the probability of error can be expressed as

$$Pr_E = 1 - Pr_C^{(\mathrm{i})} P_C^{(\mathrm{d})} .$$

3.4 Fourth Approach: Joint Source-Channel Coding

In this case, the two transmitters perform independent channel coding at the same (compressing) rate, $R = 1$. In other words, the correlation is not exploited at the transmitters, while the access point performs joint channel decoding, i.e., the correlation is exploited at the receiver. In the case of orthogonal direct channels, the achievable (common for both source nodes) channel rate, *r*, has to satisfy the following inequalities [21]:

$$rR_{y|x} \leq \frac{1}{2}\log_2\left(1 + \frac{P_x G_x}{\sigma_d^2}\right),$$

$$rR_{x|y} \leq \frac{1}{2}\log_2\left(1 + \frac{P_y G_y}{\sigma_d^2}\right),$$

$$r\left(1 + R_{x|y}\right) - \frac{1}{2}\log_2\left(1 + \frac{P_x G_x}{\sigma_d^2}\right) - \frac{1}{2}\log_2\left(1 + \frac{P_y G_y}{\sigma_d^2}\right) \leq 0 .$$

By solving the above system of inequalities for given values of *r* and ρ (equivalently, for given values of *r*, $R_{x|y}$, and $R_{y|x}$), it is possible to determine the feasibility region of the system in the bi-dimensional normalized SNRs bi-dimensional space (γ_x, γ_y). In Figure 3, an illustrative example is given by the region with the curve denoted as "theoretical limit" as the lower bound, in a scenario with $r = 0.5$ and $\rho = 0.9$. More details of this figure will be presented later. At this point, the probability of error for a fixed maximum transmitted power can be computed through Monte Carlo simulations, according to the following steps:

1. Fix the maximum transmitted energy, E_t, (or, equivalently, the maximum transmitted power for a given value of *r*) from each node.
2. Generate two fading realizations, G_x and G_y.
3. Under the assumption of ideal power control, determine the possible couples of transmitted powers required for the current fading realizations, in order for the actual SNRs to belong to the feasibility region. Since there are

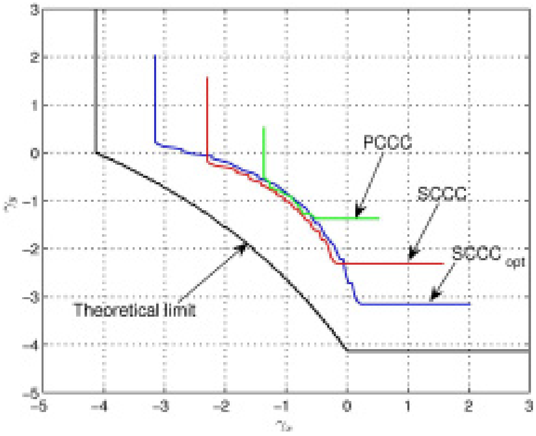

Figure 3. The feasible region (the lower bound of which is given by the curve denoted as "theoretical limit") in the bi-dimensional SNR space (γ_x, γ_y). The achievable regions, under the use of various channel-coding strategies, derived in [23] are also shown.

infinite couples of transmitted powers that allow the SNR couples to belong to the feasibility region, pick the couple of transmitted powers with minimum value of the maximum of the two.

4. The previously selected couple of transmitted powers will either satisfy the constraint on the maximum transmit power, or not. Under the assumption of ideal channel coding, if the constraint on the maximum transmitted power is satisfied, then there is no error; otherwise, there is an error.
5. By repeating the above steps a sufficiently large number of times, the probability of error can be numerically evaluated.

In the above derivation, we have assumed ideal channel coding. We now comment on this idealistic assumption. By relaxing the assumption of ideal channel coding, it is possible to evaluate the performance of specific channel codes with proper receiver structures. As mentioned in Section 1, there is intense research activity on this topic [17-21]. While in most of this work the presence of memory-less direct links was considered, in [24], various coded and uncoded schemes were analyzed in the presence of block-faded channels. In Section 4, an illustrative low-density parity-check-coded scheme, proposed in [24], will be recalled, and its performance evaluated. At this point, an interesting question is, "Does there exist an "optimal" channel code?" In order to answer this question, in [23] an innovative two-dimensional extrinsic information transfer (EXIT) chart-inspired optimization approach was proposed. This approach allows the determination of the feasibility region, in the channel SNRs bi-dimensional space, associated with the particular coding scheme under use. These results suggest that by properly designing serially concatenated convolutional codes (SCCCs), or parallel concatenated convolutional codes (PCCCs), the theoretical performance limits can be approached. In [23], preliminary results were shown in the presence of low-density parity-check coding, but their accurate optimization is one of our current research activities. In Figure 3, the feasibility regions in the case of additive white Gaussian noise communication channels[2] are shown, considering various turbo codes. The interested reader can find more details in [23]. Approaching the theoretical limits very closely remains an open problem.

3.5 Comparative Performance Evaluation

In Figure 4, the average transmitted energy, E_t, is shown as a function of ρ, considering the four possible types of schemes outlined before in this section, under the condition that $Pr_E = 0.01$. In the cooperative source-channel coding scheme, three values of θ, namely $\theta = -3$, $\theta = 0$, and $\theta = 3$ dB, are considered. In all cases, the average values $\Gamma_x = \Gamma_y = 10$ dB. Note that for $\theta > 0$ dB, i.e., when the SNR in the indirect link is higher than in the

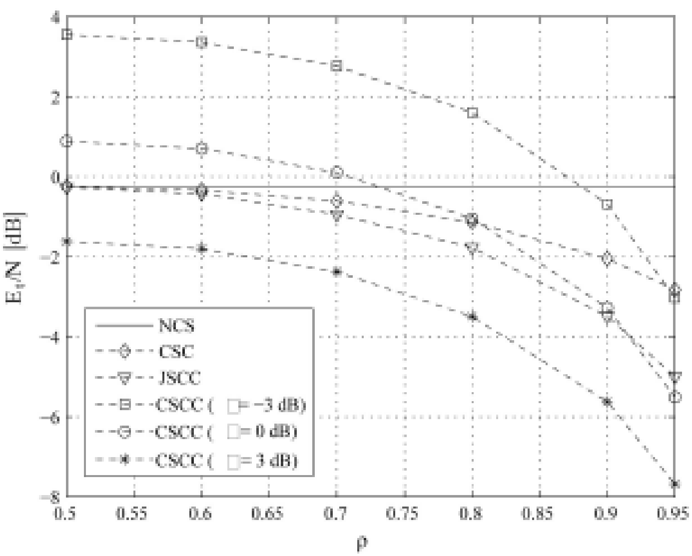

Figure 4. The average transmitted SNR, E_t/N, as a function of the source correlation coefficient, ρ, required to achieve $Pr_E = 0.01$. All four approaches are described in Sections 3.1-3.4. In all cases, $\Gamma_x = \Gamma_y = 10$ dB.

direct link, the cooperative source-channel coding scheme always outperforms the other schemes, even for very low values of the correlation coefficient ρ. Moreover, for highly correlated sources ($\rho \to 1$), the cooperative source-channel coding scheme always outperforms the non-cooperative system scheme by 14 dB, and the cooperative source-channel coding scheme by 10 dB, results that clearly show the potential benefits of cooperation. For $\theta < 0$ dB, i.e., when the SNR in the indirect link is lower than that in the direct link, then the performance of the cooperative source-channel coding scheme degrades remarkably. This is intuitive, since in this case the two sensors cannot reliably exchange information, and cooperation among them might cause even more errors (each sensor receives a "wrong suggestion" from the other one). It is worth noting that when $\rho \to 0$ (independent information signals), non-cooperative system and cooperative source-coding schemes are equivalent, and the cooperative source-channel coding scheme becomes a cooperative diversity scheme, similar to those already investigated in the literature (e.g., see [2-6]). In this case, if the channels from the nodes to the access point are the same as the channel between the nodes (i.e., $\theta = 0$ dB in our scenario), cooperation is not useful, and the non-cooperative system scheme slightly outperforms the cooperative source-channel coding scheme.

From the results shown in Figure 4, the following simple rules for the selection of the best scheme to be used, can be summarized as follows:

- If the sensors cannot cooperate with each other, then the joint source-channel coding scheme is to be preferred with respect to the cooperative source coding scheme (and, obviously, to the non-cooperative system): this is also motivated in this case by the impossibility of performing link adaptation.
- If the sensors can cooperate with each other and ϑ is sufficiently higher than 0 dB (i.e., the inter-sensor links have better quality than the direct links), then the cooperative source-channel coding scheme is to be preferred with respect to the joint source-channel coding scheme, regardless of the value of ρ.
- If the sensors can cooperate with each other but ϑ is lower than 0 dB (i.e., the inter-sensor links have worse quality than the direct links), then the cooperative source-channel coding scheme's performance degrades significantly, and the joint source-channel coding scheme is to be preferred, regardless of the value of ρ.
- If the sensors can cooperate with each other and ϑ is around 0 dB (i.e., the inter-sensor links have a quality approximately equal to that of the direct links), then the cooperative source-channel coding scheme is to be preferred only for very high values of ρ. Otherwise, the joint source-channel coding scheme is the way to go.

Although a cooperative source-channel coding scheme may guarantee a significant advantage with respect to a joint source-channel coding scheme (without cooperation), our results do not prove that considering separate source-channel coding is the best approach in the presence of cooperation between the nodes. Determining the "optimal" approach in the presence of cooperation is an interesting research direction.

4. A Practical Joint Source-Channel Coding Scheme with Distributed LDPC Coding

In this section, we propose a practical joint source-channel coding (or JCD) scheme based on the use of low-density parity-check codes. While the structure of the iterative decoder at the access point was originally proposed in a scenario with additive-white-Gaussian-noise direct links and two sources in [25, 26], this scheme has been extended to a scenario with a generic number of correlated sources and block-faded channels [24]. A scenario of this type, with M sources, is shown in Figure 5.

The information sequences are separately encoded using identical low-density parity-check codes and transmitted over the communication links. The common coding rate at the sources is $r = 1/2$. The proposed iterative

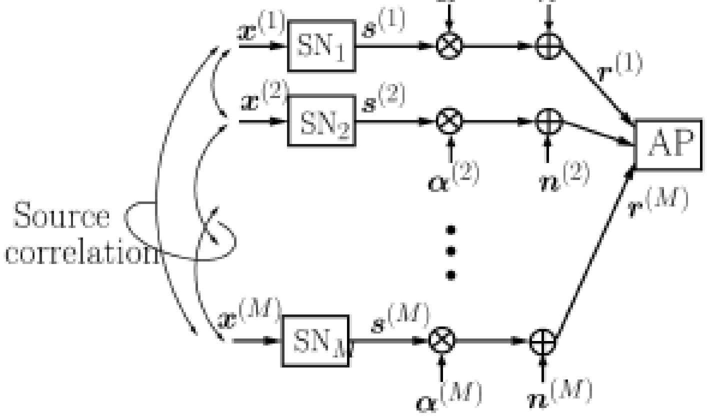

Figure 5. A scheme with a generic number of correlated sources and block-faded channels.

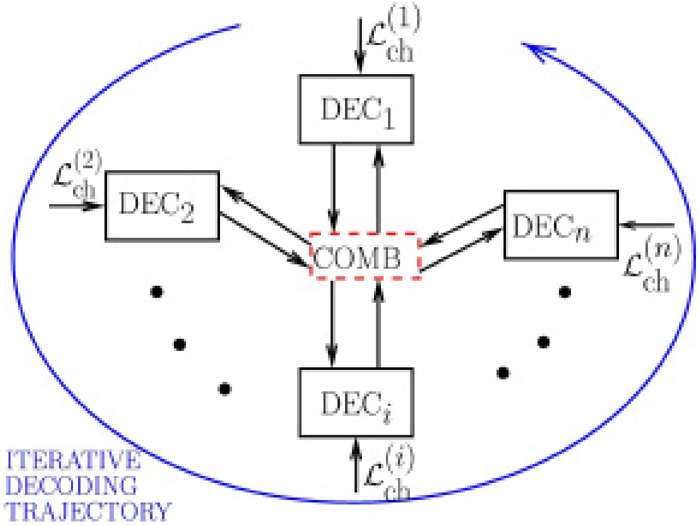

Figure 6. An iterative decoding scheme of correlated data in the absence of relay. Each component decoder, $\mathrm{DEC}_i, (i = 1, \ldots, n)$*, is a low-density parity-check decoder that receives both the channel a posteriori reliabilities and the a priori probabilities obtained by properly processing the soft-output reliability values generated by the other decoders. These processing/combining operations are carried out in the central block denoted as "COMB."*

decoding scheme at the access point is shown in Figure 6, where a low-density parity-check decoder per source is considered, and the trajectory of the iterative decoding process among these source decoders is highlighted: this decoding scheme is an extension of those discussed relative to two sources in [21]. Each low-density parity-check-coded sequence is decoded by using the classical sum-product algorithm [27]. Under the assumption of perfect channel-state information (CSI) at the receiver, the channel log-likelihood ratio (LLR) at the input of the *i*th variable node [27] can be expressed as

$$\mathcal{L}_{i,\mathrm{ch}}^{(k)} = \ln \frac{p\left[r_i^{(k)} \mid s_i^{(k)} = 1, \alpha_i^{(k)}\right]}{p\left[r_i^{(k)} \mid s_i^{(k)} = -1, \alpha_i^{(k)}\right]}$$

$$= \frac{2 r_i^{(k)} \sqrt{E_\mathrm{c}^{(k)}} \left|\alpha_i^{(k)}\right|}{\sigma^2}, \qquad (1)$$

where $\sigma^2 = N_0/2$ and N_0 is the variance of the additive white Gaussian noise sample. The maximum number of *internal* decoding iterations in each component low-density parity-check decoder is denoted as $n_{\mathrm{it}}^{\text{int-max}}$.

The a priori information about the correlation between the sources is exploited by applying the following *external* iterative decoding steps between the component low-density parity-check decoders: (i) the a posteriori reliability (i.e., the LLR) on the *information*[3] bits of the *j*th decoder is properly modified, taking into account the correlation (as will be explained later), and used as a priori reliability for the information bits at the input of the ℓ th decoder ($j \neq \ell$); (ii) at the first external iteration, the a priori reliability on the information bits at the input of the ℓ th decoder is obtained by properly modifying the a posteriori reliability of the *j*th decoder ($j < \ell$); (iii) the algorithm stops when a maximum number of external iterations (denoted as $n_{\mathrm{it}}^{\mathrm{ext}}$) is reached.

The total a posteriori reliability at the input of each variable node of the factor graph underlying the ℓ th low-density parity-check decoder can be expressed as follows:

$$\mathcal{L}_{i,\mathrm{in}}^{(\ell)} = \begin{cases} \mathcal{L}_{i,\mathrm{ch}}^{(\ell)} + \mathcal{L}_{i,\mathrm{ap}}^{(\ell)}, & i = 0, \ldots, N/2 - 1 \\ \mathcal{L}_{i,\mathrm{ch}}^{(\ell)}, & i = N/2, \ldots, N-1 \end{cases}.$$

In other words, besides the channel reliability value expressed as in Equation (1), the a posteriori reliability at the input of the variable nodes associated with the information bits ($i = 0, \ldots, N/2 - 1$) includes the "suggestion" (given by the soft reliability value $\mathcal{L}_{i,\mathrm{ap}}^{(\ell)}$) obtained from a posteriori reliability values output by the other decoders. In particular, the a priori component of the a posteriori reliability at the input of the ℓ th decoder can be written as

$$\mathcal{L}_{i,\mathrm{ap}}^{(\ell)} = \ln \frac{P\left[s_i^{(\ell)} = 1\right]}{P\left[s_i^{(\ell)} = -1\right]}, \quad i = 0, \ldots, L-1,$$

where

$P\left[s_i^{(\ell)} = \pm 1\right]$ are derived from the soft-output values generated by the other decoders, as follows. In a straightforward manner, one can rewrite $P\left[s_i^{(\ell)}\right]$ as

$$P\left[s_i^{(\ell)}\right] = \frac{1}{M-1} \underbrace{\left\{P\left[s_i^{(\ell)}\right] + \ldots + P\left[s_i^{(\ell)}\right]\right\}}_{M\text{-1 times}}. \qquad (2)$$

Using Bayes' theorem [28], the probability $P\left[s_i^{(\ell)}\right]$ can be expressed as

$$P\left[s_i^{(\ell)}\right] = \sum_{s_i^{(k)} = \pm 1} P\left[s_i^{(\ell)}, s_i^{(k)}\right]$$

$$= \sum_{s_i^{(k)}=\pm 1} P\left[s_i^{(\ell)} \mid s_i^{(k)}\right] P\left[s_i^{(k)}\right], \qquad (3)$$

$$k = 1,\ldots,N \text{ and } k \neq \ell \,.$$

Approximating the a priori probability $P\left[s_i^{(k)}\right]$ in Equation (3) with the a posteriori reliability value, denoted as $\hat{P}\left[s_i^{(k)}\right]$, output by the kth decoder ($k \neq \ell$), from Equation (3) one obtains

$$P\left[s_i^{(\ell)}\right] \simeq \sum_{s_i^{(k)}=\pm 1} P\left[s_i^{(\ell)} \mid s_i^{(k)}\right] \hat{P}\left[s_i^{(k)}\right]$$

where

$$\hat{P}\left[s_i^{(k)}\right] = \begin{cases} \dfrac{e^{\mathcal{L}_{i,\text{out}}^{(k)}}}{1+e^{\mathcal{L}_{i,\text{out}}^{(k)}}} & \text{if } s_i^{(k)} = +1 \\ \dfrac{1}{1+e^{\mathcal{L}_{i,\text{out}}^{(k)}}} & \text{if } s_i^{(k)} = -1 \end{cases}.$$

At this point, we evaluate the conditional probability $P\left[s_i^{(\ell)} \mid s_i^{(k)}\right]$ in Equation (3) using the *a priori* distribution (rather than a posteriori reliability values). By using Bayes' theorem, it follows that

$$P\left[s_i^{(\ell)} \mid s_i^{(k)}\right] = \frac{P\left[s_i^{(\ell)}, s_i^{(k)}\right]}{P\left[s_i^{(k)}\right]}$$

$$= 2P\left[s_i^{(\ell)}, s_i^{(k)}\right],$$

where we have used the fact that $P\left[s_i^{(k)} = -1\right] = P\left[s_i^{(k)} = +1\right] = 1/2$, since the BPSK symbols are supposed to be *a priori* equi-probable. Finally, Equation (2) can be approximated as

$$P\left[s_i^{(\ell)}\right] = \frac{1}{M-1}\left\{ \sum_{s_i^{(1)}=\pm 1} P\left[s_i^{(\ell)}, s_i^{(1)}\right] \right.$$

$$+\ldots+ \sum_{s_i^{(\ell-1)}=\pm 1} P\left[s_i^{(\ell)}, s_i^{(\ell-1)}\right]$$

$$\left. + \sum_{s_i^{(\ell+1)}=\pm 1} P\left[s_i^{(\ell)}, s_i^{(\ell+1)}\right] + \cdots + \sum_{s_i^{(n)}=\pm 1} P\left[s_i^{(\ell)}, s_i^{(n)}\right] \right\}$$

$$\simeq \frac{2}{M-1} \sum_{\substack{k=1 \\ k\neq \ell}}^{M} \sum_{s_i^{(k)}=\pm 1} \underbrace{\hat{P}\left[s_i^{(k)}\right]}_{[\text{from decoder } k]} \underbrace{P\left[s_i^{(k)}, s_i^{(\ell)}\right]}_{[\text{a priori source correl.}]}$$

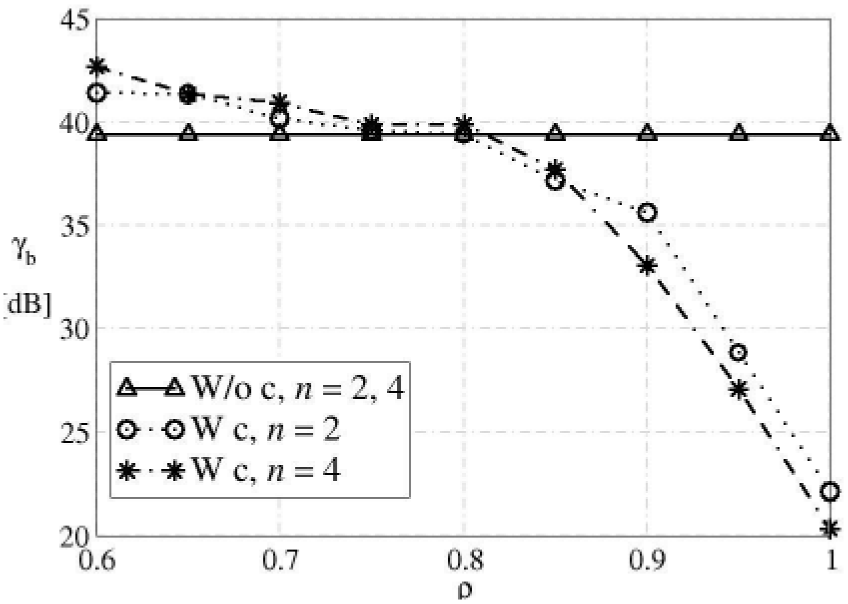

Figure 7. The SNR as a function of the correlation coefficient, ρ, required to achieve a BER equal to 10^{-4} in a low-density parity-check coded scenario with block-faded links. The number of sources, n, is set to either two or four. For comparison, the SNR required when the source correlation is not exploited at the access point (without c) is also shown.

where $P\left[s_i^{(k)}, s_i^{(\ell)}\right]$ can be obtained by marginalization of the nth dimensional a priori joint PMF $\left\{P\left[s_i^{(1)}, s_i^{(2)}, \ldots, s_i^{(M)}\right]\right\}$ of the information sequences at the input of the sources[4]. The intuition behind Equation (4) consists in modifying the input a priori probability of a single bit by taking into account – through a weighed average – the reliability values (on the same bit) generated by the other decoders. In particular, the weight of the reliability value generated by the kth decoder is given by the joint a priori probability between the ℓth and the kth decoders.

We now consider a specific scenario where each of the source sequences is encoded using a regular (3,6) low-density parity-check code with rate 1/2 and $N = 2000$. Each component decoder performs a maximum number $n_{\text{it}}^{\text{int-max}}$ of internal iterations, set to 50, whereas the number $n_{\text{it}}^{\text{ext}}$ of external iterations between the two decoders is set to 20. The low-density parity-check code is constructed in a *random* fashion, according to the following algorithm, which exploits an idea similar to the progressive-edge-growth (PEG) algorithm presented in [29]. Some potential connections, denoted as *sockets*, are drawn for all the variable and check nodes. For each variable node, a socket is then randomly chosen among all the free sockets at the check nodes, and the connection is added only if a *cycle* of a given (or lower) length is not created. In our case, the checked cycle length is equal to six.

In Figure 7, the SNR required to achieve a bit error rate (BER) equal to 10^{-4} is shown, as a function of the *correlation coefficient*, ρ, in various low-density parity-check-coded scenarios. In the relayed cases, the source-relay links are ideal. The number of sources, n, is either two or four. For comparison, the SNR required when the source correlation is not exploited is also shown. While for values of ρ lower than 0.8, exploiting the correlation leads to a (limited) SNR loss; for higher values of ρ, the SNR gain becomes significant. At very low values of ρ, the SNR loss is due to the fact that the iterative decoding scheme at the access point does not converge. In other words, the suggestions that the decoders pass to each other are not reliable and degrade the performance. However, the number of sources seems to have a limited impact on the SNR gain.

5. Concluding Remarks

In this paper, we have presented a perspective on cooperative coding for distributed radio systems. In particular, in a scenario with correlated sources, four possible approaches have been analyzed to exploit the source correlation: a non-cooperative system, cooperative source coding, cooperative source-channel coding, and joint source-channel coding. The differences between these approaches reside on how source correlation is exploited: either at the sources and/or at the access point. Obviously, the system complexity concentrates at the system position where correlation is exploited. In all approaches considered, we have derived the probability of error, i.e., the probability of incorrect delivery of the sources' information signals, and a comparative analysis has been carried out. In the joint source-channel coding case, a few idealistic assumptions were considered: ideal power control and ideal channel coding. However, it has been observed that joint source-channel coding systems are very attractive in wireless scenarios where the communication channels between the sources may be very noisy. As an illustrative example of the feasibility of these schemes, we have considered a low-density parity-check-coded joint source-channel coding system in a scenario with block-faded direct links and an arbitrary number of sources.

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[1] In [16], the communication links are affected by additive white Gaussian noise, whereas in this work we extend this approach to encompass the presence of fading.

[2] Note that the ultimate feasibility region, obtained in the presence of faded channels with ideal power control, coincides with that in the presence of additive white Gaussian noise links.

[3] Note that only the information bits are considered in the exchange of reliability information between the component low-density parity-check decoders, since the coded bits are not directly correlated.

[4] The joint PMF of $\left\{ s_i^{(k)} \right\}_{k=1}^{M}$ can then be obtained directly from the correlation between the information sequences. Note that Equation (4) is an approximation, since heuristically the first probability in the summation on the right-hand side is obtained from the reliability values generated by the other decoder, whereas the second probability is a priori.

Concepts and Main Applications of High-Altitude-Platform Radio Relays

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Abstract

Novel radio systems using high-altitude platforms (HAPs) at altitudes around 20 km above the ground are proposed as a way of overcoming the drawbacks of conventional satellite and terrestrial radio systems. This paper provides an introduction to the field, acting as a precursor to several extended papers on HAPs that will be featured in a future *Radio Science Bulletin* special issue. The concept, genesis, development, and recent status are detailed, and the proposed applications and advantages of different categories of HAPs are discussed. The possibilities of microwave-energy transmission from ground stations to receiving rectifier antennas (*rectennas*) on HAPs are described as a way of overcoming the long periods of darkness, when solar cells are not usable. Finally, the paper describes the new HAPS projects being realized, followed by a short forecast of future trends in the area. A thorough comparison of radio communication and remote sensing using GEO and LEO satellites, terrestrial radio systems, and HAPS is presented in tabular form.

1. Introduction

Radio communication systems are one of the main promoters of the modern economy and social growth. Their importance is also predominant for defense and security issues. Nowadays, the global numbers of mobile phones and of mobile PC subscribers alone have exceeded two billion and four hundred million, respectively. Their numbers are still increasing, as shown in Figure 1 [1, 2]. Most mobile radio communication equipment and systems are terrestrial. However, for long distances, reliable communication, and for scarcely populated rural regions and oceans, geostationary (GEO) and low-Earth-orbit (LEO) satellites are used as base-station relays in the sky [3, 4].

GEO satellites are located 36,500 km above the ground in an equatorial orbit, and are geostationary related to the Earth. Their main applications are for good-quality long-range mobile radio communication, and for broadcasting

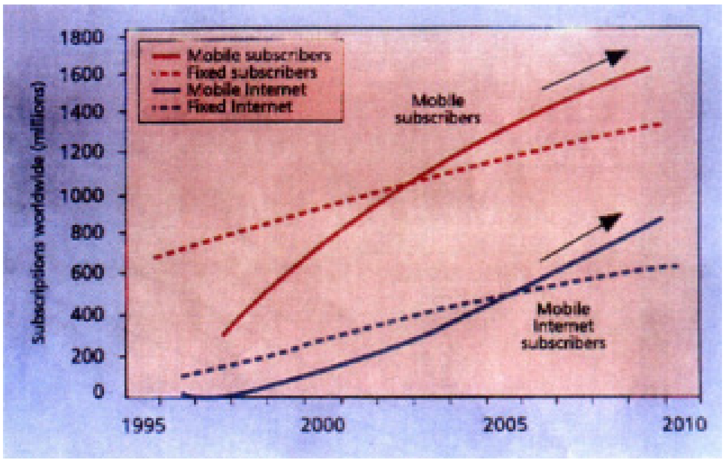

Figure 1. The global growth of mobile and fixed radio-communication subscribers [1].

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This is an invited paper.

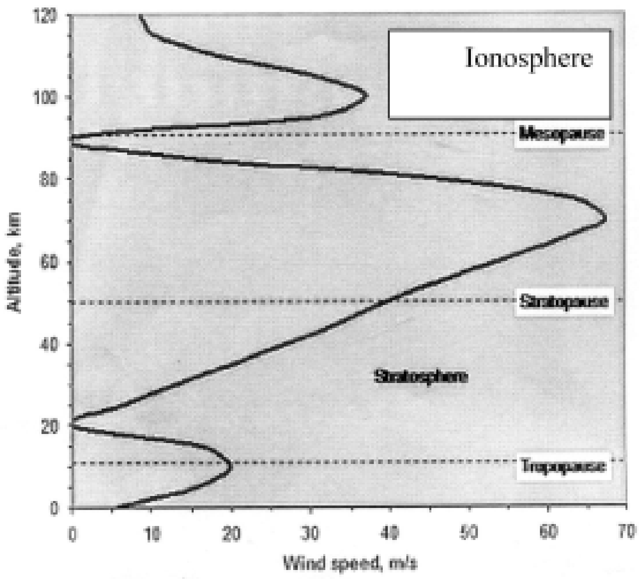

Figure 2. The annual average wind speed as a function of altitude: values vary with season and locations [8] (http://nssdc.gsfc.nasa.gov/space/model/atmos/cospar1.html).

[5, 6]. Nowadays, more than 400 GEO satellites are active, despite their high price and high launching and insurance costs. GEO satellites are characterized by high dispersion losses in the range of 200 dB, and time delays in the range of 0.3 s, due to their long distance from the ground stations [7]. The long delay causes synchronization problems, and limits high-speed Internet transmission.

Due to the tremendous growth in radio communications and the danger of the GEO satellite orbit saturation, LEO satellites were designed. These are located around 1000 km above ground. The launch of a single LEO satellite is significantly simpler and cheaper than for a GEO satellite, and the dispersion losses and delay times are smaller. However, the LEO satellite radio systems require handover, and the whole system operation needs 12 to 66 satellites [3]. The big LEO Iridium and Globalstar systems were technological successes but economic failures. The main reasons were the high cost of the LEO system, and the rapid development and improvement of terrestrial mobile-radio communication systems. Nowadays, these LEO systems are still operational, due to the important requirements for secure mobile-radio communication.

Today third-generation terrestrial mobile communication systems are expanding quickly. However, they have several limitations, mainly linked to designs intended as a way of overcoming the main limitations due to the lack of line-of-sight (LOS) propagation conditions. In most cases, even with high base-station antennas, mobiles are shadowed and affected by multipath and fading, which limit the quality and the operational range of terrestrial radio-communication systems [7, 8]. Future terrestrial mobile-radio design improvements will require a significant increase in the number of base stations. This will negatively impact the environment in dense urban zones, and more in scarcely rural zones, where the level of transmitted radiation is usually higher. The operational range is limited. Even for a high-altitude terrestrial base station, it is in the order of only 30 km [8, 9].

A new technique for efficient mobile-radio communication systems and other applications is the development of high-altitude platforms (HAPs), located in the stratosphere at an altitude of the order of 20 km for relay radio stations [3]. These HAPs stations will be stabilized relative to the ground as stratospheric quasi-stationary platforms (SQ-SP), due to significant progress in the technology [3, 10]. The performance and advantages of HAPs are, in several aspects, better than for GEO and LEO satellites and conventional terrestrial stations, as will be explained in this paper. This paper is intended as an introduction to the challenge of HAPS. It will be followed by a special *Radio Science Bulletin* issue, which will include several extended papers on the subject.

2. HAPs Concepts

HAPs are high-flying aircraft or airships, which will operate from altitudes of 17 to 24 km. The main reason for this operating altitude is that numerous wind-speed measurements show that the slowest wind speeds occur at these heights, extending above the tropospheric jet-stream wind altitudes, as shown in Figure 2 [11]. Less power is therefore required for the platform's stabilization. A second minimum wind zone occurs at around 90 km altitude, but this is less advantageous for HAP operation [3, 11].

Numerous temperature measurements show that the temperature decreases with altitude, and reaches a local minimum near the tropopause layer, around an altitude of 18 km. The average temperature is quite constant, around −56° C, at this lower stratospheric layer where HAPs will

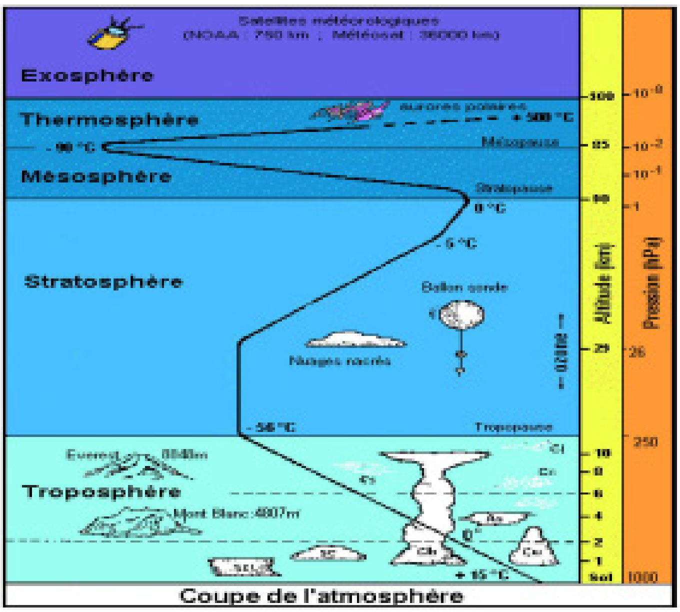

Figure 3. The average temperature as a function of altitude [11] (NASA/COSPAR).

operate, as shown in Figure 3. Therefore, the noise temperature and insertion losses of the antennas and the front-end circuits of the receivers (RX) for HAPS are reduced to a minimum [7, 10], and the receiver sensitivity is maximized at altitudes of 17 to 24 km.

The radiation effects – especially due to the sun's activities – are significantly reduced at an altitude around 21 km. This is due to the long-distance separation from the intense radiation of the Van Allen belts at altitudes ranging from 1500 to 5000 km, and from 13000 to 20000 km, which affect satellites [7, 9]. The extensive shielding seen on satellites for enhancing operational lifetimes is therefore not necessary for HAPs.

However, the ultraviolet (UV) radiation and ozone concentrations are higher at 21 km than at lower altitudes. It is therefore recommended that light composite materials, made of carbon fibers, be used in the construction of the HAPs. The exposed electronic devices must also be hardened against UV radiation [5, 11].

A general concept of the HAP is shown in Figure 4 [3]. The operation of the HAP requires electrical energy for the motors used to stabilize the platform in a quasi-stationary position related to the ground. Electrical energy is also required for the payload, which especially includes the transmitters (TX), receivers, power supply, and sensors to relay information to ground [3].

For GEO satellites, the maximum sun eclipse time is 1.2 hours and occurs only twice each year. LEO satellites suffer only from very short eclipses. In contrast, HAPs will suffer from very long eclipses. The HAPs Earth shadow results in eclipses near to 12 hours per day for locations in proximity of the equator, and up to 24 hours per day in the winter for locations in proximity of the Earth's poles [7, 10].

HAPs therefore need significantly larger and heavier energy storage systems than do satellites. The electrical power required by HAPs is in the range of 10 to 150 KW [8, 11]. This power can be obtained by efficient solar cells [12]. During eclipse periods, batteries, such as NiH2 or LiIon can be use. Another alternative is fuel cells, which are very heavy and are still not sufficient for long eclipse periods [8]. A possible solution may be the transmission of microwave (MW) energy from the ground. This would be converted at the platform to direct current using rectifier antennas (rectennas), characterized by very high efficiency in energy conversion [13, 14]. The rectenna function is shown in Figure 4, and will be explained briefly in Section 3.

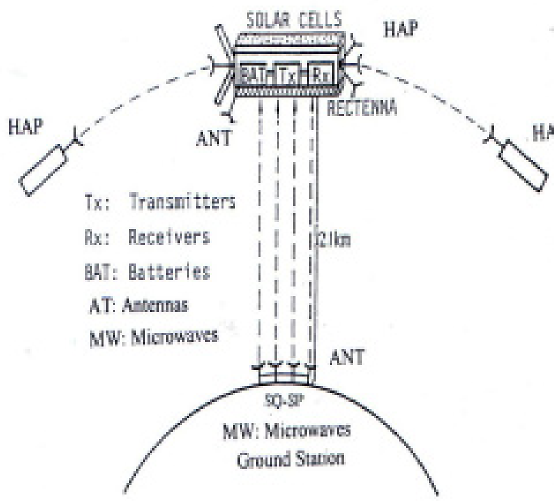

Figure 4. The concept of the HAPs [3].

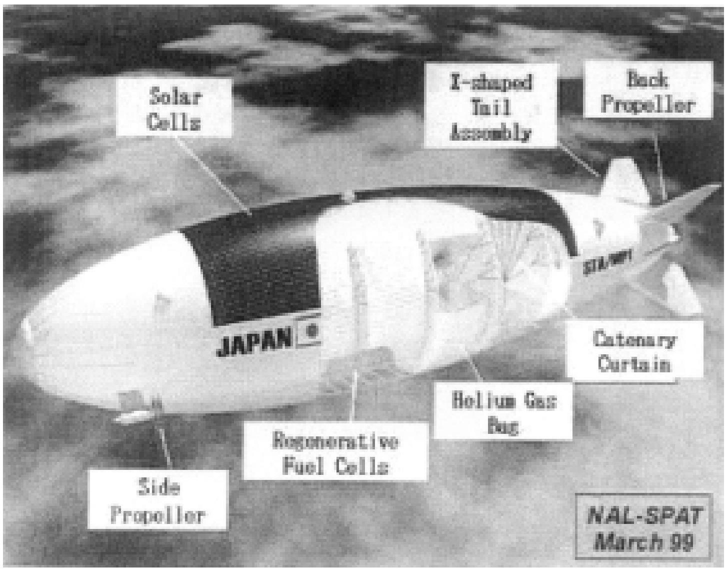

Figure 5. Japan's future HAP airship for a mission duration of three years. It is intended to operate at an altitude of 20 km, it has a length of 245 m, and a diameter of 60 m [26] (ASTAP02/FR06/EG.HAPS.04; http://www.aptsec.org/astap).

From geometric principles, the maximum HAP radio-operation radius under line-of-sight conditions for a zero elevation angle and an altitude of about 20 km, is around 600 km. To cover larger areas as do satellites, information can therefore by relayed between HAPs, as shown in Figure 4 [3]. Inter-HAP communication can use high-capacity, millimeter-wave or infrared-beam carriers, which are not affected by moisture, clouds, and precipitation losses in the stratosphere [7, 9].

An important advantage of HAPs over terrestrial radio systems is the mostly favorable line-of-sight propagation conditions, similar to satellites. However, the dispersion losses of HAPs – which are proportional to the square of the separation distances – and the delay times are significantly lower than for satellites [9]. HAPs can therefore provide much higher rates of information and faster Internet transmission using significantly lower transmitter power, and smaller and less-costly terrestrial mobile handsets, than for satellites. These handsets can operate even inside buildings without external antennas, which cannot be achieved for satellites [7, 8]. The full comparison between GEO satellites, HAPs, LEO satellites, and terrestrial radio systems is given in Table 1.

At the HAPs altitude, the atmospheric pressure is low, but the air density is still sufficient for activating aircraft propellers operated by electrical motors from solar energy or fuel. The relatively smooth-flowing air stream, combined with state-of-the-art propulsion, aerodynamic, thermodynamic, and material designs, will provide a stable and controlled flight. This will result in accurate position maintenance and minimal axial (pitch, roll, yaw) rotation. The rate of change of the velocity of stratospheric winds is well within the capability of the propulsion and control systems of HAPs for maintaining the desired position and heading [11].

There are two categories of HAPs:

Lighter-than-air (LTA) HAPS are usually balloon aerostats, or airships filled with helium gas, as shown in Figure 5. The lighter-than-air HAPS need less energy for launching and stabilization over a fixed spot [7, 10].

Heavier-than-air (HTA) HAPS are manned airplanes or UAVs (unmanned aerial vehicles). According to Bernoulli, a lift force keeps the heavier-than-air airplane in the air. An adequate forward thrust is provided by propellers activated by electric motors, jet engines, or other thrusters. The heavier-than-air vehicle has to move in order to float. Most HAPs will fly in a circle of up to 2 km radius, controlled from the ground [11, 14].

The International Telecommunication Union (ITU) has recognized the HAP as a separate category of radio station, the High Altitude Platform Station (HAPS), and has assigned the 2, 31/28, and 47/48 GHz frequency ranges for its operation [8].

The majority of the experimental and operational commercial HAPs are unmanned, solar-powered UAVs, e.g., those built by NASA, as shown in Figure 6 [10]. The Russian M55 GN stratospheric single-seat aircraft, which was designed and built as a piloted, manned HAP, is an alternative. To maintain continuous operation, pilots may operate in eight-hour shifts, with replacement aircraft taking over to provide continuous operations over 24 hours [10, 15].

Table 1. A comparisons of GEO and LEO satellites, HAPS, and terrestrial radio communications systems

Parameters	GEO Satellites	HAPS	LEO Satellites	Terrestrial Base Stations
1. Launch and insurance costs	High (–) $40-$80M	Very low(+) <$0.3M	High (–) $100-$200M for the whole LEO system	Very low(++) No launch required
2. Satellites required for system operation	1	1	12-66	1
3. Average system construction cost	Very high (–) $150M per satellite	Low (+) $5M; less for mass construction	Low (–) Satellite $10M; high for system	Very low (++) especially for the power supply
4. Dispersion losses at L band (1.5 GHz) Ka band (30 GHz)	Very high (– –) 185-190 dB 210-215 dB	Low (++) 130-145 dB 155-170 dB	High (–) 160-170 dB 185-195 dB	Very low (++) In many cases, no LOS conditions
5. Transmitter power requirements	High (–)	Low (++)	Medium (+)	Low (+) depends on distances and LOS
6. Receiver sensitivity requirements	High (– –)	Very low (++)	Medium (–)	Medium (–)
7. Indoor coverage	No available (– –)	Available (+) most cases	No available (–) generally	Available (+) most cases
8. Handover requirements	No (++)	Seldom (+)	Yes (–)	Yes (–)
9. Hop delay time	High (– –) ~ 250 msec	Very small (+) generally < 1 msec	Medium (–) ~ 10 msec	Very small (+) neglected in urban and LOS scenarios
10. Energy requirements	Solar cells (+) and expensive Batteries	Solar cells (– –) expensive and heavy fuel cells or rectenna	Solar cells (+) and expensive batteries	Terrestrial (++) electricity supply and simple batteries
11. Coverage of one satellite	Up to 40% (++) of Earth's surface or 8,000 km radius	Up to 500 km (–) radius	Up to 2500 km (+) radius	Local (– –) Up to 30 km radius; small coverage in urban areas
12. Transmission rate per user*	1-155 Mbit/s (–)	(++) 10-1000 Mbit/s	0.1-10 Mbit/s (– –) Mbit/s	0.1-55 Mbit/s (+)
13. Full coverage	Global (++) 3 satellites can cover most of Earth's surface	Local (–) or regional several inter-connected platforms	Global (+) 12-66 satellites to cover the Earth's surface	Local (– –) cannot cover Earth's surface
14. Technology	Proven (++) available on the shelf	Not yet (–) but high probability of achievement	Under (+) advanced development	Proven (++) available off the shelf
15. Vulnerability to destruction	Very low, (++) for the space segments	High (–)	Low for (+) the whole space segment	Very high (– –)

* Data rates depend upon application and hardware complexity.

For GEO satellites, $h = 36000$ km, $d = 38000$ km; for HAPS, $h = 21$ km, $d = 70$ km;

for LEO satellites, $h = 700$ km, $d = 1000$ km; for terrestrial base stations, $2 < h < 30$ m, $0.1 < d < 20$ km.

3. Genesis and Development of HAPS

3.1 The Preliminary Stage: Up to 1990

The first steps for the realization of HAPs were carried out in the USA, with the development of high-altitude airplanes, manned and unmanned, such as the U-2 plane, shot down by the Soviet Union in 1960. This was followed by the SR71 Blackbird plane, which flew at an altitude above 25 km in 1976 [15]. In the beginning of the 1960s, a few high-altitude echo balloons were also launched to reflect broadcasting for Bell Labs, but the results were disappointing [5].

The possibility of HAP realization was studied by NASA, first for military and meteorological purposes, and later for radio-communication and remote-sensing applications. The US also developed lighter-than-air tethered balloons for military and homeland security supervision, used at altitudes up to 3 km, for operational times of a few days [8, 16].

The long-term operation of unmanned UAV HAPs may require a microwave energy source and large antennas from a ground station below the platform. It may also require a rectenna on the HAP to convert the *microwave energy to direct current (dc) for charging the batteries, and activating the payload and propellers to stabilize the platform, especially in case of stratospheric winds. This is in addition to efficient solar photovoltaic cells, exposed on the top of the platforms, as shown in Figure 4 [7, 14].

The pioneers in the development of rectennas were R. H. George and W. C Brown from Raytheon. They developed very power-efficient microwave rectennas, with efficiencies exceeding 85% at the ISM frequency of 2.45 GHz [13, 14]. Rectennas will also be used for future solar power satellites (SPS) [17].

At the beginning of the 1990s, NASA also invested in the development of several preliminary HAP projects, especially for military remote sensing, intelligence, metrology, and radio communication [15, 16]. The US department of defense and NASA developed a manned heavier-than-air HAP with a crew of two pilots. The first gasoline model was the Rutan Voyager, which made a flight of 216 hours without interruption in 1986 [11, 16].

The first commercially oriented project for radio-communication HAPs, the Stratospheric High Altitude Relay Platform (SHARP), was initiated by the Communication Research Center (CRC) group in Ottawa, Canada, in collaboration with the University of Toronto, and several local firms. They succeeded in the operation of a one-eighth-scale model unmanned heavier-than-air HAP in 1987. It was a radio-controlled aircraft with a 4.5 m wingspan, powered from a ground-based microwave transmitter operating at 2.45 GHz. The carrier beam energized a disc-shaped rectenna located on the aircraft, providing 25 kW of dc power to a 100 kg payload, and to motor-driven propellers [8, 14]. However, no final 21-km altitude HAP was developed, because of funding problems [3, 15].

3.2 The Development Stage: From 1990 to Present

The successor of the U-2 and the Blackbird, the U-2R "Dragon Lady," is still operational today. Designed for high-altitude intelligence and reconnaissance missions, this aircraft can fly above 20 km altitude. It can carry some of the most advanced long-range reconnaissance gear available today. Its mission payloads include the highest-resolution synthetic-aperture radar (SAR), as well as sophisticated signals intelligence (SIGINT) systems. The U-2R provides near-real-time imagery to war fighters and national authorities. The "Dragon Lady" was used extensively during operation Iraqi Freedom in 2003, and provided important damage-assessment information after Hurricane Katrina struck the Gulf of Mexico in 2005 [15].

Figure 6. The Helios NASA experimental heavier-than-air solar-powered HAP flight over Hawaii, at an altitude of around 21 km in 2001 [15].

The later US manned canard aircraft HAP model was the Proteus. It was capable of lifting a 1000 kg payload to an altitude of 20 km, and could circulate there for many hours [7, 18]. This manned High Altitude Long Operation (HALO) option, shown in Figure 7, was proposed by Angel Technologies for broadband communication in 1998. With this option, full-time operation of 24 hours, seven days per week may be possible, using a time shift of three Proteus aircraft [16, 18]. Currently, the Proteus is owned by Northrop Grumman, and can also be operated as an UAV.

Several UAV heavier-than-air HAPS models were have been developed in the US:

- The NASA Pathfinder project Helios solar-powered UAV model HAP, shown in Figure 6 [8, 11], used motors and propellers fed by solar cells located on large transparent wings of over 35 m span. It reached an altitude of 30 km in 2001, and was expected to stay in operation without landing for more than six months. However, the Helios model crashed over Hawaii in 2003, and the project was discontinued [15].

- Smaller versions, but not less effective than the Helios and especially for radio-communication applications, are the Sky Tower HAPs, The Sky Tower development was a joint project between NASA and the Japan Ministry of Communications with the cooperation of the Aerovironment Corp. [19].

- The NASA/Northrop Grumman project Global Hawk UAV uses gasoline engines, operates at an altitude of 20 km, and remains on station for more than 24 hours [10, 13]. The Global Hawk UAV supports the USAF, NATO members, and Australia in the global war against terrorism; monitors wild fires and hurricanes; and other applications. These HALE (high-altitude, long-endurance) UAVs, combined with satellite and line-of-sight communication links to ground forces, permit worldwide operation. The Global Hawks have flown more than 24,000 hours on operational missions. Improved versions – the RQ-4B and RQ-4N – are under contract. They can carry payloads of nearly 1000 kg payload [20].

- Lockheed-Martin, Boeing, Raytheon, and General Atomics Aeronautical Systems Inc. are companies that are also involved in HAP design and development, especially for military applications [11, 15, 21].

In 2006, the QinetiQ company in the UK built the Zephyr, an ultra-light high-altitude long-endurance UAV for the US and UK ministries of defense. The Zephyr is a carbon-fiber UAV, launched by hand, as shown in Figure 8 [22]. This low-cost, light UAV recently achieved an 82 hour flight at an altitude up to 20 km. It is useful for reconnaissance and relaying radio communications for homeland security, and in the worldwide fight against terrorists [15].

The first enterprise to make commercial use of near-space HAPs using lighter-than-air balloons for radio communication was Space Data, in Chandler, Arizona [23, 24]. The Space Data balloons are extremely low cost, and can be launched within minutes by just one person from almost anywhere. The balloon has a diameter of about 8 m, a total weight of about 6 kg, and a payload weight of about 3 kg. The radius of operation is in the range of 8 to 320 km. The balloons operate at altitudes from 20 to 30 km. The 2008 Space Data commercial operating US network comprises 11 launch sites, three remote ground stations, a network operation center, and numerous balloons. It has logged more than 250,000 flight hours over the US. Space Data has also developed a worldwide military combat version of their lighter-than-air HAP system, called Skysat [24].

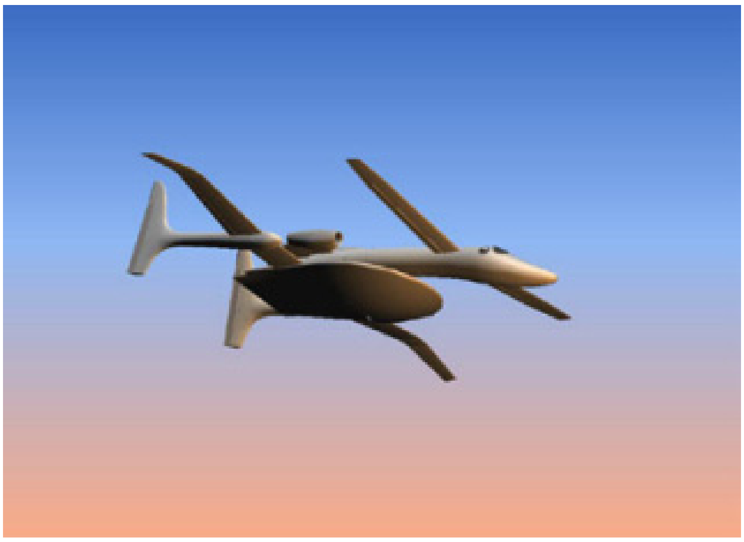

Figure 7. The Proteus Manned Canard aircraft HAP, owned by Northrop-Grumman [18]. The Proteus program is supported by the NASA Dryden Flight Research Center.

Figure 8. The launch of the Qinetiq ultra-light low-cost Zephyr high-altitude long-endurance UAV [22] (copyright QinetiQ).

The Sanswire Corp. specializes in the design and development of big lighter-than-air HAPs airships, which they call Stratellites. A prototype was tested in 2005, with a length of 75 m, a width of 44 m, and a height of 26.5 m. The Stratellite is powered by electric motors, and energized by solar energy and batteries. The expected payload weight is up to 1400 kg. It holds its position using six onboard GPS units for a duration of up to 18 months [15, 25]. In 2005, Sanswire joined the German TAO company, and they operate from the USA [25].

The Japanese have cooperated with Canada and the US in the design and development of HAPs [16]. Japan invested in lighter-than-air airship HAPs, and developed the airship shown in Figure 5 [26, 27].

Russia is also involved in lighter-than-air and heavier-than-air HAP developments. The Russian ROS Aero Systems Corp. developed several models of lighter-than-air airship HAPs named Berkut. These can carry up to 1200 kg payload, especially for communication and surveillance applications. These solar-powered HAPs are expected to supply up to 15 kW for payloads and 100 kW for position stability, with a flight endurance of up to four months [28]. The Russians also developed the M55 GN stratospheric manned single-seat aircraft in 1993. In 2007, QUCOMHAPS, an Irish company, intended to order 50 M55 GN aircraft to supply Malaysia, Indonesia, Saudi Arabia, and the AER with 24 hour by seven day full-time coverage for surveillance and radio-communication applications [29].

The European Union has recently invested in the study and realization of HAP systems. The European HELINET project, which ran from 2000 to 2003, aimed at developing a fast Internet network served by HAPs [10]. This was followed by CAPANINA, which ran from 2003 until January 2007. This tested the feasibility of delivering broadband communications from different types of HAPs. Practical trials were held in the UK (tethered balloon), Sweden (free-floating balloon), and the USA (unmanned UAV). The project also developed detailed design and analysis for equipment, especially antennas, and system architectures. This work is continuing through the COST 297 project, which is a research discussion forum for "High Altitude Platforms for Communications and Other Services" (HAPCOS) [30]. The membership mainly comprises European academic organizations and companies, and is led by the University of York. The activities are supported by three working groups: Radio Communications, Optical Communications, and Aeronautics and Other Applications [10]. Several HAP prototype models are in the realization process [30, 31].

Several countries such as South Korea, Israel, and China are also involved in HAP development, especially for military applications [32-34].

3.3 The Realization and Commercialization Stages: From 2010 to the Future

The realization and commercialization of HAPs seem to be very near. Several organizations and countries are planning to use HAPs for commercial purposes, especially for military radio-communication and remote-sensing purposes.

QinetiQ's Zephyr project in the UK, discussed earlier, is expected to continue through to the commercialization stage. It will be used for civilian Earth observation, small-scale communications relay, and military missions [22].

The StratXX organization in Switzerland is aiming to deliver a range of lighter-than-air HAPs and lower-altitude aerial platforms. The purpose is to deliver a combined remote-sensing and radio-communications system [36].

Figure 9 : An artist's impression of the Lockheed-Martin future high-altitude long-endurance HAP using a thin-film solar-cell array [21]

A small Italian company, ERS-SRL, is planning to deliver communications and environmental monitoring services, using the German GROB future high-altitude, long-endurance manned or UAV aircraft [37]. The QUCOMHAP project for communication relays and reconnaissance in Asian countries, using Russian manned M-55 aircraft HAPs, is still at the paper stage [29].

The Sanswire company claims to be able to commercialize its products by 2010, and to become a leader in civilian and military HAPS applications [25] The US Space Data and UK QinetiQ companies have a strong potential for becoming leaders in commercial HAPs operational systems [23, 24].

The Japanese are also involved with the NASA in the development of a long-term space solar-power (SSPs) program. This will require huge rectennas [13, 14], microwave power transmission up to 5 GW from geostationary satellites to the ground, and huge solar-cells arrays. Japan also initiated the Skynet project [26]. This intends to develop a network of from five up to 15 HAPS, to cover almost the total Japanese territory, with radio communication, fast Internet, and broadcasting [26, 27]. However, no activities have been recently published on the Skynet project.

Most of the development and investment in HAPs systems remains in the military and homeland-security domains. The US Department of Defense Rapid Eye program is an exploratory development program. It has the overall goal of developing and demonstrating the ability to deliver persistent remote-sensing and radio-relay capabilities anywhere on the globe within one hour, and to remain on-station until relieved or until the mission is completed [38, 39].

Boeing has recently commenced studies of solar-cell power generation, produced by its subsidiary, Spectrolab. By 2010, the company plans to offer space solar cells with efficiencies as high as 33%. These cells are five times more efficient than the solar arrays used to propel the Helios UAV [11]. This could provide primary power sustaining an airborne platform for nearly unlimited times. It could contribute to the future development of commercial and military HAP systems [39]. Boeing and Aurora Flight Sciences, in the joint Orion program, are developing a heavier-than-air UAV aircraft powered by hydrogen fuel. It may operate up to 100 hours at an altitude of around 21 km [15, 38].

Lockheed Martin is continuing to make progress with lighter-than-air HAP technology. This is being funded by the US Army, for the delivery of remote sensing and communications for tactical scenarios. Their newly designed HAP, shown in Figure 9, will have a very long endurance using thin-film solar cells [40, 41].

Aurora Flight Sciences plans to build the Odysseus unmanned HAP based on a radical idea. Instead of one single plane, three separate planes will take off individually.

Figure 10. The Aurora Flight Sciences Corp. Odysseus future high-altitude long-endurance HAP concept. This is based on launching three separate planes and joining them together in the stratosphere [40].

Once aloft, they will join together at the wing tips to create one giant aircraft design of 150 m wingspan. Odysseus will allegedly be powered by a large array of solar panels, positioned on the aircraft's wings. These will have the ability to pivot, so that the sunlight collection area is maximized. The Odysseus UAV, shown in Figure 10, will have two flight configurations. In the "Z-mode," the wing is partially folded, so the solar cells are canted to catch as much sunlight as possible during the day. In the flat configuration, the wing span is maximized to reduce drag and power consumption during the night, as shown in Figure 10 [39, 40].

Recently, the US Defense Advanced Research Projects Agency initiated an ambitious Vulture UAV high-altitude, long-endurance project. The Vulture is intended to keep a 450 kg payload powered by a constant 5 kW supply from solar and fuel cells for up to five years of operation, at an altitude of 27 km. It is intended to operate like a pseudo-satellite. This UAV has to be kept precisely on station at least 95% of the time [41]. The three main competitors for this project are the Aurora Flight Sciences Corp.'s Odysseus, concept shown in Figure 10; a larger modification of the QinetiQ Zephyr UAV, shown in Figure 8, is being proposed by Boeing; and Lockheed Martin plans a modified version of their lighter-than-air UAV shown in Figure 9 [21, 42].

4. Important HAP Applications

HAPs will have numerous applications. The most important are the following:

1. National or limited regional coverage of commercial and military radio-communication systems. In particular, cellular, voice, Internet, and video could be provided for extended rural coverage up to 1000 km in diameter, in comparison with 20 to 40 km for a common terrestrial rural mobile or cellular system. In addition, very high speed Internet could be supplied for limited urban areas, operating using the $3G^{+}$ or 4G mobile-radio communications. This could also provide emergency communication in case of natural disasters and conflicts [8, 10]. Radio communication is thus the main HAPs application [7, 15].
2. Permanent military, intelligence, surveillance, and reconnaissance (ISR) missions, with very high resolution, in cooperation with national or international homeland-security missions [11, 39]. Examples include coastal or border-guard observation, and missile-launch detection [16, 40].
3. Extended local or national television broadcasting coverage, including high-definition TV (HDTV), voice, and digital audio, for operational distances up to 1000 km [10].
4. Local or regional radio-communication services, such as personal computers, data communications, distance learning, and financial transactions. These services can be extended to global coverage using multiple HAPs cooperating with global GEO or LEO satellite systems [3].
5. Persistent surveillance for illegal activities such as drug traffic, signal gathering, and spectrum monitoring, which are important for homeland security [15, 16].
6. Monitoring ground vehicles, aircraft, and ships for traffic control, security surveillance, and position information, including special missions such as detection and localization of stolen cars for the police or defense authorities [7, 15].
7. Remote sensing of regional weather conditions for forecasting and atmospheric data gathering, including hurricane detection and monitoring [3].
8. Automatic control and telemetry data collection for electrical utilities.
9. Surveillance of pollution for environmental-conservation purposes, including monitoring concentrations of carbon dioxide, ozone, and radiation levels, even at high altitudes [8, 16].
10. Space observations and radio-signal monitoring from space and from the ground [3, 8].
11. Providing significant enhancement in reliability and accuracy to the Global Positioning System (GPS) by adding Differential GPS sources to the HAPs payload [3].
12. Scientific exploration of the moon, Mars, and other planets [43].

The feasibility of HAPs systems and the fulfillment of the applications described, depend on significant research achievements, funding, and technological improvement. Improvements are needed in aeronautical science, control, electromagnetic compatibility (EMC), wireless power transmission, payload efficiency, and radio-frequency interference (RFI) reduction techniques [3, 8, 9].

5. Conclusions

A description of the use of HAPs for radio-system operations has been presented in this paper, with reference to GEO and LEO satellites and terrestrial mobile radio-communication systems. The wireless microwave-power-transmission concept, using rectifier antenna (rectenna) elements, has been presented. Significant efforts for the realization of HAP systems have been made, starting from the 1950s, beginning in the US and Canada [13, 14]. Actual progress on HAP system realization has been achieved, especially in the USA, Japan, Russia, and also in Europe [8, 40]. The important applications of HAPs have been described.

HAP systems can be used efficiently in multiple radio applications with dependent service areas of diameters from 60 to 1000 km, which represent the maximum limit for line-of-sight reception. However, these can form part of an enhanced global communication network, by coordinating the operation of multiple HAPS, GEO satellites, and terrestrial systems. The HAP systems are thus not competing with the operation of GEO or LEO satellites or terrestrial systems, but instead complement them. This is due to the several unique advantages of HAPs, specifically the local and regional services, as presented in this paper and in Table 1.

Today, numerous organizations, companies, and research and academic institutes are involved in the design, development, and construction of HAPs. Several prototypes have been developed, and have operated successfully for a limited time. The actual status of HAPs is similar to the status of the GEO satellites following the Early Bird launch, and before the establishment of the INTELSAT organization. Forecasters predict that in spite of the deep current economic crisis, in a few years, the early growing pains of HAP systems will be resolved. This will be followed by commercial success, and the achievement of numerous civilian and military HAPs applications, especially in radio communication, remote sensing, and homeland security [39, 40].

6. Acknowledgment

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Structure and Performance of the Future Galileo Civil Signals

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J-L Issler

Abstract

Galileo is the European initiative in the field of global satellite-based navigation systems. It will come as a complementary alternative to the current US GPS and Russian GLONASS systems, which are being modernized, and to the upcoming Chinese COMPASS and the Indian IRNSS-GINS. This article aims at presenting the Galileo's signal structure, and explaining the rationale behind it in terms of signal-processing performance. This will be done by comparing the designed Galileo signals with the main civil signal available today, the GPS C/A signal. Understanding the basic processing and inherent limitations of the latter will provide us with the necessary background to introduce the innovations present in the Galileo signals, and their impacts on Galileo's performance.

1. Introduction

Recent years have seen an increasing interest by the civilian community towards positioning and navigation applications through the development of numerous systems using the US Global Positioning System (GPS). Its economic and strategic role as a civilian and military asset has led to many international initiatives in this field. This can be observed through the modernization/development of several global navigation satellite systems (GNSSs) and regional navigation satellite systems (RNSSs). Current initiatives for global navigation satellite systems are the modernization of the US GPS and the Russian GLObal Navigation Satellite System (GLONASS), and the development of the European Galileo, the Chinese COMPASS, and the Global Indian Navigation System (GINS). The regional navigation satellite systems initiatives include, for instance, the Japanese Quasi-Zenith Satellite System (QZSS), and the Indian Regional Navigation Satellite System (IRNSS).

Global navigation satellite systems use a medium-Earth-orbit (MEO) constellation of about 25 to 30 satellites, with between six to 12 satellites visible to a user on the Earth's surface (considering an open area). Assuming that the satellites' positions are known, the receiver has to estimate the propagation time – or, equivalently, the range – between each satellite antenna's phase center and its own antenna's phase center, in order to retrieve its position using multi-lateration techniques. This propagation time is computed using the ranging signals transmitted by each satellite. Obviously, the choice of these signals is crucial to accurately estimating the satellite/receiver distance. The design of the GPS C/A signal, the only civil global navigation satellite system signal currently fully available, was realized decades ago, based on the technology and processing power available at that time. The current rising demand for positioning and navigation applications shows how relevant was that design. However, as the expectations from the users are constantly increasing, and the applications' requirements are more and more stringent, it has become obvious that a new generation of signals is required. This is the reason why a significant part of the innovations brought by the new/modernized systems concern the structure of the signals.

Since 1999, the Galileo signal design has been taken care of by the Galileo Signal Task Force of the European Commission. Although some of the details have not yet been finalized – in particular, concerning the navigation messages – the main lines of the Galileo signal structure are well known. It is already possible to assess the performance of the Galileo signals with respect to the legacy civil-signal GPS C/A. The goal of this article is (1) to review the typical processing of the GPS C/A, in order to introduce the critical parameters affecting signal tracking; (2) to present the innovations implemented by the Galileo's civil signals; and (3) to quantify their impact on the receiver performance. It is important to point out that the innovations presented in this article did not all originate from Galileo. Some of them will also be present, in the same or a different form, in other global navigation satellite systems.

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The content of this article is as follows. In the following section, the GPS C/A signal structure and its model will be reviewed. The third section will analyze the typical global navigation satellite system tracking processes, as well as their critical parameters. The fourth section will introduce the innovations that are used by Galileo, as well as their impact on signal tracking. Finally, the Galileo signals will be presented, emphasizing their performance with respect to the GPS C/A signal.

2. The GPS C/A Signal Structure

The GPS C/A signal is based on direct-sequence spread-spectrum (DS-SS) techniques using a code-division multiple-access (CDMA) scheme [1, 2]. It is the product of three elements [3-5]:

1. A useful signal, *d*, specific to each satellite of the GPS constellation and referred to as the navigation message, which is necessary for the user to compute its position. (For example, it contains a model for computing the position of the transmitting satellite, as well as the drift of its clock with respect to the GPS time reference [6].) It is a binary message, encoded as a non-return-to-zero (NRZ) rectangular waveform. Its rate is 50 bits per second (bps), meaning that one bit lasts for 20 ms.

2. A binary spreading sequence, *c*, also referred to as a pseudo-random noise (PRN) sequence. The spreading sequence is composed of 1023 chips. Each satellite uses its own spreading sequence, which is taken from a Gold family of length 1023 [7]. The pseudo-random noise sequence is encoded as a non-return-to-zero rectangular waveform, and has a chipping rate, f_c, equal to 1.023 MHz. It thus repeats every 1 ms, and there are 20 repetitions of the spreading sequence within one navigation data bit. The pseudo-random noise sequences are chosen to provide good auto- and cross-correlation properties, in order for the receiver to be able to process the signals from one satellite without being disturbed by the signals from the other satellites. Examples of these correlation functions are shown in Figure 1. As a first approximation (the pseudo-random noise sequence is assumed to be infinite), the normalized autocorrelation, R_{cc}, of any pseudo-random noise sequence can be modeled as

$$R_{cc}(x) = \begin{cases} 1-|x| & \text{if } |x| \leq 1\text{chip} \\ 0 & \text{elsewhere} \end{cases} . \quad (1)$$

3. A carrier, the frequency of which is the L1 frequency, f_{L_1}, equal to 1575.42 MHz.

Consequently, the overall GPS C/A signal uses a binary phase-shift keying (BPSK) modulation with a rectangular waveform. Because the frequency, $f_0 = 1.023\,\text{MHz}$, is a fundamental frequency in a global navigation satellite system, the GPS C/A modulation is often referred to as BPSK ($1 \cdot f_0$) or BPSK(1). The signal generated at the satellite's level can thus be modeled as

$$s_{L1}(t) = d(t)c(t)\cos\left(2\pi f_{L_1} t\right) \cdot \quad (2)$$

The signal is then amplified and filtered (in order to respect the allowed transmitted bandwidth), and transmitted using a directional antenna array pointing constantly towards the Earth [4, 8]. As shown in Figure 2, the propagation channel comprises the satellite antenna, the propagation through free space, the propagation through the atmosphere (the ionospheric and tropospheric layers), the reflection or diffraction at obstacles, intentional or unintentional interference, and the receiver's antenna. The signal received by the user's antenna is composed of all the GPS signals coming from the satellites in view, which have been modified by the propagation channel.

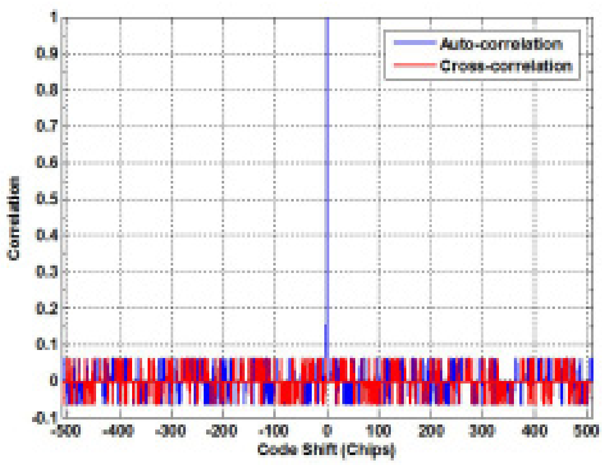

Figure 1a. The cross-Correlation between GPS C/A PRN 1 and PRN 2 and the autocorrelation of GPS C/A PRN 1.

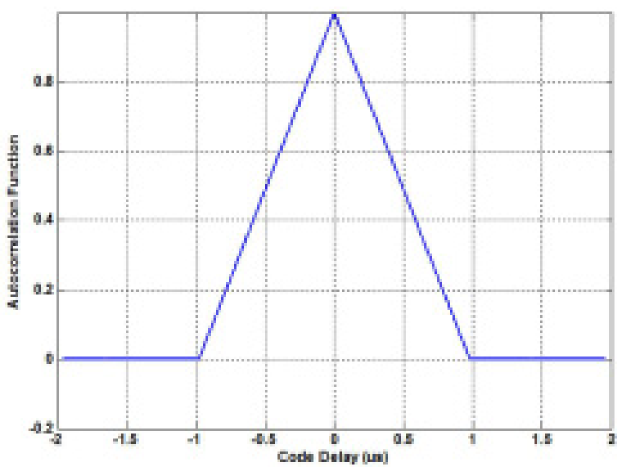

Figure 1b. A close-up of the autocorrelation of GPS C/A PRN 1.

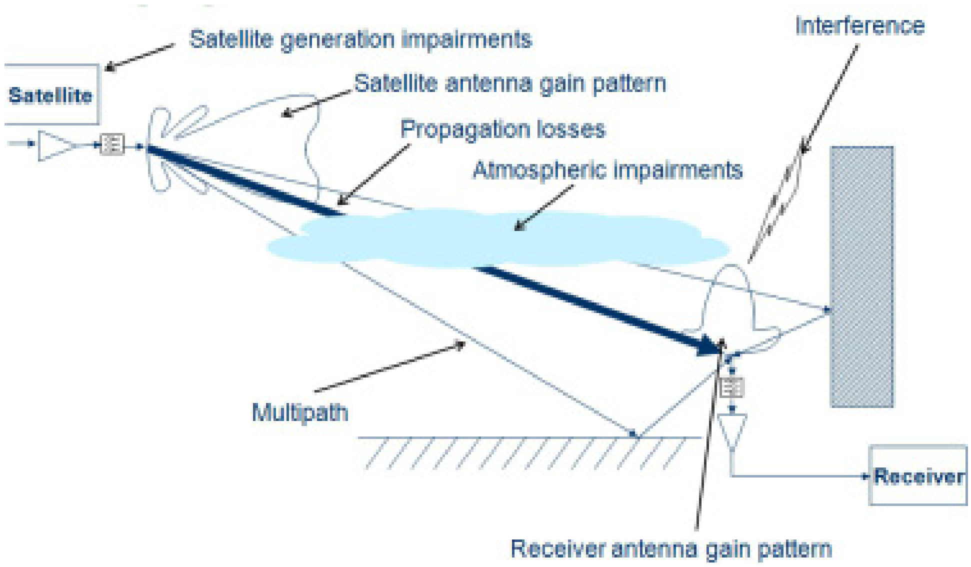

Figure 2. A typical global navigation satellite system propagation channel.

In order to compute a position, the receiver first needs to estimate the distance between receiver's antenna and each satellite's antenna. This is done by estimating the propagation time of the direct signal coming from each satellite. The receiver thus needs to allocate a specific processing channel to each potential satellite in view.

For each channel, the receiver processing sequence is: (1) acquire the signal, (2) track the line-of-sight (LOS) signal to estimate its propagation time, and (3) decode the navigation message in order to exploit it for positioning and navigation operations. In the frame of this paper, and for clarity, only one receiver channel will be investigated, and the focus will be put on the performance of the tracking operations.

Once the signal is received by the user's antenna, it goes through the receiver's front end, the basic functions of which are to amplify, filter, down-convert, quantize, and sample the signal [4, 8, 9]. Taking all the disturbances into account, the signal at the output of the front end can be modeled as (maintaining the continuous notation)

$$s_{L1}(t)$$

$$= \sqrt{2C} d\left[t-\tau(t)\right]\tilde{c}\left[t-\tau(t)\right]\cos\left[2\pi f_{IF} t+\varphi(t)\right]+n(t) \quad (3)$$

where C is the power of the incoming signal at the front end's output, τ is the delay due to the propagation channel at the front end's output, φ is the phase of the received signal at the front end's output (includes the Doppler effect), f_{IF} is the intermediate frequency used by the receiver, n represents the thermal noise at the front end's output, and $\tilde{c} = c \otimes h$ is the spreading sequence filtered by the equivalent front-end filter ($\otimes$ being the convolution operation).

The envelope of the GPS C/A signal power spectral density (PSD) can be obtained by computing the Fourier transform of Equation (1). It is shown in Figure 3, and is given by

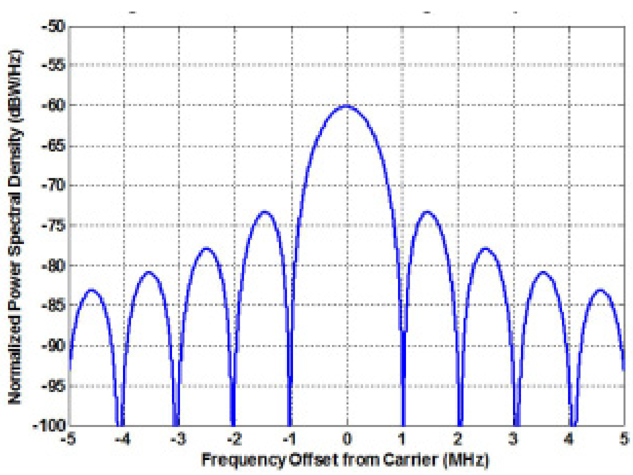

Figure 3. The power spectral density of the GPS C/A signal.

$$G_{C/A}(t) = T_c \, \text{sin}_c^2(\pi f T_c), \tag{4}$$

where $T_c = 1/f_c$ is the duration of a chip, and $\text{sin}_c(x) = \sin(x)/x$.

As will be seen later on, the choice of the receiver's front-end filter bandwidth impacts the signal-processing performance. It can be seen that the minimum possible front-end filter bandwidth is 2 MHz (double-sided), although a wider bandwidth could be chosen to also receive the sidelobes. The front-end filter bandwidth directly impacts the choice of the sampling frequency, and thus the receiver's power consumption. At this stage, it is important to remember that a typical GPS C/A receiver architecture will use around 12 channels in parallel, in order to track all the satellites in view. The choice of the sampling frequency will thus play an important role in the power consumption of the receiver, especially when considering handheld devices that need long autonomy.

3. Basics of Global Navigation Satellite System Signal Processing

3.1 The Correlator-Output Model

The receiver estimates the signal-propagation time using a correlation operation between the incoming signal and a local replica of the incoming signal. This local replica can be expressed as

$$r(t) = c\left[t - \hat{\tau}(t)\right] \exp\left\{j\left[2\pi f_{IF} t + \hat{\varphi}(t)\right]\right\}, \tag{5}$$

where the "hat" represents the code delay and phase generated by the receiver.

The correlation operation is done by multiplying the incoming signal with the local replica, and then integrating the resulting signal over a given duration, T_I, referred to as the coherent integration time. A thorough description of the correlation operation can be found in [3-5, 8]. It is important to understand that the correlation operation should be realized over an integer number of pseudo-random-noise sequences, in order to keep the auto- and cross-correlation properties of the spreading sequence. Moreover, it is also important to ensure that the integration is done over a segment of the received signal that does not include a data-bit transition, so that the correlator's output is not degraded. Consequently, the integration is generally limited to 20 ms for a GPS C/A receiver, once bit-synchronization has been achieved. Assuming no data-bit transition and an infinite front-end filter, the normalized output of the correlation operation during the *k*th correlation time interval can be modeled as [8]

$$K(k) = \sqrt{2T_I \frac{C}{N_0}} R_{cc}\left[\varepsilon_\tau(k)\right] \tag{6}$$

$$D(k) \, \text{sin}_c\left[\pi \varepsilon_f(k) T_I\right] \exp\left[j \varepsilon_\phi(k)\right] + n_K(k)$$

where $\varepsilon_\tau = \tau - \hat{\tau}$ is the code-delay error in chips, ε_ϕ is the carrier-phase error in the middle of the integration interval in radians, ε_f is the frequency error between the Doppler frequency of the incoming signal (assumed constant during the integration interval) and the local VCO frequency in hertz, *D* is the value of the navigation data bit during the correlation interval, N_0 is the level of the single-sided thermal noise power spectral density, and $n_K = n_I + jn_Q$, with n_I and n_Q being independent Gaussian noise with a unity variance.

There are several important observations to make regarding the correlator-output expression shown in Equation (6):

- It is clear that the correlator output is a very good indicator of the synchronization of the local replica with the incoming signal. Indeed, given a sufficient C/N_0, the correlator output will only be significantly above the noise level if ε_τ, ε_ϕ, and ε_f are small, so that R_{cc}, $\cos\left[\varepsilon_\phi(k)\right]$, and $\text{sin}_c\left[\pi\varepsilon_f(k)T_I\right]$ are close to one. Moreover, when all these conditions are fulfilled, not only is the signal synchronized, but it is also possible to decode the current navigation data bit from the sign of the real part of the correlator output.

- The incoming signal is usually composed of signals coming from different satellites that have significantly different received-power levels. It is thus extremely important to have cross-correlation functions between spreading sequences that have very low peaks. The GPS C/A code family provides an isolation of −21.9 dB [8, 9], meaning that if there is a power difference that is greater than this value, the receiver might mistake the desired signal for the interfering signal.

The correlator's output is a key value used by the whole signal processing chain: acquisition, tracking, and demodulation. Acquisition and demodulation are not discussed in this paper; they are well documented in [4, 8, 9]. Tracking is usually based on the use of two parallel loops: a phase-tracking loop that estimates the phase (and frequency) of the incoming signal, and a code-delay-tracking loop that estimates the delay of the spreading sequence. The implementation of these two loops is a suboptimum version of the maximum-likelihood estimator of the phase and code delay. More details on the optimum estimator can be found in [3]. Note that some receivers use a frequency-tracking loop instead of the phase-tracking loop, because they are more robust [1, 2, 4, 5]. However, this is not considered hereafter. The code-delay and phase-tracking loops will be investigated in the following sections.

3.2 Phase Tracking

The goal of phase tracking is to generate a carrier the phase of which is perfectly synchronized with the phase of the incoming signal. Based on the correlator's output, the phase-locked loop (PLL) uses a phase-discriminator function that roughly estimates the phase-tracking error. This discriminator output is then filtered by a low-pass filter. The resulting quantity is used to drive a voltage or numerically controlled oscillator (VCO or NCO), which will generate the local replica of the carrier for the next correlation period. Because the functioning of a typical phase-locked loop is well documented in numerous books and articles [1, 2, 4, 8, 9], and because the goal of this article is to show how Galileo (and, indirectly, other modernized/new global navigation satellite systems) will improve tracking performance, this section will concentrate on only two of the phase-locked loop's critical elements: the phase discriminator and the coherent integration time.

3.2.1 Critical Elements

The first critical element of the phase-locked loop is the phase discriminator. It uses the correlator's output in order to extract the current phase-tracking error (as suggested by Equation (6)). However, because the sign of the navigation data bits provides a 180° uncertainty in the phase of the correlator's output, it is necessary for a receiver to rely on a phase discriminator that is independent from the navigation data bit sign. Such phase discriminators are the Costas discriminator (which uses the product of the real and imaginary part of the correlator's output), and the arctangent discriminator (which uses the arctangent of the ratio of the imaginary part to the real part of the correlator's output) [1, 2, 4, 8-10]. The corresponding phase-locked loop is usually referred to as a Costas loop. The crucial characteristics of these types of discriminators are:

- The stability region within which the output of the discriminator has the same sign as the phase-tracking error. For a Costas loop, it is contained within $[-\pi/2;\pi/2]$.
- The double-sided linearity region, L_ϕ, within which the discriminator's output is proportional to the phase-tracking error. The linearity region of a discriminator is crucial, since it defines the level of errors tolerated in order to have a properly functioning phase-tracking loop. The width of this region will therefore directly impact the resistance of the loop to loss of lock [1, 10, 11]. The double-sided linearity region of a Costas loop is significantly lower than 2π.
- The intrinsic ambiguity of π. A Costas loop has an infinite number of stability regions, separated by π radians. When the loop loses lock and re-locks shortly after on a different tracking point, it is referred to as a cycle slip. Looking at Equation (6), it can be seen that this could result in the inversion of the sign of the demodulated bits, and could thus corrupt the use of the navigation message (the navigation-message parity check will likely fail).

The second critical parameter is the coherent integration time. Its value impacts the signal-to-noise ratio (SNR) at the correlator's output, as can be seen in Equation (6). However, using a long coherent integration duration might not be the right choice, because (1) the coherent integration time should be smaller than the data-bit duration, and (2) the phase-locked loop might not be able to follow fast signal dynamics or the receiver's oscillator drift. Due to the presence of a 50 Hz navigation message, the typical coherent integration time duration for the GPS C/A is 20 ms [4, 5, 8].

Other important phase-locked loop parameters, such as the loop order or the equivalent loop bandwidth, are not discussed in this section. This is because their values will have limited influence on the phase-locked loop, whether future or current navigation signals are received. A typical equivalent loop bandwidth for a global navigation satellite system receiver is between 10 and 20 Hz, and the phase-locked loop is usually a third-order loop [4, 5, 8].

3.2.2 Overall Phase-Tracking Performance

The three typical phase-locked-loop performance criteria are the tracking accuracy, the tracking threshold, and the cycle-slip occurrence rate.

Oscillator phase noise, oscillator vibration, signal dynamics, thermal noise, and multipath are the main sources of error impacting phase-tracking accuracy [4, 5, 8]. The levels of errors induced by oscillator phase noise and signal dynamics, although very important to setting the phase-tracking loop [4, 5, 8, 10-12], will not be significantly changed by the use of the future global navigation satellite system signal structures. They are thus omitted in this section. Of interest to the rest of the paper is the impact of thermal noise and multipath on the phase-locked loop accuracy:

The expression of the phase-tracking error variance due to thermal noise for a Costas loop is given by [1, 4, 5, 8]

$$\sigma^2_{PLL,Th} = \frac{B_L^{PLL}}{\frac{C}{N_0}}\left(1+\frac{1}{2\frac{C}{N_0}T_I}\right)\ [\mathrm{rad}^2], \qquad (7)$$

where B_L^{PLL} is the phase-locked-loop equivalent loop bandwidth in Hz.

Due to the high frequency of the carrier, the resulting tracking error in meters is very small, on the order of a few millimeters. It can be seen that the tracking-error variance depends directly on the equivalent loop bandwidth. Also, the second term in Equation (7) is referred to as the squaring losses, meaning that it is due to the multiplication or division of the correlator outputs, thus increasing the noise level. It can only be mitigated using long integration times, which is not possible in the case of the GPS C/A signal.

Regarding multipath, the usual figure of merit is the multipath envelope. It is obtained by looking at the phase-discriminator stability point, assuming that only the direct signal and one multipath with a given amplitude, delay, and phase relative to the direct signal are present [13]. Because the correlation operation is linear, the resulting correlator output equals the sum of two correlator outputs that would separately process the direct signal and the multipath. This leads to a distorted correlator output that will impair phase tracking. It can be shown that the resulting multipath error is bounded by the cases where the multipath is in-phase and out-of-phase with respect to the direct signal. The ensuing envelope is shown in blue in Figure 4 for the GPS C/A case. In this figure, it can be seen that the multipath envelope is null when the multipath delay is greater than one chip, which is due to the correlation property of the spreading sequence. The shape of the phase multipath envelope is the same as the shape of the spreading-sequence autocorrelation function: decreasing linearly with the multipath delay for the GPS C/A case.

The tracking threshold represents the minimum C/N_0 value at the correlator output that can be accepted by the loop before losing lock. The typical criterion to assess this threshold is given by [4, 5] as

$$3\sqrt{\sigma_{PLL,Th}^2 + \sigma_{PLL,Vib}^2 + \sigma_{PLL,Osc}^2} + \theta_{Dyn} \le \frac{L_\phi}{2}, \quad (8)$$

where $\sigma_{PLL,Vib}^2$ is the variance of the phase-tracking error due to the oscillator vibrations, $\sigma_{PLL,Osc}^2$ is the variance of the phase-tracking error due to the oscillator phase noise, and θ_{Dyn} is the phase-tracking bias due to a constant signal dynamic.

The justification behind Equation (8) is that the tracking error should remain within the discriminator's linear range in order for the loop to exhibit a "normal" behavior. It is therefore possible to compute the tracking threshold as a function of the loop bandwidth for given maximum signal dynamics and oscillator phase noise. Assuming an equivalent loop bandwidth around 10 Hz and typical mass-market users (temperature-compensated oscillator and "reasonable" dynamics), the typical phase-locked-loop tracking threshold for the GPS C/A signal is between 27 and 30 dB-Hz [10, 11].

The last criterion to assess the correct behavior of the loop is the cycle-slip occurrence rate. It has already been mentioned in Section 3.2.1 that a cycle slip could be very detrimental to the decoding of the navigation message, or to the use the of phase measurements as a source of differential ranging. Reference [10] assessed the occurrence of cycle slips as a function of the tracking-error level and the discriminator stability region. As expected, it clearly showed that a wide discriminator stability region and a small tracking error resulted in a lower cycle-slip occurrence rate.

To conclude, phase-tracking error is at the centimeter level, and is thus extremely accurate. The counterpart is that it is very sensitive to cycle slipping and is ambiguous. This makes the phase-locked loop not very robust, and therefore it is usually the first element to lose

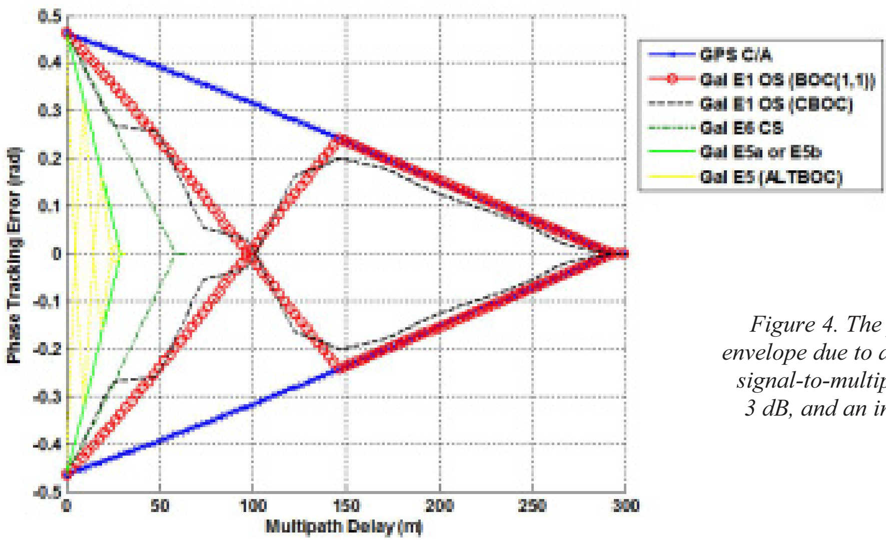

Figure 4. The phase-tracking error envelope due to a single multipath with a signal-to-multipath amplitude ratio of 3 dB, and an infinite front-end filter.

lock in difficult conditions. This is a crucial point for the overall receiver robustness, since the receiver usually loses lock shortly after a phase-locked-loop loss of lock [4, 5, 8].

3.3 Code-Delay Tracking Performance

Code-delay tracking is very similar to phase tracking in its principles, but uses a different discriminator. Code-delay tracking is based on a delay-locked loop (DLL) architecture that uses several correlator outputs in parallel. The receiver needs to generate two complex local replicas symmetrically with respect to the elected prompt replica: one that is delayed, and one that is advanced. The corresponding correlator outputs are referred to as the early and late correlator outputs. The delay, δ, between the corresponding local replicas is referred to as the correlator spacing. Choosing a correlator spacing lower than one chip, it is possible to form a discriminator that uses the symmetry of the spreading-waveform autocorrelation function to estimate the code-delay tracking error. More information can be found in [1, 2, 4, 5, 8].

3.3.1 Critical Elements

Due to the conceptual proximity of the delay-locked loop with the phase-locked loop, the delay-locked loop and phase-locked loop critical elements are very similar, and will not be further described in this section. However, the delay-locked loop's performance also depends significantly upon the correlator spacing, the impact of which will be investigated in the following section.

There are several discriminators that can be used in a delay-locked loop. They are usually categorized as coherent (sensitive to carrier-phase offset) or non-coherent (insensitive to carrier-phase offset) [1, 2, 4, 5, 8]. Typical receivers use the latter, such as the early-minus-late-power (EMLP) discriminator, because of their increased robustness.

3.3.2 Overall Code-Tracking Performance

Because phase tracking is the weak link in a global navigation satellite system receiver, and since the delay-locked loop discriminators are not ambiguous (in principle they are, since the spreading sequence repeats itself, but the resulting ambiguity can be very easily solved), the main figure of merit for the code-tracking loop is its accuracy. Since the delay-locked loop uses the output of the phase-tracking loop in most of the receivers in order to compensate for the signal dynamics, the delay-locked-loop equivalent loop bandwidth can be significantly reduced compared to the phase-locked loop bandwidth (only the residual dynamic error and atmospheric delay variations have to be tracked). A typical delay-locked-loop equivalent loop bandwidth is around 1 Hz. Consequently, the dominant sources of errors are thermal noise and multipath.

Assuming a non-coherent early-minus-late-power discriminator and an infinite front-end filter, the theoretical code-tracking noise variance can be estimated as [11]

$$\sigma_{DLL,Th}^{2} = \frac{B_L^{DLL}\delta}{2\frac{C}{N_0}\alpha}\left[1+\frac{2}{\frac{C}{N_0}T_I\left(2-\alpha\delta\right)}\right] \text{ [Chip}^2\text{]}, \quad (9)$$

where α is the slope of the normalized autocorrelation function of the spreading waveform around $\delta/2$ chip, and B_L^{PLL} is the delay-locked-loop equivalent loop bandwidth.

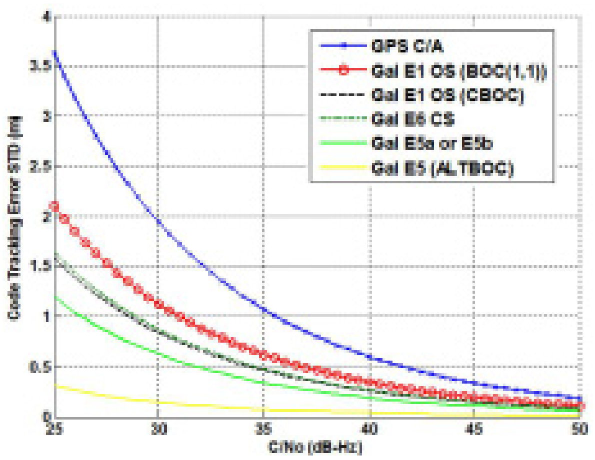

Figure 5a. The code-tracking error due to thermal noise assuming an infinite front-end filter (a correlator spacing of 1/12.2276e6 sec was chosen except for the ALTBOC, where 1/40.92e6 sec was chosen).

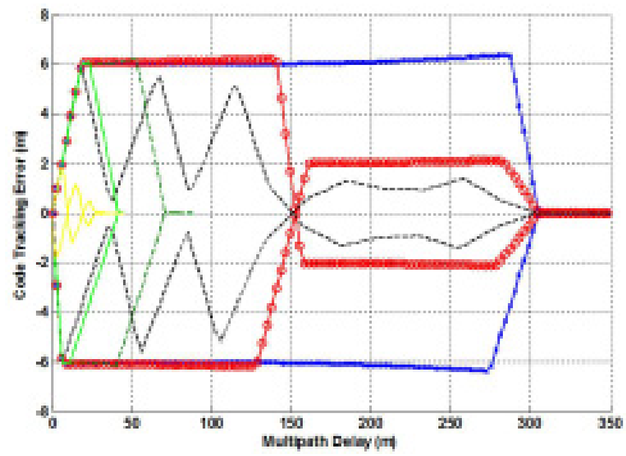

Figure 5b. The code-tracking error envelope due to a single multipath with a signal-to-multipath amplitude ratio of 3 dB, assuming an infinite front-end filter (a correlator spacing of 1/12.2276e6 sec was chosen except for the ALTBOC, where 1/40.92e6 sec was chosen).

Equation (9) shows that:

- The impact of thermal noise on the delay-locked loop depends upon the shape of the spreading-sequence correlation function. It can be seen that the steeper the correlation function, the lower the code-tracking noise. It is important to remember here that the sharpness of the autocorrelation function's main peak is directly related to the spectral repartition of the signal's power spectral density: the higher the power repartition at high frequencies, the sharper the correlation function [14]. In the case of the GPS C/A, α equals one (assuming an infinite front-end filter), as can be seen in Figure 1.
- It is beneficial to have a narrow correlator spacing. This is generally true for a very wideband front-end filter, but not for narrowband front-end filters. Indeed, by filtering out the high-frequency components of the spreading sequence, a narrowband receiver will round off the peak of the correlation function and prevent the option of a very narrow correlator spacing [4, 5, 9, 15, 16]. This would reduce the tracking accuracy.
- There is a squaring-loss term, as for the phase-locked-loop tracking error, which can be reduced using long correlation periods. However, for the GPS C/A this period is limited to 20 ms.
- The tracking accuracy will depend upon the chipping rate. A higher chipping rate will result in more accurate tracking in meters.

The resulting tracking accuracy in thermal noise is shown in Figure 5 for the case of the GPS C/A signal, a 1-Hz equivalent loop bandwidth, a 20-ms correlation time, and a narrow correlator (1/12th of a chip). Note that typical mass-market GPS C/A receivers have a 2 MHz front-end filter and a one-chip correlator spacing.

Regarding the impact of multipath on the delay-locked loop, the same methodology as that shown for phase tracking in Section 3.2.3 is generally used. The result is again a multipath envelope, as shown in Figure 6. This envelope becomes null when the multipath delay goes above $1+\delta/2$ chips. Moreover, the amplitude of the envelope also depends upon the correlator spacing: a narrow correlator spacing means better multipath rejection. As already explained previously, the choice of the correlator spacing is conditioned by the front-end filter's bandwidth.

3.4 Conclusions

In this section, the basics of code and phase tracking of the GPS C/A signal were exposed, and their main constraints were emphasized:

- The spreading-sequence properties (rate, length), which
  - influence the resistance towards multipath (no impact for a delay greater than one chip),
  - impact the isolation of the cross-correlation peaks with respect to the autocorrelation main peak, and
  - constrain the value of the minimum coherent integration time.
- The presence of the navigation data bit, which
  - prevents the reduction of the squaring losses through a longer integration time, and
  - calls for a phase discriminator that is insensitive to 180° phase jumps.

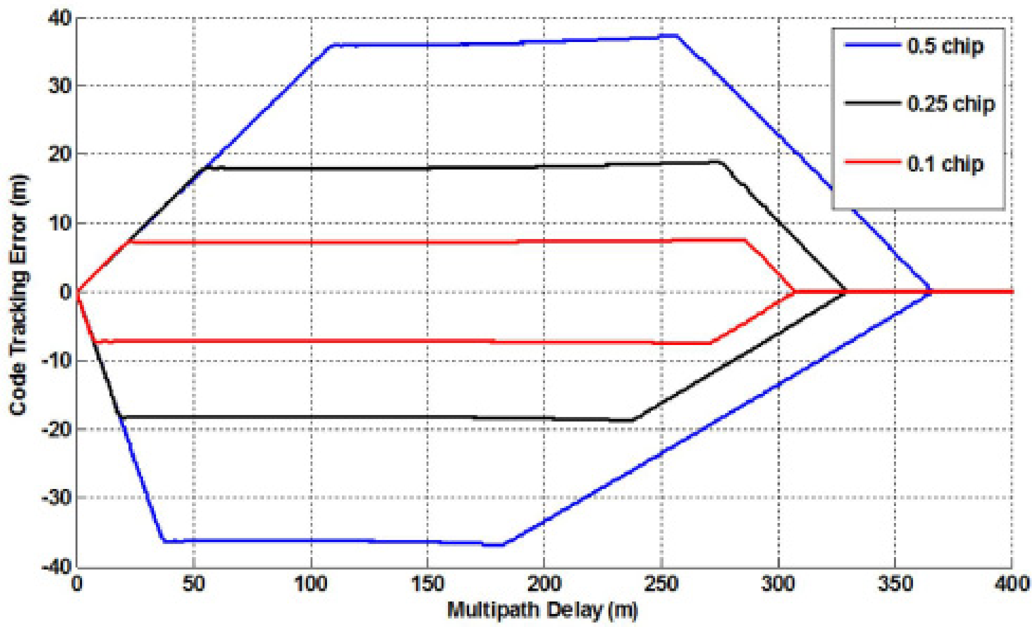

Figure 6. The code-tracking multipath envelope for several correlator spacing values, an infinite front-end filter, and a signal-to-multipath amplitude of 3 dB.

- The shape of the spreading-waveform autocorrelation function, which
  - impacts the multipath resistance of the code delay and phase-tracking loop, and
  - influences the resistance of the code-delay tracking loop towards thermal noise.

These three points can be tackled through three innovations that will be present in the future Galileo: (1) the use of new spreading codes; (2) the presence of a data-less, or pilot, channel; and (3) the use of binary-offset carrier (BOC) modulation. These three points will be analyzed in the next section.

4. Innovations Present in Galileo

Based on the conclusions presented in Section 3.4, most of the future and/or modernized global navigation satellite systems have started implementing new signal structures, in order to improve the overall receiver performance and to provide measurements to the user that are of better quality. The following presents the main innovations used by Galileo (and that can be found in other systems).

4.1 Code Properties, Code Length and Secondary Codes

It was seen in Section 3.1 that the Gold code family used by the GPS C/A signal is 1023 chips long, and provides a cross-correlation isolation of around 21.9 dB. When looking towards new applications developed in challenging environments, such as indoor or urban canyons, it is more and more common to have power differences between signals that are close to or even greater than that value (for instance, a signal arriving through a window, while another signal from another satellite is arriving through a concrete wall [17, 18]). It is thus important to offer more isolation with future signals. There are different ways to improve the code properties:

- Increase the code length. The theoretical minimum of the maximum values of the auto- and cross-correlation side-peaks for a set of M codes of length n is given by the normalized Welch bound [19, 20]:

$$\theta_{welch} = \sqrt{\frac{M-1}{nM-1}} \xrightarrow[n>>1]{} \frac{1}{\sqrt{n}} \qquad (10)$$

  Consequently, increasing the code length also increases the isolation of the autocorrelation main peak with respect to the highest side-peak (pending the right set of codes being selected!).

- Find a better code family: the ultimate choice is the use of memory codes. They correspond to a family of codes that have been selected among all other codes because they offer the best performance according to predefined constraints [19].

Regarding the first bullet, it is usually not possible to increase the spreading-sequence length too much, as it would be detrimental for the acquisition time [4, 5]. It is then possible to further improve the spreading-sequence auto- and cross-correlation properties by using a tiered code. This consists of multiplying the pseudo-random-noise sequence by a secondary binary code. Each chip of the secondary code is multiplied by the full "primary" pseudo-random-noise sequence, as shown in Figure 7. By doing so, the tiered code imitates a longer code based on a shorter primary sequence. However, the resulting auto- and cross-correlation properties are not as good as those of a spreading code the length of which is the product of the lengths of the primary and secondary codes. The receiver can then either use a correlation over the primary code only (this is typically

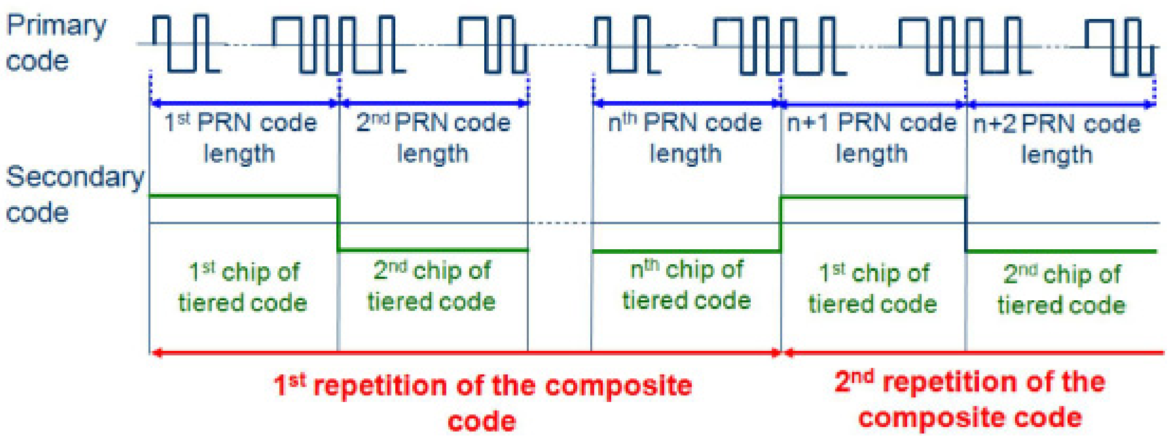

Figure 7. The generation of tiered codes.

used for acquisition purposes, when the receiver is not yet synchronized), or a correlation over the whole tiered code to improve cross-correlation properties or better mitigate narrowband interference, for example [21, 22].

Finally, an easy way to improve tracking performance is to increase the spreading-code rate. As seen throughout Section 3, the code-tracking performance was given in units of chips (Equation (9), for instance). This means that using a higher chipping rate would improve the tracking performance. It is then possible to better reject multipath and thermal noise for the delay-locked loop, and multipath for the phase-locked loop. However, in this case, the receiver will also be more complex.

4.2 Pilot Channel

A pilot, or data-less, signal represents a signal that does not contain any navigation message. The reason for using such a signal is to remove the limitations related to the presence of the navigation data that were exposed in Section 3 (limited integration time, phase discriminator insensitive to data-bit transition, etc.). The counterpart is obviously the absence of the navigation message, which means that with only a pilot signal, it is impossible to compute the user's position. Consequently, in a global navigation satellite system, a data signal is always sent synchronously with a pilot signal. The resulting signal can be expressed at baseband as

$$s(t) = a d(t) c_D(t) + b c_P(t), \qquad (11)$$

with $a^2 + b^2 = 1$, where c_D and c_P are the data- and pilot-channel spreading sequences, respectively.

The power required to send the composite signals has then to be split between the data and pilot channels. Since the power resources in a satellite payload are limited, it is then important to make sure that the resulting performance of the pilot signal for code and phase tracking is greater than that of a data signal with the full available power.

4.2.1 Advantage for the Phase-Tracking Loop

The presence of a pilot channel means that a true phase-locked loop can be used. This has primarily two impacts:

- The phase discriminator does not have to be insensitive to 180° phase jumps, since no data is modulating the carrier. This means that more optimal discriminators can be used. In particular, it is possible to find discriminators that have a wider linear region. Examples of such phase discriminators are the Q and four-quadrant arctangent discriminators [11, 22]. They have several advantages over the Costas discriminator:
  - a wider stability region, and
  - a 2π ambiguity instead of a π ambiguity for the Costas discriminator.
- It is possible to use a longer correlation time. This means that it is possible to increase the SNR at the correlator's output. Obviously, this does not mean that the correlation time can be infinite, since the tracking loops still have to track the signal dynamics, as well as the oscillator phase error. Assuming a long-enough correlation time, the resulting phase-tracking noise variance becomes

$$\sigma^2_{PLL,Th} = \frac{B_L^{PLL}}{\frac{C}{N_0}} \ [\mathrm{rad}^2]. \qquad (12)$$

The result is a phase-locked-loop tracking threshold improved by approximately 6 dB compared to a Costas loop that processes a data signal with the same C/N_0 [11, 22]. Consequently, even if the signal power is equally split between the data and pilot channels, there is still an improvement of the tracking threshold of about 3 dB with respect to a Costas loop processing a data channel with the full power. This is tremendously important, as the phase-tracking loop is the weak part of the receiver-tracking process. Thanks to the wider stability region, the cycle-slip-occurrence rate using a pilot channel is also significantly reduced, compared to the case of a data channel [11, 22]. Consequently, phase tracking will be more reliable, and thus phase measurements will be easier to use as a differential-ranging source.

Moreover, it has to be kept in mind that the data and pilot components are sent synchronously. Tracking the pilot component's phase thus means that phase synchronization is also intrinsically achieved on the data component. Because there is a 2π ambiguity using the pilot channel, the data demodulation will never be ambiguous.

4.2.2 Advantage for the Code-Delay Tracking Loop

Code-delay tracking mostly uses the pilot channel in order to limit the squaring losses in Equation (9). Indeed, since a typical delay-locked loop is aided by the phase-locked loop, the residual dynamic error is very limited, and a very long correlation time can be used. It is then also possible to integrate over the whole secondary code if a tiered code is used (it should be remembered that according to the oscillator quality and signal dynamics, that might not be possible on the phase-locked loop). This is very important for reducing the level of noise affecting the delay-locked loop and its resistance to narrowband interference.

4.2.3 Conclusions

This section showed that both code and phase tracking can be significantly improved by the presence of a pilot channel. In particular, the resulting lower tracking threshold of a pure phase-locked loop compared to a Costas loop means that the overall receiver-tracking threshold will be improved. This means that a typical receiver will track the incoming signal based only on the pilot channel, while using the data channel only for demodulation purposes.

Also, the use of long correlation durations is equivalent to the use of a low-pass filter. Consequently, all the disturbances that change quickly over time, such as interference or multipath, can be significantly reduced. This is particularly true for dynamic users.

4.3 Binary Offset Carrier Modulation

Binary offset carrier (BOC) modulation [14, 23] is obtained from the multiplication of the pseudo-random-noise sequence by a square-wave subcarrier (which can be derived from a sine or a cosine function), with certain constraints on this sub-carrier:

- The subcarrier is synchronized with the spreading sequence, and
- Twice the subcarrier frequency, f_{sc}, is a multiple of the spreading-sequence rate, f_c.

In global navigation satellite systems, the frequencies are usually expressed with respect to the frequency 1.023 MHz. In the present case, the following notation is generally used: $f_{sc} = mf_0$ and $f_c = nf_0$. The BOC modulation is then typically referred to as sine- or cosine-BOC(m,n). Using the same notations as in Equation (2), the baseband temporal expression of a sine-BOC(m,n) signal is

$$s_{BOC(n,m)}(t) = d(t)c(t)SC(t), \qquad (13)$$

where c has a chipping rate of $f_c = nf_0$, and $SC_m(t) = \text{sign}\left[\sin(2\pi m f_0 t)\right]$.

The use of a BOC modulation impacts the autocorrelation function of the pseudo-random-noise sequence and, obviously, its power spectral density. BOC theoretical power spectral densities were provided in [14, 24]. Examples of BOC power spectral densities are given in Figure 8 for different combinations of m and n. As can be seen, they are characterized by their two main sidelobes, offset by f_{sc} from the carrier frequency. The width of these lobes is equal to $2nf_0$ MHz. This split spectrum has the benefit of avoiding an overlap with existing legacy signals' power spectral densities, which are usually located on the carrier frequency, as for the GPS C/A (see Figure 3) [14]. Using the flexibility of f_{sc}, it is possible to "organize" the signal so that it does not interfere with other signals in the same band, even if other BOC signals are present. This is important for respecting compatibility issues with other systems [25]. It is also interesting to note that a single sidelobe of a BOC(m,n) signal can be tracked as a BPSK(n) signal, thus providing the receiver with an interesting means of resistance against interference, at the expense of a slightly degraded correlation operation [11, 26].

The BOC autocorrelation function is a multi-peak autocorrelation function, the characteristics of which depend upon the values of m and n, and whether a sine or a cosine function is used to generate the square subcarrier. The higher is the ratio m/n, the higher is the number of side peaks and the magnitude of the central peaks [14]. Examples

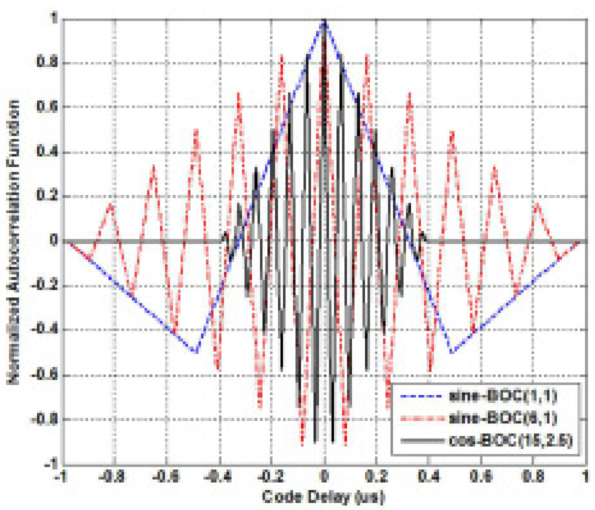

Figure 8a. An example of a BOC autocorrelation function.

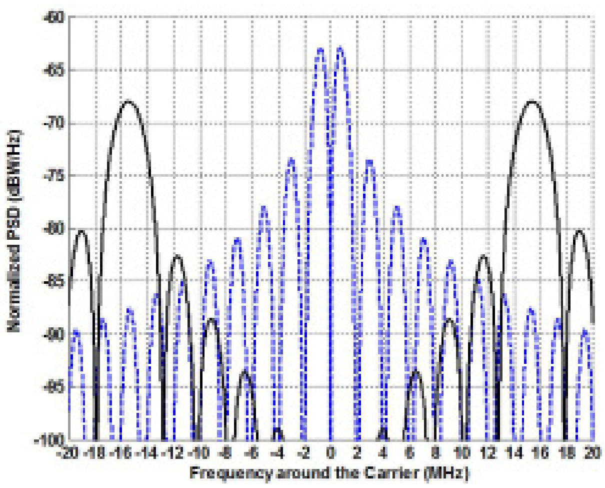

Figure 8b. An example of a BOC power spectral density function.

of BOC autocorrelation functions are shown in Figure 8. It can be seen that the main peak of the BOC(m,n) autocorrelation function will always have a steeper slope than that of the BPSK(n). This has several impacts on code-delay tracking, since, as seen in Section 3.3, the resistance to multipath and to thermal noise is dependent upon the signal's autocorrelation function. In particular, looking at Equation (9), it is clear that the delay-locked-loop tracking variance due to thermal noise will be reduced by a factor equal to the slope of the BOC autocorrelation's main peak.

There are primarily two drawbacks associated with the typical BOC tracking technique. The first is due to its wider spectral occupation, which forces a BOC(m,n) receiver to have a minimum front-end filter that is greater than that of a BPSK(n). This means that BOC(m,n) receivers will generally be more complex than BPSK(n) receivers. The second drawback is due to the multiple peaks of the BOC autocorrelation function. Indeed, a typical delay-locked loop is designed to track the peak of the autocorrelation function. In the case of a BOC signal, this means that if no care is taken, it is possible for the receiver to track the wrong peak and to deliver a biased measurement to the user, thus affecting the user's ability to position itself accurately. Methods were developed to remove this problem [11, 25, 27, 28]; however, these are once again at the expense of receiver simplicity.

Now that the three main innovations contained in Galileo have been introduced in a general way, they will be placed in the actual context of Galileo's signal structure.

5. Galileo's Signal Structure and Performance

5.1 Overview of Galileo's Services and Frequency Plan

Galileo will broadcast signals on four different frequency bands [29, 30]:

- The E1 band (1559 MHz to 1591 MHz) is centered around 1575.42 MHz. This band is an Aeronautical Radio Navigation Service (ARNS) band, with exclusivity to radio navigation satellite systems, and thus global navigation satellite systems. An ARNS band has the advantage of limiting the in-band interference environment, since it is regulated by stringent aviation constraints. The E1 band is superimposed onto the GPS L1 band (1563 MHz to 1587 MHz). It is a crucial global navigation satellite system band, because it is located where the GPS L1 C/A signal is located. Consequently, all the current receivers built for civil use have an architecture adapted to this band, in particular, the antenna and the front end. Consequently, this band will likely remain the global navigation satellite system band where most of the signals dedicated to the mass market are located.

- The E6 band is located within (1260 MHz to 1300 MHz). It is centered around 1278.75 MHz. It is not an ARNS band, and GPS does not broadcast in this band. It is thus a differentiator of Galileo with respect to GPS (however, other systems, such as QZSS, will be present in this band).

- The Galileo E5a (1164 MHz to 1191.795 MHz), centered on 1176.45 MHz, and E5b (1191.795 MHz to 1214 MHz), centered on 1207.14 MHz, bands. They are part of the same transmitting band, the Galileo E5 band (1164 MHz to 1214 MHz), the central frequency of which is 1191.795 MHz. The E5 band is the largest global navigation satellite system band. It is also an ARNS band, but with no exclusivity to radio-navigation satellite systems. In particular, other aeronautical RF systems, such as distance measuring equipment (DME), are deployed in this band, thus providing some interference to Galileo's (or any global navigation satellite system's) signals. Note that GPS will broadcast its L5 signal in the same frequency band as E5a. On the other hand, E5b is a differentiator from GPS.

The choice of Galileo signals has been made around the notion of services. Four services will be delivered by the Galileo system [29, 30]:

- The Open Service (OS), which will be free of charge, and which is intended for mass-market users. Three signals in three different frequency bands will support this service (E1, E5a, E5b). The first draft of the Open Service Signal-in-Space Interface Control Document is provided in [31].

- The Commercial Service (CS), which will be dedicated to commercial applications, with the need to bring an extra value to the user through the navigation message (it is not freely accessible) and the number of frequencies used. Four signal signals in four different frequency bands will support this service (two are common with the Open Service/Safety-of-Life – E1, E5a and E5b – and one is specific – E6). Some applications of the Commercial Service could be limited to E6 Commercial Service, Open Service/Safety-of-Life E1, and E5a or E5b.

- The Safety-of-Life service (SoL), freely available, which is aimed at applications requiring integrity such as civil aviation, rail, or sailing applications. Two signals in two different frequency bands will support this service (common with the Open Service and Commercial Service: E1 and E5b). Moreover, Safety-of-Life users could use three Galileo signals: the two Safety-of-Life signals, and the remaining Open Service signal (E5a), also situated in an ARNS band.

Table 1 : The characteristics of Galileo Signals

Frequency Band (central frequency)	Service	Chipping Rate (MHz)	Primary Code Length	Secondary Code (length)	Nav. Data Rate (sps)	Modulation	Number in Fig. 9	Multiplex Scheme
E5a (1176.45 MHz)	OS	10.23	10230*	Yes (100)	pilot	BPSK(10)	1	ALTBOC (15,10)
		10.23	10230*	Yes (20)	50	BPSK(10)	2	
E5b (1207.14 MHz)	OS, CS, SoL	10.23	10230*	Yes (100)	pilot	BPSK(10)	3	
		10.23	10230*	Yes (4)	250	BPSK(10)	4	
E6 (1227.6 MHz)	CS	5.115	5115	No	pilot	BPSK(5)	5	N/A
		5.115	5115	Yes (100)	1000	BPSK(5)	6	
	PRS	5.115	N/A	N/A	N/A	Cos-BOC(10,5)	7	
E1 (1575.42 MHz)	OS, CS, SoL	1.023	4092*	Yes (25)	pilot	CBOC(6,1,1/11,'−')	9	Interplex
		1.023	4092*	No	250	CBOC(6,1,1/11,'+')	10	
	PRS	2.5575	N/A	N/A	N/A	Cos-BOC(15,2.5)	8	

* *Memory Codes*

- The Public Regulated Service (PRS), which is aimed at public-security applications. Two signals in two different frequency bands are exclusively dedicated to this service (E6 and E1), and the spreading sequences will be highly encrypted in order to be delivered only to the authorized users.

In total, six signals will be transmitted by each satellite in four different frequency bands. It can be deduced from this that certain signals will be used by several services. It is thus necessary to broadcast a navigation message that is adapted to this constraint, which is also an innovation brought by Galileo. It can be seen that all the proposed services are supported by at least two signals present in two different frequency bands. This means that the signal structure offers frequency diversity against interference to the user, as well as the possibility of removing the impact of the ionosphere on pseudo-range measurements [4, 5, 30].

5.2 Galileo's Signal Structure

Table 1 summarizes the characteristics of all the Galileo signals. The spectral occupation of Galileo's signals is presented in Figure 9. It can be seen that all the signals use at least one of the three new components that have been described in Section 4. The civil signals all have a data/pilot architecture favorable to better phase tracking, with 50%/50% power sharing. To compensate for the potentially lower power of the data channel, the navigation-message encoding scheme used by Galileo has been improved, in order to increase the demodulation threshold [29, 31]. A convolutional coding scheme of length seven and rate one-half is used. Also, all the signals use a primary code length greater than 1023, with a chipping rate at least equal to 1.023 MHz. Finally, the BOC modulation is widely used.

In the following, a more-detailed description of the Galileo signals will be given, band by band.

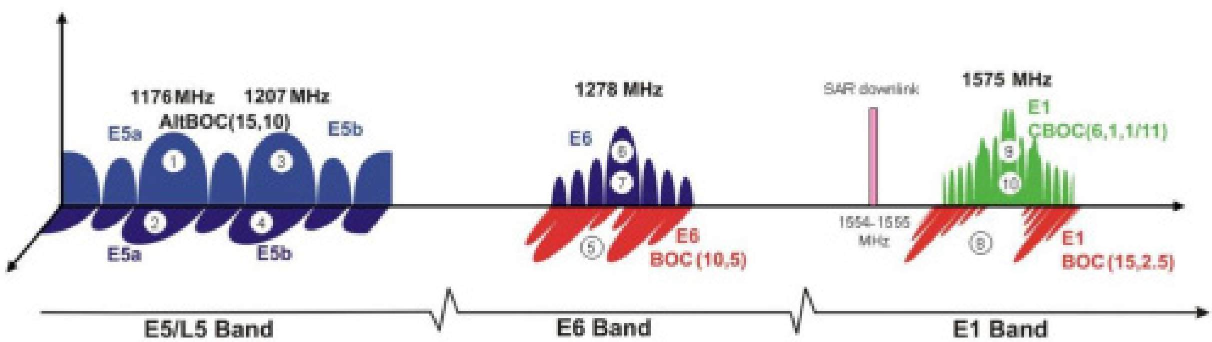

Figure 9. A representation of the spectral occupation of Galileo.

5.2.1 Galileo E1

As noted in Table 1 and Figure 9, the Galileo E1 signal is composed of an Open Service/Commercial Service/Safety-of-Life signal and a Public Regulated Service signal. The Public Regulated Service signal will use a cos-BOC(15,2.5) modulation, which brings the sidelobes of its power spectral density to the edge of the band, thus isolating them from the center of the band where the other global navigation satellite system signals are located.

The Open Service/Commercial Service/Safety-of-Life signal is composed of a data and a pilot channel with 50%/50% power sharing. Both channels use a primary code length of 4092 chips, with a chipping rate of 1.023 MHz. The pilot channel also uses a secondary code. It can be seen that the modulation used by both channels is denoted as composite BOC (CBOC). This modulation is compliant with the US/EU interoperability agreement signed in 2004, deciding that Galileo and GPSIII (third generation) will each broadcast new civil signals on the E1/L1 band that share the same power spectral density. This is referred to as multiplexed BOC (MBOC), and defined as [32, 33]

$$G_{MBOC}(f)=\frac{1}{11}G_{BOC(1,1)}(f)+\frac{10}{11}G_{BOC(6,1)}(f). \quad (14)$$

As seen in Equation (14), the MBOC power spectral density is the sum of 1/11th of a sine-BOC(6,1) power spectral density and 10/11th of the sine-BOC(1,1) power spectral density. Because this modulation is used for the Open Service signal in the E1 band, it was necessary to keep a narrowband signal for the mass-market applications, which is the reason for the sine-BOC(1,1) component. However, it was also interesting to add a signal that could significantly improve the signal performance for more-advanced applications, which can use wideband receivers. The addition of a sine-BOC(6,1) was thus considered as a relevant choice, considering the radio-frequency-compatibility issues related to the occupation of the L1/E1 band by several systems [34]. However, the MBOC is only defined as a power spectral density, and does not specify the type of modulation required to obtain this power spectral density. This is why GPS and Galileo proposed different MBOC implementations for their signals. GPS proposed the time-multiplexed BOC (TMBOC) modulation [32], while Galileo proposed the composite BOC (CBOC) [32, 33]. Herein, only the CBOC is explained. Following the same notations as Equation (13), it can be expressed at baseband as

$$s_{E1OS}(t)=\frac{1}{\sqrt{2}}\left[d(t)c_D(t)CBOC_{\left(6,1,\frac{1}{11},'+'\right)}(t) - c_P(t)c_{Sec}(t)CBOC_{\left(6,1,\frac{1}{11},'-'\right)}(t)\right] \quad (15)$$

with

$$CBOC_{\left(6,1,\frac{1}{11},'+'\right)}(t)=\sqrt{\frac{10}{11}}SC_1(t)+\sqrt{\frac{1}{11}}SC_6(t),$$

$$CBOC_{\left(6,1,\frac{1}{11},'-'\right)}(t)=\sqrt{\frac{10}{11}}SC_1(t)-\sqrt{\frac{1}{11}}SC_6(t),$$

and where c_{Sec} is a secondary code.

This Galileo E1 Open Service pilot channel is represented in Figure 10 for the first 10 chips. It can be seen that it is an unconventional signal, due to its non-binary nature. The receiver will thus have to process a signal

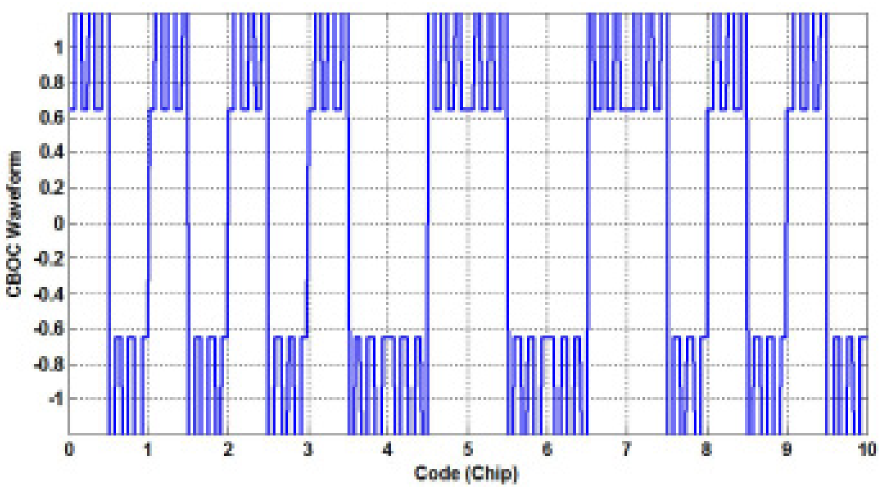

Figure 10. The CBOC(6,1,1/11,'-') subcarrier time representation.

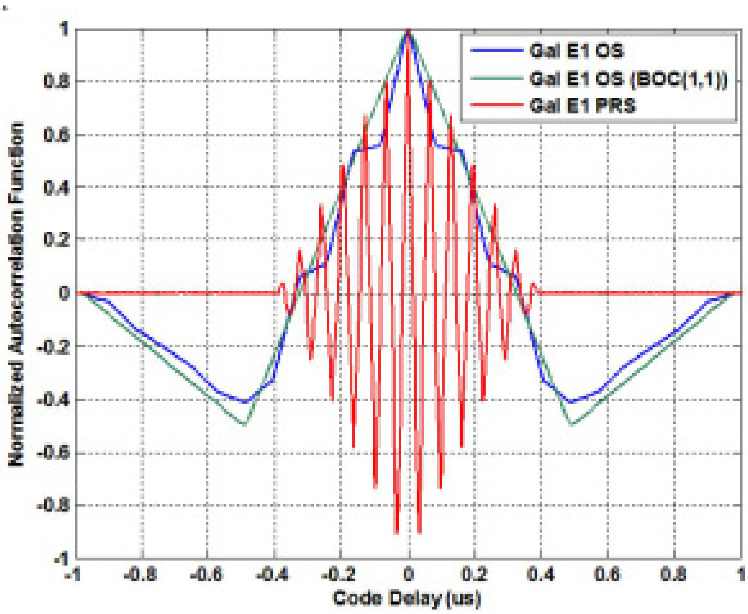

Figure 11. The normalized autocorrelation functions of Galileo signals in the E1 band.

significantly different from the usual binary signals. Note that a narrowband receiver can still track only the sine-BOC(1,1) component, with only a very limited power loss. For comparison, the TMBOC implementation retained by GPS for the GPS L1C signal is composed of a pure sine-BOC(1,1) on the data channel (with 25% of the total power), and a time-multiplexing of the sine-BOC(1,1) and sin-BOC(6,1) components on the pilot channel (with 75% of the total power). Four chips out of every 33 chips are modulated by a sine-BOC(6,1) subcarrier on the pilot channel, thus avoiding a multi-bit subcarrier, but requiring time multiplexing. More information on the TMBOC can be obtained from [30].

The autocorrelation function of the Galileo E1 signals is represented in Figure 11. It can be seen that the Galileo E1 Public Regulated Service (PRS) has a very sharp main peak, which is very appropriate for accurate tracking. It can also be seen that the sine-BOC(6,1) component of the Galileo E1 Open Service/Commercial Service/Safety-of-Life signal significantly modifies the autocorrelation function of the signal, providing a sharper correlation function.

To summarize, five binary components have to be broadcast coherently on Galileo E1: the Public Regulated Service signal, the Open Service sine-BOC(1,1) on the data and pilot channels, and the Open Service sine-BOC(6,1) on the data and pilot channels. It is well known that in order to optimize the power consumption of the on-board high-power amplifier (HPA) and to limit its degradation of the useful signal, it broadcasting a constant-envelope signal is recommended [35, 36]. To do so, Galileo decided to use a five-element interplex modulation – that is, a phase-shift-keyed/phase modulation (PSK/PM) [35, 36] – which is also specific to Galileo. This solution requires the generation of an extra (or sixth) binary signal, referred to as the intermodulation (IM) signal, to ensure a constant-envelope modulation. This intermodulation term is not a useful signal and thus wastes some of the transmitted power, but in a very limited way.

5.2.2 Galileo E5a and E5b

Both Galileo E5a and E5b have a data/pilot structure, with 50%/50% power sharing. All the components use secondary codes. Finally, both signals use a QPSK(10) modulation. They thus have a wide spectrum that will enable very good tracking capabilities. In order to transmit these two adjacent signals, the idea was to broadcast a unique signal on E5 with a constant envelope, for the reasons exposed in Section 5.3.1. The proposed solution was the constant-envelope ALTernative-BOC(15,10) (ALTBOC(15,10)) [34]. This is an alternative version of the BOC modulation that allows carrying four distinct channels (with their own spreading sequence, navigation message, etc.) in the same signal. The ALTBOC sidelobes then correspond to the E5a and E5b signals, which are associated with different Galileo services. This modulation is currently only present in Galileo, and other global navigation satellite system systems are considering its usage. The great advantage of this multiplexing scheme is that the resulting signal can carry different services, and can be processed in two different ways:

- Each sidelobe is considered separately as a QPSK(10) signal. In this case, the user can independently access the E5a or E5b signals and services with minimal degradation.
- The whole ALTBOC(15,10) signal can be processed as a single navigation signal. This means that the user will have access to both E5a and E5b messages, and will use a 50-MHz wide signal for navigation purposes. This will result in unmatched tracking performance, at the expense of a complex receiver. High-end precision

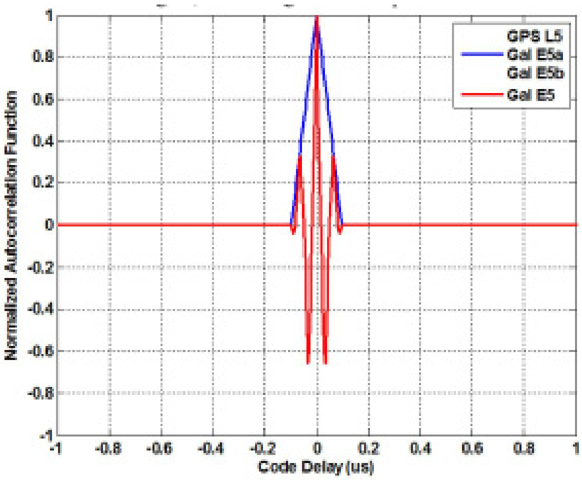

Figure 12. The normalized autocorrelation functions of Galileo signals in the E5 band.

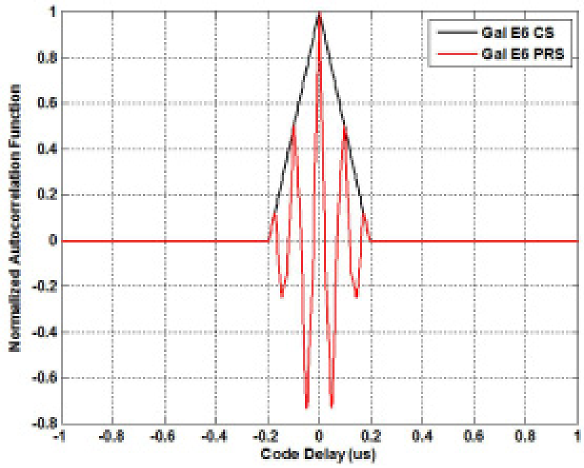

Figure 13. The normalized autocorrelation functions of Galileo signals in the E6 band.

receivers will be more complex if they exploit the full transmitted bandwidth of the ALTBOC, which will be close to 100 MHz.

Similarly to the Galileo E1 interplex, an intermodulation term is necessary to form the ALTBOC modulation and to ensure a constant envelope [36]. Figure 12 represents the normalized autocorrelation function of the Galileo signals on E5, with the same *x* axis as Figure 11, for comparison.

Note that the Galileo E5a signal has the same modulation as the future GPS L5 signal, thus promoting the interoperability of both systems. Another example considers the Galileo E5b signal, which has the same frequency and modulation as the future COMPASS B2 signal, which might be interoperable with Galileo E5b.

5.2.2 Galileo E6

There are three signals broadcasting two Galileo services in this band: two signals dedicated to the Commercial Service, and one signal dedicated to the Public Regulated Service. The Public Regulated Service uses a cosine-BOC(10,5) modulation. The Commercial Service has a data/pilot architecture using a QPSK(5) modulation. These two signals have a wide bandwidth compared to the GPS C/A, thus providing good tracking accuracy. Note also from Table 1 that the Commercial-Service data rate is 1000 sps, since more information is passed to the user to provide the user with added value. Figure 13 shows the normalized autocorrelation function of the Galileo signals on E6.

Now that every Galileo signal has been presented, it is important to quantify their code delays and phase-tracking performance.

5.3 Galileo Signal Performance

5.3.1 Phase Tracking

Since the data/pilot power share of all the Galileo civil signals is 50%/50%, this means that the phase-tracking threshold of all the civil Galileo signals will be around 3 dB below that of GPS C/A (in terms of C/N_0 at the correlator's output), as already mentioned in Section 4.2. Consequently, a typical Galileo civil receiver will have a phase-locked-loop tracking threshold around 24 to 27 dBHz.

It was seen in Section 3.2 that the signal waveform had an impact on the multipath rejection by the phase-tracking loop. The phase multipath envelopes are shown in Figure 4 for the Galileo civil signals, along with the GPS C/A envelope. It can be seen that the envelopes are all significantly reduced, in particular for wideband signals.

The other phase-tracking improvements (cycle-slip occurrence rate, tracking accuracy) were already mentioned in Section 4.2. Since phase tracking is the weak link of a global navigation satellite system receiver, this clearly means that these receivers will be more robust to signal degradation compared to the current GPS C/A civil receivers.

5.3.2 Code-Delay Tracking

Concerning code tracking, the new modulations will impact the mitigation of thermal noise and multipath, since their autocorrelation functions are different. However, it is important to compare their performance in a relevant way. For instance, it is important to consider the same correlator spacing in seconds to really compare the performance of the signals. Because the early and late correlators have to be on the linear part of the main peak of the signal's autocorrelation function, it was decided to choose a correlator spacing adapted to the CBOC autocorrelation function, which is 1/12.2276e6 sec (or one-twelfth of a chip for a spreading code rate of 1.023 MHz). However, for the ALTBOC(15,10), due to its very narrow autocorrelation function, a correlator spacing of 1/40.92e6 sec was chosen. The selected correlation time is 20 ms, and the delay-locked-loop equivalent loop filter bandwidth is 1 Hz. The resulting tracking performance in thermal noise and multipath are shown in Figure 4. Once again, it can be seen that the future Galileo signals bring significant tracking improvement, with exceptional performance for the ALTBOC signal. However, it has to be kept in mind that the tracking of these signals usually requires a more-complex receiver, as seen in Section 4.3, and a front-end filter that is wider than that of a GPS C/A receiver.

5.4 Interoperability and Compatibility with Other Global Navigation Satellite Systems

As was presented in the introductory part, there are several global navigation satellite system/regional navigation satellite system initiatives currently being modernized and/or developed (Galileo, GPS, GLONASS, COMPASS, IRNSS-GINS, QZSS, etc.). These systems are all planning to broadcast several signals on multiple frequencies [37-43]. Because the number of frequency bands allocated to radio-navigation satellite systems is limited, this means that there is a need to coordinate the signal structure of each system, in order to ensure a minimum level of compatibility and interoperability among them. Forums to discuss such aspects exist, such as the International Committee on Global Navigation Satellite Systems (ICG).

A definition of compatibility (not definitive) given by the ICG [44]

> ...refers to the ability of global and regional navigation satellite systems and augmentations to be used separately or together without causing unacceptable interference and/or other harm to an individual system and/or service.

- The International Telecommunication Union (ITU) provides a framework for discussions on radiofrequency compatibility. Radiofrequency compatibility should involve thorough consideration of detailed technical factors, including effects on receiver noise floor and cross-correlation between interfering and desired signals.
- Compatibility should also respect spectral separation between each system's authorized service signals and other systems' signals. Recognizing that some signal overlap may be unavoidable, discussions among providers concerned will establish the framework for determining a mutually acceptable solution.
- Any additional solutions to improve compatibility should be encouraged.

Although compatibility between systems is essential, interoperability is also a key element of system design. However, it is a more-complex topic, since it is dependent upon the readiness of different global navigation satellite system organizations to share common views on the system architecture (e.g., realization of a common reference frame), services (e.g., complementary service to benefit the user), or signal design (e.g., same signal broadcast, or same frequency band used in order to facilitate the development of dual-system receivers). Obviously, the goal of interoperability is to provide benefits to the user while maintaining the system's objectives. A proposed definition of interoperability given by the ICG (not definitive) [44]

> ...refers to the ability of global and regional navigation satellite systems and augmentations and the services they provide to be used together to provide better capabilities at the user level than would be achieved by relying solely on the open signals of one system.

- Interoperability allows navigation with signals from different systems with minimal additional receiver cost or complexity.
- Multiple constellations broadcasting interoperable open signals will result in improved observed geometry, increasing end user accuracy everywhere and improving service availability in environments where satellite visibility is often obscured.
- Geodetic reference frames realization and system time steerage standards should adhere to existing international standards to the maximum extent practical.
- Any additional solutions to improve interoperability are encouraged.

Even if the ICG definitions of compatibility and interoperability might be slightly modified, it is easy to understand that these two notions are key elements of system design, especially in the current context of five global navigation satellite systems. Galileo is pursuing multilateral (e.g., through the ICG) and/or bilateral discussions with the US, Russia, China, Japan, and India, among others. This is very important for ensuring the relevance of the current Galileo design and its potential evolutions. An example of the US/EU compatibility and interoperability efforts can be seen through the two frequency bands shared by GPS and Galileo with interoperable signals. Also, as seen in Section 5.2.1, the US and EU signed a bilateral agreement in 2004 deciding that Galileo and GPSIII (third generation) will broadcast civil signals on the L1/E1 frequency that will have the same power spectral density. This means that GPS/Galileo receivers can have a unique front end for the E1/L1 frequency, thus reducing the complexity (and the price) of the receiver. The E6 frequency and modulation, common to GALILEO and QZSS, is another notable example of interoperable signals, aimed at broadcasting precise positioning information, among other data. E6 is also considered as a signal option for one version of the GPS III system. Interoperability also concerns navigation and integrity message designs. An important challenge in the future will be not only the search for compatibility and interoperability in global navigation satellite system L bands, but also in the S band, targeted by GALILEO, Beidou/Compass, and IRNSS in a band already used by GLOBALSTAR. This is an extension of the interoperability definition to include synergy between mobile communications and navigation.

6. Conclusions

This article showed the structure of the GPS C/A signal, the only fully available global navigation satellite system civil signal, as well as its limitations when it comes to tracking. Several ideas were developed to overcome these limitations and to propose a more-robust and accurate signal for more-demanding user communities. Three main innovations were presented in this article: the use of better spreading-code properties, the use of a pilot channel, and the use of BOC-based modulations. These three innovations were shown to significantly improve the tracking ability of the receiver.

Based on Galileo's signal structure, it has been shown how these three ideas were used by the European system to considerably improve the receiver's performance for all users. In particular, a Galileo receiver will provide to the user code-delay measurements that are less affected by multipath and thermal noise, thanks to the BOC-based modulations and the tiered codes. It will also allow a lower receiver-tracking threshold, since phase tracking is improved by the use of a pilot signal. Finally, better spreading sequences will provide users with better performance in difficult environments.

Finally, it is important to remind the reader that this article used the Galileo system to show most of the improvements that will be brought by the next generation of global navigation satellite systems. Parallel initiatives are under development in the US, Russia, China, Japan, and India that will also likely have improved performance with respect to the GPS C/A.

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The Formation and Early Years of URSI Commission K on Electromagnetics in Biology and Medicine

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Abstract

At the time of the next URSI General Assembly in Istanbul, Turkey, it will have been 20 years since Commission K became the International Union of Radio Science (URSI) Commission on Electromagnetics in Biology and Medicine. The latest Commission of URSI held its inaugural scientific sessions at the XXIVth URSI General Assembly, hosted by the historic capital city of Japan, Kyoto, in 1993. How did all of this come about? The answer is given in this paper, written by three radio scientists who were there at the beginning. It also traces the role of URSI in the development of this "new" field of science.

1. Introduction

Indeed, URSI's role in the development of this "new" field of science can be traced as far back as the 1972 URSI National Radio Science Meeting, held in Williamsburg, VA, sponsored by the US National Committee (USNC), where scientific sessions on selected topics of bioelectromagnetics were held. Encouraged by the positive experience of USNC, suggestions were put forward that URSI should take an interest in the subject of the interaction of electromagnetic fields with biological systems. An Inter-Commission Working Group (between Commissions A and B) of URSI, on "Measurements Related to the Interaction of Electromagnetic Fields with Biological Systems," was established in 1975 to take up the subject. The Chair was Saul Rosenthal of the Brooklyn Polytechnic Institute of New York. The Inter-Commission Working Group was instrumental in organizing the International Symposium on Biological Effects of Electromagnetic Waves, held at Airlie House, VA, in 1977. It brought together members of the scientific community from North America, Europe, and the Soviet Union.

The Bioelectromagnetics Society (BEMS) was officially created in 1978, as a multidisciplinary membership society to serve the interests of engineers, physicists, biologists, and physicians studying electromagnetics and its interaction with biological systems. In 1979, Commissions A and B of URSI/USNC were cosponsors of the Bioelectromagnetics Society's first official meeting, held jointly with the URSI National Radio Science Meeting.

The General Assemblies of URSI are held at three-year intervals, at the international level. The main objective of the General Assembly is to review current trends in research, to present new discoveries, and to make plans for future research work or for specific projects, especially where it seems desirable to arrange for cooperation on an international scale. Many radio scientists often attend the General Assemblies solely for the scientific activities, which are open to anyone interested in radio science, whether or not they are connected with a national member committee of the Union.

In this case, the General Assemblies provided a triennial forum for the scientific community to convene to review recent accomplishments, discuss the latest results and identify developing trends in bioelectromagnetics. It also furnished a meeting structure for URSI and BEMS to encourage the cooperation that preceded the creation of BEMS. The collaboration was manifested in the organization of the 1984 URSI Open Symposium on Biological Effects

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of Electromagnetic Waves, during the XXIst URSI General Assembly, held in Florence, Italy. It also led to the organization of the Joint Symposium on Interaction of Electromagnetic Waves with Biological Systems, chaired and organized by James C. Lin of the University of Illinois-Chicago, held in Tel Aviv, Israel, during the XXIInd URSI General Assembly, in 1987. The presentations of the latter meeting were published in the book *Electromagnetic Interaction with Biological Systems*, edited by James Lin [1]. It is noteworthy that both symposia prominently featured presentations of research by scientists from around the world, working in this interdisciplinary field.

Following the General Assembly in Tel Aviv, to further URSI's role in fostering international forums for cooperation and discussion in this emerging field, the Inter-Commission Working Group (Chair: Saul Rosenthal; Vice Chair: Paolo Bernardi), was given the mandate to examine the possibility of creating, within the URSI framework, an ad hoc Commission devoted to the interaction of electromagnetic waves with biological systems. Unfortunately, in the interim, Saul Rosenthal died before the next General Assembly.

2. The Formation of URSI Commission K

During the XXIIIrd General Assembly in Prague, Czechoslovakia (now in the Czech Republic), in 1990, an ad hoc group was appointed by Jørgen Bach Andersen. As Vice President of URSI at the time, he served as Chair of the group, with three members: Paolo Bernardi, James Lin, and Maria Stuchly from the University of Victoria in Canada. The general consensus was that the time had come to explore the formation of a new Commission, and to enlarge URSI's involvement in this emerging interdisciplinary field of science. The group convened an open business meeting to present its recommendations; about 50 delegates were in attendance [2]. Following discussions, the delegates strongly endorsed the formation of a new Commission, with a suggested name of Electromagnetics in Biology and Medicine. Subsequently, the URSI Council decided to create Commission K during the Prague General Assembly [3].

The decision took into consideration the increasing interest and activities in the field of bioelectromagnetics and its interdisciplinary character, and the need for a truly international forum where biologists, engineers, physicians, and physicists would be able to interact. The role of URSI in establishing such an international forum had already been recognized by the community involved, The scope of the Inter-commission Working Group set up to cover the metrology aspects of bioelectromagnetics had become much wider over the years. Moreover, several URSI Commissions could contribute usefully to the development of bioelectromagnetics by bringing together diverse expertise in the engineering, mathematical, and physical fields.

The resolution to create the new Commission K on Electromagnetics in Biology and Medicine suggested that the precise terms of reference for the new Commission K would be defined later. It should include the study of interactions between electromagnetic radiation and living systems in the entire electromagnetic frequency spectrum (namely, dc to optical frequencies), and applications in medicine. It also designated Jørgen Bach Andersen (Denmark) as Interim Chair of the new Commission, and Maria Stuchly (Canada) as Interim Vice Chair. Later, balloting was held within the ranks of the official members of the new Commission K. Maria Stuchly and Paolo Bernardi were elected as the first Chair and Vice Chair of Commission K for a term ending at the 1993 General Assembly in Kyoto, Japan [4]. The URSI Scientific Commission K on Electromagnetics in Biology and Medicine was thus formally launched, and became operational at the beginning of 1991.

The initial charges and functions (terms of reference) of Commission K were the study of interactions of electromagnetic radiation with living systems from dc to optical frequencies and their applications in medicine, and promotion of research and development in the domains of (a) physical interactions of electromagnetic fields with biological systems; (b) biological effects of electromagnetic fields; (c) interaction mechanisms; (d) human exposure assessment; (e) experimental exposure systems; and (f) medical applications [5]. Moreover, a particular attribute of Commission K was its interdisciplinary character, which enabled it to foster research cooperation among various disciplines, and to interact with fellow URSI Commissions and other relevant organizations in its domains of interest.

It should be noted that during the XXVth General Assembly, Commission K made changes to its terms of reference to further refine and reflect developments in the subjects of interest to the Commission [6]. In particular, *interaction mechanisms* was changed into *mechanisms of the effects*, and the term *medical applications* became *medical and biological applications*. Finally, a new term of reference was added: *electromagnetic field interference with medical devices either implanted or connected to the human body*.

3. Activities During Commission K's First Decade

The new Commission K was well positioned to face the challenges bestowed upon it. This section highlights activities that took place during the first decade, including the 1999 General Assembly in Toronto, Canada, and encompassing three General Assemblies. During this period, the Chair and Vice Chair of Commission K and their terms in office were

1991-1993: Maria Stuchly (Canada) and Paolo Bernardi (Italy)

1993-1996: Paolo Bernardi (Italy) and James Lin (USA)

1996-1999: James Lin (USA) and Shoogo Ueno (Japan)

3.1 Participation in Triennial General Assemblies

A major function of the Commission is the development of an international scientific program at the triennial General Assembly.

The XXIVth URSI General Assembly, hosted by the historic capital city of Japan, Kyoto, in 1993, was the inaugural event of Commission K. The scientific program included an invited tutorial lecture by Ross Adey on the interaction of electromagnetic fields with living systems and medical applications. Commission K, led by its Chair, Maria Stuchly, organized four platform and one poster sessions, and two joint sessions. A total of 88 out of 117 submitted papers were scheduled for presentation [7]. A joint session devoted to "Exposure Assessment and Measurement in Complex Environments" was convened by Lars-Erik Paulsson from Sweden (Commission K) and Moto Kanda from USA (Commission A). It included such topics as measurement of contact current and transient fields, exposure assessment for mobile communication, and epidemiological studies at power-line frequencies. The other joint session, entitled "Computational Electromagnetics in Biology and Medicine," was convened by James Lin of USA (Commission K) and S. N. Horsleth of Denmark (Commission B). Topics focusing on analytic and numerical techniques used for computing energy deposition in biological tissues at low and high frequencies from various environmental, occupational, and medical exposures were the subjects of discussion.

The titles and conveners of the four platform sessions were: "Interaction Mechanisms of Electromagnetic Fields (T. Tenforde, USA); "Health Effects" (R. Saunder, UK, and D. J. Szabo, Hungary); "Electromagnetic Fields in Medical Diagnosis" (M. Saito, Japan, and G. J. Beers, USA); and "Therapeutic Applications of Electromagnetic Fields (B. Veyret, France, and C. K. Chou, USA).

The unprecedented growth and spread of wireless communication services brought into the foreground two main concerns related to the subjects of interest of Commission K. The first concern dealt with questions on the safety of human exposure to the electromagnetic fields used in wireless communications, and, as an extension, to low-frequency fields. In fact, the growth of wireless communications allowed large numbers of people on the move to be constantly linked to the rest of the world. At the same time, the increasing presence of electromagnetic fields in the environment raised public concerns about possible health effects of these fields. The second concern, also linked to the increasing presence of electromagnetic fields in the environment, was the possible malfunctioning of medical devices due to interference problems. These concerns, and the research and development that went on during the triennium, played an important role in the organization of scientific programs for the next two URSI General Assemblies.

The scientific program of Commission K at the XXVth URSI General Assembly in Lille, France, in 1996, was organized by the Commission Chair, Paolo Bernardi. It consisted of a general lecture on "Mobile Communication Systems and Biological Effects on their Users" by Maria Stuchly; and four sessions organized by Commission K and three sessions organized jointly with Commissions A, B, and E. Commission K led the organization of the first two joint sessions [6]. The four Commission K sessions consisted of 14 invited presentations, 25 contributed papers, and 29 poster papers. The topics were: "Biological Effects and Mechanisms of Interaction," "Safety of ELF and LF Fields," "Safety of Wireless Communications," and "Medical Applications of EM Waves." The joint sessions had 17 invited presentations, 10 contributed papers, and two posters on the topics of human-exposure assessment and related measurements, EM modeling in bioelectromagnetics, and characterization of EM sources and design to minimize coupling to the human body. In addition, Commission K was the sponsor of a tutorial lecture, attended by an estimated 600 participants. The number of participants at each session varied somewhat, but was estimated to range from 60 to 140. Clearly, the sessions were well attended, and the participants were enthusiastic.

The titles and convener(s) of the four Commission K organized scientific sessions were:

K1 "Biological Effects and Mechanisms of Interaction
(A. Chiabrera, Italy, and B. Veyret, France)

K2 "Safety of ELF and LF Applications"
(C. Polk, USA, and L. D. Szabo, Hungary)

K3 "Safety of Wireless Communication"
(N. Kuster, Switzerland, and J. C. Lin, USA)

K4 "Medical Applications of EM Waves"
(K. M. Reineck and S. Ueno, Japan)

The titles and convener(s) of the three jointly organized scientific sessions were:

KA "Human Exposure Assessment and Related Measurements"
(L. E. Paulsson, K-Sweden, and S. Tofani, A-Italy)

KB "Electromagnetic Modeling in Bioelectromagnetics"
(O. P. Gandhi, K-USA, and P. Excel, B-UK)

EK "Characterization of EM Sources and Design of Equipment for Minimum Coupling with the Human Body"
(R. De Leo, E-Italy, and H. Korniewicz, Poland)

As in previous triennia, the XXVIth General Assembly, held in Toronto in 1999, was organized by the Commission K Chair, James Lin. It scheduled a record number of 137 platform and poster presentations [8], including one tutorial lecture, and 78 platform and 58 poster presentations, divided into eight oral sessions and five poster sessions. Four of the sessions were organized jointly with Commissions A, B, C, and E, expanding Commission K's cooperation further to include four other Commissions. The number of participants at each session varied somewhat, but was very high. It was estimated to range from 60 to 140. Clearly, the sessions were well attended, and the participants enthusiastic. A brief listing of the program is given below.

Tutorial Lecture: "The Biology and Epidemiology of Human Exposure to Power-Line Electromagnetic Fields," by Russel Reiter, USA

Platform Sessions (and conveners):

K1 "Mechanisms and Modeling of Electromagnetic Interaction with Biological Systems"
(C. Polk, USA, and G. D'Inzeo, Italy)

K2 "Biological Effects of Electromagnetics Fields"
(L. Kheifets, USA, and R. Korenstein, Israel)

K3 "Hazard Assessment for Wireless Communication"
(P. Bernardi, Italy, and B. Veyret, France)

K4 "Biomedical Applications of Electromagnetic Fields and Waves"
(C. Gabriel, UK, and S. Ueno, Japan)

Joint Sessions:

KA "Exposure Assessment for Cellular and Personal Telecommunications"
(C. Chou, K-USA, and M. Taki, A-Japan)

KB "Computation of Electromagnetic Fields in the Human Body"
"O. Gandhi, K-USA, and Y. Rahmat-Samii, B-USA)

KC "Health Effects of Mobile Telephones"
(R. Adey, K-USA, and N. Kuster, C-Switzerland)

KE "Electromagnetic Interference with Medical Devices"
(D. Witters, K-USA, and O. Fujiwara, E-Japan)

It is noteworthy that during the 1993-1996 triennium, URSI celebrated its 75th anniversary, in April, 1995. As part of the celebrations held in Brussels, Belgium, in the presence of HM the King of Belgium Albert II, Maria Stuchly gave one of the official presentations. It was entitled "Bioelectromagnetic Effects on Users of Mobile Communication Systems," and was proposed by Commission K.

3.2 International Scientific Symposia

Following the practice of other Commissions in organizing scientific meetings between two General Assemblies, the first inter-assembly meeting, Microwaves in Medicine, was organized by the Chair of Commission K (Paolo Bernardi) in Rome, Italy, in October 1993. The meeting was organized in cooperation with the IEEE Microwave Theory and Techniques Society (MTT-S), and co-sponsored by the European Bioelectromagnetics Association (EBEA) and the Italian Society for Radiation Research (SIRR). The symposium was opened by an invited lecture on "Minimally Invasive Microwave Technology for Therapeutic Medicine" by James C. Lin, President-elect of the Bioelectromagnetics Society, followed by 11 scientific sessions with a total of 69 presentations [9].

The Chair of Commission K (James Lin, 1996-1999), with the cooperation of the IEEE MTT-S, organized the subsequent inter-assembly meeting. This was the International Symposium on Electromagnetics in Medicine, held in Chicago, USA, in November 1997 [10]. After observing the technology changes and receiving much feedback from attendees of the Rome meeting and the XXVth URSI General Assembly, it was decided to expand the meeting to cover both low and high frequencies, instead of only microwaves. To reflect this expansion of the meeting's scope, the name of the scientific meeting was changed. The meeting had a total of 76 attendees, which included 21 students. The quality of presentations and level of discussions were very high.

The technical program covered three full days. A single-session format was adopted to encourage interdisciplinary dialogue. Sessions began on Monday morning and ran through Wednesday afternoon. An invited lecture started the meeting each morning, on a topic of emerging interest. It was followed by sessions each having six contributed papers.

3.3 Cooperation and Co-Sponsorship of Meetings with Other Organizations

From the start, Commission K actively participated in and co-sponsored international conferences with other relevant organizations. Indeed, one of the most rewarding undertakings of Commission K during this period was the co-sponsorship of national and international conferences and symposia with technical support or, in some cases, with both financial and technical support. A chronological listing of these, with references to articles reporting on them in the *Radio Science Bulletin* where appropriate, is given below.

In June 1992, Commission K co-sponsored sessions at the First World Congress for Electricity and Magnetism in Biology and Medicine, held in Orlando, FL, USA. The Congress was organized by the Bioelectroamgnetics Society, Bioelectrical Repair and Growth Society, Bioelectrochemical Society, European Bioelectromagnetics Association (EBEA), and IEEE Engineering in Medicine and Biology and Power Electronics Societies [11]. The session topics ranged from biological effects of electromagnetic fields, interaction mechanisms, dosimetric modeling, and exposure assessment to molecular devices and medical applications. The congress programs consisted of plenary and tutorial sessions in the morning, with five parallel platform sessions, poster sessions, and posters in the afternoon.

During the 1993-1996 triennium, Commission K agreed to collaborate with other groups for the development of common scientific activities and organization of special sessions at their annual meetings. Commission K organized various special sessions on subjects close to the terms of reference of the Commission at many international conferences and symposia, including the following (in what follows, a single asterisk indicates technical support provided by Commission K; a double asterisk indicates both technical and financial support):

- 2nd EBEA Congress in Rome, Italy, in 1993*
- BEMS annual meeting in Copenhagen, Denmark, in 1994*
- Physics and Engineering of Millimeter and Sub-Millimeter EM Waves in Kharkov, Ukraine, in 1994**
- 3rd EBEA Congress in Nancy (France) in 1995.* URSI Commission K joined with EBEA and European COST Project 244 in organizing two special sessions on "Human Exposure in Mobile Communication Systems" and "Biological Responses to ELF Fields."
- International Symposium on Electromagnetic Compatibility (EMC) in Rome, Italy, in 1994*
- EMC International Symposium in Zurich, Switzerland, in 1995*
- URSI Commsphere in Eilat, Israel, in 1995*
- International Symposium on EMC Rome, Italy, in 1996*
- COST 244 Workshop on "Biological Effects Relevant to Amplitude Modulated RF Fields" in Kuopio, Finland, in 1996**
- Congress on "Biophysical Aspects of Coherence" in Prague, Czech Republic, in 1996**
- International EMC Conference in Wroclaw, Poland, in 1996**

Between 1997 and 1999, Commission K co-sponsored 10 conferences and symposiums: six meetings with no financial support, and five meetings with financial and technical support. A chronological listing of these, with references to articles reporting on them in the *Radio Science Bulletin* where appropriate, is given below:

- URSI Commsphere, Lausanne, Switzerland, February 11-14, 1997 [12]**
- 2nd World Congress for Electricity and Magnetism in Biology and Medicine, Bologna, Italy, June 11-13, 1997 [13]
- Radio Africa, Nairobi, Kenya, August 4-8, 1997 [14]**
- Symposium on Electromagnetic Fields in Biological Systems, Prague, Czech Republic, September 13-16, 1997 [15]**
- EMC Roma, Rome, Italy, September 14-18, 1998 [16]**
- Physics and Engineering of MM and Sub-MM EM Waves, Kharkov, Ukraine, September 15-17, 1998 [17]**
- 4th EBEA Congress, Zagreb, Croatia, November 19-21, 1998 [18]**
- Asian Pacific Microwave Conference, Yokohama, Japan, December 8-11, 1998
- URSI Commsphere, Toulouse, France, January 25-28, 1999**
- Day on Diffraction, St Petersburg, Russia, June 1-4, 1999

3.4 *Review of Radio Science*

The *Review of Radio Science* was first published in 1993 as an edited volume with peer-reviewed chapters summarizing the status and future of radio science, both for those working in the field and those who wished to know what was of current interest. The book was organized into sections by Commission, and the topics, author, and referees were selected by the Commissions. Prior to 1993, it was published triennially at the time of the General Assembly of URSI in the form of only a selected, annotated bibliography. With the new book format, a computer diskette with the bibliographic database was provided with the volume of reviews. By design, the Vice Chair of the Commission usually served as the section editor.

For the 1993 General Assembly in Kyoto, Commission K, under the leadership of Paolo Bernardi, contributed 100 pages in five chapters:

- "Biophysical Mechanisms of Interaction" by S. F. Cleary
- "Biological Effects of Electromagnetic Fields and Radiation" by Z. J. Sienkiewicz, N. A. Cridland, C. I. Kowalczuk, and R. D. Saunders
- "Diagnostic Applications of Electromagnetic Fields" by J. C. Lin
- "Therapeutic Applications of Electromagnetic Fields" by J. W. Hand and R. Cadossi
- "Human Exposure Assessment and Dosimetry" by L. E. Paulsson.

Commission K's bibliography diskette was prepared by P. Bernardi with help from others [19].

For the 1996 General Assembly in Lille, the Commission K chapters for the book *Review of Radio Science, 1993-1996*, reviewing the developments of radio science related to the triennium, were edited by J. C. Lin [20]. The book contained two Commission K chapters:

- "Medical Applications of Electromagnetic Energy" by C. K. Chou and W. W. Wang
- "Biological Effects of Electromagnetic Fields" by R. de Seze and B. Veyret

The Commission K part of the disk of references was edited by S. Pisa, using contributions provided by the Commission K official Members, together with references selected from international journals and publications. It consisted of more than 500 references, each quoted with some keywords. Although the disk was not intended to be an exhaustive bibliography for the years covered, it gave a broad representation of the international literature published in scientific journals, books, or as full papers in the proceedings of international symposia.

The Commission K portion of the 1999 triennial URSI *Review of Radio Science, 1996-1999*, was edited by S. Ueno. There were five chapters in the Commission K section [21]. The titles and the authors of these chapters were:

- "Cell and Molecular Biology Associated with Radiation Fields of Mobile Telephones" by W. R. Adey
- "Review of Exposure Assessment for Handheld Mobile Communications Devices and Antenna Studies for Optimized Performance" by M. Burkhardt and N. Kuster
- "Bioelectric and Biomagnetic Measurements" by T. Katila and R. J. Ilmoniemi
- "Medical Applications of Static and Low-Frequency Pulsed Magnetic Fields" by S. Ueno
- "Biomedical Applications of Electromagnetic Fields and Waves: Radio Frequencies and Microwaves" by J. C. Lin

Commission K's diskette of references of archival publications during 1996-99 was edited by Masao Taki. The final list of 643 references included in the K section consisted of suggestions from official members, and references selected from journals and other archival publications. Although the disk was not intended to be an exhaustive bibliography for the years covered, it gave a broad representation of the international literature published during the triennium.

The Commission K section of the *Review of Radio Science, 1999-2002*, was edited by Bernard Veyret and Niels Kuster. There were two chapters in the Commission K section [22]. The titles and the authors of these chapters were:

- "Possible Exposures from Future Mobile Communication Systems" by J. Bach Andersen, P. Mogensen, and G. F. Pedersen
- "Biological Effects of Microwaves: Animal Studies" by Zenon Sienkiewicz

From 2002 forward, the *Review of Radio Science* papers have been published in the *Radio Science Bulletin*.

3.5 URSI Resolutions and Recommendations

Commission K recognized the problem, the scientific uncertainty, and the public concern about health effects of all RF systems. It recommended and caused the URSI Council's adoption of specific resolutions throughout the first decade. The objectives of the resolutions were to stimulate domestic research, encourage international cooperation, and to provide coordination, if necessary.

At the XXIVth General Assembly in Kyoto, considering (1) that there was a rapid development of new technologies such as local-area networks (LAN), cellular telephones, personal communication services (PCS), cordless telephones, etc., and their expansion was anticipated; (2) that there existed scientific uncertainties about the potential impact on human health of electromagnetic fields for wireless communication; and (3) that there was public concern about health effects of all electromagnetic systems, Commission K recommended [5] that broadly based research programs should be established nationally and internationally to address the following key questions: (1) what are the interaction mechanisms with living systems of weak electromagnetic fields of various characteristics; (2) what biological effects – and particularly potentially harmful effects – are caused, and under what exposure conditions; and (3) how to evaluate the exposure through proper measurements and dosimetric modeling.

Commission K was instrumental in the URSI Council's adoption of a resolution to establish an Inter-Commission Working Group on the Safety of Medical Devices in the Presence of Electromagnetic Fields, during the XXVth General Assembly in Lille, France [6]. The resolution noted that there was increasing evidence that electromagnetic fields from wireless communication devices may affect the operation of some medical devices – either implanted or connected to the human body – and as a result may pose a problem to operation and health. The Working Group consisted of representatives from Commissions K and E. It aimed to study the specific behavior of implanted equipment; the characteristics of connected medical equipment; modeling methods; specific measurements; and the influence of the person on electromagnetic interference.

During the 1996-99 triennium, Commission K recommended accelerated scientific and industrial research to ensure the safety of medical devices in the presence of electromagnetic fields. Specifically, studies were to be aimed at clarifying (a) the specific behavior of implanted equipment; (b) the characteristics of connected medical equipment; (c) modeling methods; (d) specific measurements; and (e) influence of the person on electromagnetic interference (EMI). To ensure that these global research programs responded to local environmental concerns and societal goals, URSI Commission K encouraged each member Committee to actively promote research in their territory and to urge appropriate authorities to create research centers devoted to electromagnetics in biology and medicine. Moreover, Commission K sponsored the URSI Council's Resolution [8] that stated the following:

Recognizing:

a) That all lives on Earth thrive in a natural electromagnetic environment. Over the past few decades, we have learned to understand some of its characteristics and we have applied them in abundant ways to embellish our lives. Indeed, we have come to depend on the electromagnetic environment for life, health, safety, information, comfort, and conveyance.

b) Bioelectromagnetic research has developed a unique body of new knowledge and it is crossing a threshold from the traditional boundaries of biological and biophysical sensitivities. This new knowledge provides an invaluable bridge between health hazards of exposures to electromagnetic fields and waves and new diagnostic and therapeutic uses of electromagnetic fields and waves.

c) As scientific understanding of the interaction of electromagnetic interaction with biological systems increases, the prospect for its use in biology and medicine also becomes greater;

Resolves

that URSI Member Committees encourage appropriate international and national organizations to promote research on the effects of electromagnetic fields and waves in biology, and their uses in diagnostic and therapeutic medicine, for the benefit of human society.

These recommendations and resolutions have been significant in the development of the scientific discipline of bioelectromagnetics in many countries. Very few countries have the resources capable of sustaining a large-scale research program, let alone linking it to the rest of the world. As an international organization with a long-standing interest, URSI has been instrumental in fostering the development of this scientific area.

4. Conclusions

As an international organization with a long-standing interest, URSI has fostered the development of the scientific area of "Electromagnetics in Biology and Medicine" for many years. It has provided a platform for cooperative research, exchange of data, and the sharing of knowledge. Through this process, the research efforts of national Commissions were connected to the global effort by URSI Commission K, both to facilitate and provide knowledgeable scientific expertise in each member Committee, coupled with the-state-of-knowledge in the global community.

As we look forward to the next General Assembly in Istanbul, Turkey, it is reassuring that the Commission created in 1990 to deal with the subject of electromagnetics in biology and medicine has had four more extremely successful sessions at the XXVth, XXVIth, XXVIIth, XXVIIIth, and XXIXth General Assemblies in 1996, 1999, 2002, 2005 and 2008, respectively, in Lille, France; Toronto, Canada; Maarstricht, the Netherlands; New Delhi, India; and Chicago, USA.

It is evident that Commission K on Electromagnetics in Biology and Medicine – although the youngest among URSI Commissions – has played a pioneering role. It was a significant player in the development of the scientific discipline of bioelectromagnetics, with a focus on the interaction of electromagnetic fields with biological systems. Commission K will soon be celebrating its 20th anniversary. It is our sincere belief that it will continue to succeed in organizing meetings and sessions to discuss scientific ideas, exchange knowledge, and debate issues of interest to the Commission in years and General Assemblies to come, including the XXXth General Assembly and Scientific Symposium in Istanbul, Turkey, during 2011.

5. Acknowledgment

The authors thank URSI Executive Secretary, Mrs. Inge Heleu, Belgium, for her assistance in providing some of the records used in the preparation of this article.

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Conferences

CONFERENCE REPORT

IHY-AFRICA/SCINDA 2009

Livingstone, Zambia, 7 – 12 June 2009

The workshop was hosted by the University of Zambia in collaboration with Hermanus Magnetic Observatory, South Africa, the North-West University, South Africa, and the Air Force Research Laboratory, USA. It was sponsored by the National Aeronautics and Space Administration (NASA), Greenbelt, Maryland, USA, the Air Force Research Laboratory (AFRL), Hanscom Air Force Base, Massachusetts, USA, the National Research Foundation (NRF), Pretoria, South Africa, the European Office of Aerospace Research and Development (EOARD), London, UK, the National Science Foundation (NSF), Arlington, Virginia, USA, the International Center for Theoretical Physics (ICTP), Trieste, Italy, Climate and Weather of the Sun-Earth System (CAWSES)/International Committee on Solar-Terrestrial Physics (SCOSTEP), the International Union of Radio Science (URSI) Commission G and H, Belgium, the University of Colorado, Cooperative Institute for Research in Environmental Sciences (CIRES), Boulder, Colorado, USA, MTN Mobile Solutions, Pretoria, South African, the North-West University (NWU), Potchefstroom, South Africa, Hermanus Magnetic Observatory (HMO), Hermanus, South Africa, Rhodes University, Grahamstown, South Africa, the University of Zambia (UNZ), Lusaka, Zambia, the National Science and Technology Council (NSTC),Lusaka, Zambia and North Carolina A&T State University, North Carolina, USA.

Introduction

As a continuation of the very successful International Heliophysical Year (IHY)-Africa 2007 workshop, which was held in Addis Ababa, Ethiopia in November 2007, the IHY-Africa 2009 workshop was organized in Livingstone, Zambia during the period 7 – 12 June 2009. The purpose of the workshop was to facilitate scientific interaction between space science instrument donors and host institutes and promote space science in Africa, with a strong focus on education and public outreach. Additional aims of the workshop were to converge African and other international scientists under one roof, facilitate the deployment of new observational infrastructure to study space weather, spark interest in space science education and research, and encourage the next generation of African scientists to become interested in the space sciences. The workshop was also the ideal forum to review and document the ongoing instrument deployment and data retrieval progress during the IHY period. This workshop also marked the end of the IHY period and the beginning of the International Space Weather Initiative (ISWI) which will take over the IHY work.

The workshop was aimed at scientists and students from African countries, and those international scientists interested in working together with African scientists, who are involved in all aspects of Space Physics, and included four main science sessions, two major poster sessions and four discussion forums.

The workshop was held in conjunction with the Scintillation Network Decision Aid (SCINDA) 2009 meeting and sessions of particular interest to SCINDA were held on the first two days of the meeting.

Attendees and Logistics

The workshop was organized through three committees, the International Advisory Committee, the International Organising and Programme Committee, and the Local Organising Committee (LOC). A list of the members of these committees appears in Appendix A.

A total of 116 delegates attended the workshop in Zambia. Of these delegates, 79 represented 19 different African countries and 37 travelled from 8 different countries outside of Africa. The list of countries represented at the workshop is given in Appendix B, and a complete list of delegates can be found in Appendix C. What is worth mentioning here is that 24 of the delegates were postgraduate students and 10 were undergraduate students from the University of Zambia.

Figure 1: Prof Nat Gopalswarmy (left) from NASA in the USA in discussions with Jean Uwamahoro, a young Rwandian student.

Due to the generous sponsorship that the workshop received, 74 African delegates and 13 rest of world delegates were sponsored either partially or in full. Travel arrangements were made for the selected sponsored African scientists using a travel agency in South Africa who made all the flight arrangements for the sponsored delegates. The same travel agency also assisted when changes were needed to flights during the workshop. The travel arrangements for the sponsored US scientists were made in the US through UCLA, and were coordinated by Dr Endawoke Yizengaw. Travel arrangements for the Zambian based participants were coordinated by UNZA staff.

The workshop was held in the New Fairmount Hotel (NFH) in Livingstone. Of the 116 delegates, 110 were accommodated in the NFH with 6 choosing to make their own arrangements. The LOC coordinated all the accommodation arrangements with the NFH, who also arranged all the meals and tea breaks.

The opening ceremony was held on the afternoon of Monday 8 June 2009, and included presentations in the conference hall followed by a sunset cruise on the Zambezi River. The opening ceremony was chaired by Prof Roger Smith, Director of Geophysical Institute of University of Alaska and was attended by the Dean of the School of Natural Sciences of the University of Zambia, Dr Samuel Banda, the Vice Chancellor of the University of Zambia, Prof Stephen Simukanga, and the Honourable Zambian Minister of Science and Technology, Honourable Gabriel Namulambe. Presentations were made by Dr Joseph Davila, the International Heliophysical Year Secretariat as well as a few representatives from the sponsors. The three Zambian guests of honour also made presentations which were motivating for science in Zambia and illustrated the Zambian government's commitment to science and technology and their appreciation for the presence of the IHY-Africa/ SCINDA 2009 workshop in Livingstone.

Scientific Programme

The first one and a half days (Sunday and Monday

Figure 2: Mpho Tshisaphungo, a South African student, and Daniel Okoh, a Nigerian student, with Prof Sandro Radicello (center) from ICTP in Italy.

Figure 3: Group picture taken at the workshop

morning) were dedicated to the SCINDA workshop presentations. Monday afternoon was dedicated to the opening ceremony. The scientific programme of the workshop included four main Heliophysics science sessions, as follows with convener names:

- Total Electron Content - Tim Fuller-Rowell and Christine Amory-Mauzaudier
- Irregularities and scintillations – Lee-Anne McKinnell and Keith Groves
- Sun-Heliosphere connection – Marius Potgieter
- Magnetosphere-Ionosphere coupling - Endawoke Yizengaw and Vafi Doumouya

These four science sessions included 40 invited oral presentations with the main focus on space science over the African region.

In addition there were 2 poster sessions which were convened by Pierre Cilliers and chaired by postgraduate students. A total of 46 posters were presented with 23 in each session. The two poster sessions were divided according to the same main science sessions mentioned above with the one exception that a poster session with a general theme was included so that poster abstracts that did not fit into one of the main sessions could also be included.

Included in the program were four one hour open discussion forums, as follows with convener names:

- Infrastructure and Communications in Africa – Victor Chukwuma and Daniel Moeketsi
- Science Education and Outreach – Carolyn Sumners and Patricia Reiff
- Instruments and Databases – Christine Amory-Mazaudier and Justin Mabie
- Establishment of African Geophysical Society – Babatunde Rabiu and Baylie Damtie.

Outreach

The IHY-Africa/SCINDA 2009 workshop had a large outreach and public awareness component. On the Saturday prior to the workshop a NASA Geophysical Information for Teachers (GIFT) workshop was held at the New Fairmount Hotel in Livingstone, Zambia. The workshop was entitled "Sun-Earth Relationships and the International Heliophysical Year" and was attended by 22 local school teachers. Several teaching aid materials were given to those teachers who participated in the GIFT workshop and these will help them in their teaching and learning processes within their classroom. The workshop program and organizing committees are listed in Appendix E.

In addition one of the local high schools, Hillcrest National Technical High School, brought approximately 30 learners to each poster session evening. These learners were introduced to the international scientists who spent time with them, and they had access to a portable planetarium which was installed in the venue throughout the week especially for the workshop. The planetarium, from Rice University, was brought by the GIFT workshop organizers for a GIFT workshop demonstration. At the end of the workshop the planetarium was donated to the University of Zambia, which makes Zambia the second country, next to Ethiopia, in Africa that has a portable scientific planetarium dome. Several of the scientists and students attending the workshop also visited Hillcrest during the week to give presentations and talk to the learners.

Conclusion

The workshop was successful in meeting the aim of bringing together African Space Scientists and providing a forum in which they could meet and make contacts with international colleagues in the field (Figures 1 and 2 are examples). A number of collaborations have been established as a result of the workshop, and a number of students were exposed to the Space Science field and to successful African scientists in the field. The problems of database access and infrastructure requirements were also discussed, and scientists with instrumentation had access to expertise with whom they could discuss problems. Meetings like these form a very valuable part of scientific research and should be encouraged and supported.

This workshop would not have been possible without the support of the sponsors who are listed at the front of this report, and would not have been successful without the support and hard work of the Zambian and South African organizing team.

CONFERENCE ANNOUNCEMENTS

Interantional conference on Electromagnetics in Advanced Applications (ICEAA 2010 - Off-Shore)

Sydney, Australia, 20 – 24 September 2010

Scope

For the first time in its history, the International Conference on Electromagnetics in Advanced Applications (ICEAA) is moving off the shores of Italy to Sydney, Australia. The twelfth edition of the International Conference on Electromagnetics in Advanced Applications (ICEAA 2010) will consist of invited and contributed papers, as well as workshops and short courses. The conference is supported by the Politecnico di Torino and by the Istituto Superiore Mario Boella, with the principal technical co-sponsorship of the IEEE Antennas and Propagation Society and the technical co-sponsorship of the International Union of Radio Science.

Topics

1. Active and smart antennas
2. Electromagnetic applications to biomedicine
3. Electromagnetic applications to nanotechnology
4. Electromagnetic measurements
5. Electromagnetic modeling of devices and circuits
6. Electromagnetic packaging
7. Electromagnetic properties of materials
8. EMC/EM1/EMP
9. Finite methods
10. Frequency-selective surfaces
11. Integral equation methods
12. Intentional EMI
13. Inverse scattering and remote sensing

14. Metamaterials
15. Microwave antennas
16. MIMO antenna systems
17. Optoelectronics and photonics
18. Phased and adaptive arrays
19. Plasma and plasma-wave interaction
20. Printed and conformal antennas
21. Radar cross section and asymptotic techniques
22. Radar imaging
23. Radomes
24. Random and nonlinear electromagnetics
25. Statistics in electromagnetics
26. Technologies for mm and sub-mm waves
27. Wireless communications

Deadlines and Contact

The deadline for abstract submission is **February 26, 2010**. Inquiries should be directed to

Prof. Roberto D. Graglia
Chair of ICEAA Organizing Committee
Dipartimento di Elettronica Politecnico di Torino
Corso Duca degli Abruzzi, 24 10129 Torino, Italy
E-mail: roberto.graglia@polito.it

Prof. Piergiorgio L. E. Uslenghi
Chair of ICEAA Scientific Committee
Department of ECE (MC 154)
University of Illinois at Chicago
851 South Morgan Street, Chicago, IL 60607-7053, USA
E-mail: uslenghi@uic.edu

Prof. Paul D. Smith
Chair of ICEAA 2010
Local Organizing Committee
Department of Mathematics, Building E7A
Macquarie University, Sydney, NSW 2109, Australia
E-mail: pdsmith@maths.mq.edu.au

Current information about the conference can be found at http://www.iceaa-offshore.org.

10th International Workshop on Finite Elements for Microwave Engineering

New England, USA, 12 – 13 October 2010

Scope

The International Workshop on Finite Elements for Microwave Engineering is a highly-focused biannual event. It provides an ideal meeting place for researchers who are active in either the theoretical development of finite-element methods or their application to radio-frequency and microwave-engineering problems. All presentations will be oral. The workshop will be held in the town of Meredith, New Hampshire, USA. The conference site, Mill-Falls Inns & Spa, is located on the beautiful shores of lake Winnipesauke, with scenic views of New England's fall foliage.

Topics

Technical areas covered by the Workshop include, but are not limited to:

- Adaptive methods
- Advanced FEM techniques: formulations, solvers, discrete representations
- Optimization techniques, parameter space sweep
- Time-domain FEM theory and applications
- Mathematical aspects of FEM
- FEM applications: antennas, materials and metamaterials, bio-electromagnetics, electromagnetic compatibility, circuits, and circuit boards
- Hybrid methods – theory and applications: FE-BI/FDTD, FE-FV/circuit simulators/high-frequency techniques, coupled physics FEM for RF-MEMS
- FEM applications: waveguides, components, active devices, lumped elements
- Multi-grid and domain-decomposition methods
- CAD/meshing advances and tools
- FEM applications in other disciplines

Organizers

The General Chairs are M. N. Vouvakis, University of Massachusetts, and D. Weile, University of Delaware. The Technical Program Committee is chaired by B. Shanker Michigan State University, and R. W. Kindt, US Naval Research Lab.

Information

Information on paper submission and registration is available at http://www.regonline.com/fem2010. The abstract submission deadline is **March 19, 2010**.

URSI CONFERENCE CALENDAR

URSI cannot be held responsible for any errors contained in this list of meetings.

October 2009

International Conference on Radar
Bordeaux, France, 12-16 October 2009
Contact: SEE / CONGRESS DEPARTMENT, Béatrice Valdayron - Valérie Alidor - Caroline Zago - Morgane Melou, Fax: + 33 (0)1 56 90 37 08, E-mail : radar2009@see.asso.fr , Web: http://www.radar2009.org

November 2009

IRI Workshop 2009
Kagoshima, Japan, 2-7 November 2009
E-mail: iri2009@ep.sci.hokudai.ac.jp, Web: http://www.ep.sci.hokudai.ac.jp/~iri2009/

ESWW6 - Sixth European Space Weather Week
Brugge, Belgium, 16-20 November 2009
Web: http://sidc.oma.be/esww6

December 2009

AEMC09 - IEEE Applied Electromagnetics Conference and URSI Commission B meeting
Kolkata, India, 14-16 December 2009

ICMARS 2009 - International Conference on Microwaves, Antenna, Propagation and Remote Sensing
Jodhpur, India, 19-21 December 2009
http://www.radioscience.org/ICMARS-2009FirstCircular.pdf

February 2010

META '10 - Second International Conference on Metamaterials, Photonic Crystals and Plasmonics
Cairo, Egypt, 22-25 February 2010
cf. Announcement in the Radio Science Bulletin of March 2009, p. 65.
Contact: Dr. Said Zouhdi, Laboratoire de Génie Electrique de paris, LGEP-Supélec, Plateau de Moulon, 91192 Gif-sur-Yvette Cedex, France, Fax+33 1 69 418318, E-mail: said.zouhdi@supelec.fr, Web: http://meta10.lgep.supelec.fr

March 2010

MicroRad 2010
Washington, DC, USA, 1-4 March 2010
Web: http://www.microrad2010.org/

8th International Nonlinear Waves Workshop
La Jolla, CA, USA, 1-5 March 2010
Contact: William E. Amatucci, Plasma Physics Division, Code 6755, Naval Research Laboratory, Washington, DC 20375, USA, Fax : 202-767-3553, E-mail : bill.amatucci@nrl.navy.mil

3rd Workshop on RFI Mitigation in Radio Astronomy
Groningen, The Netherlands, 29-31 March 2010
Contact: Prof. W.A. BAAN, Netherlands Foundation for Research, in Astronomy - Westerbork Observatory, P.O. Box 2, NL-7990 AA DWINGELOO, NETHERLANDS, Fax : +31 521-595 101, E-mail: baan@astron.nl

April 2009

AP-EMC 2010 - Asia-Pacific EMC Symposium
Beijing, China, 12-16 April 2010
Contact: Web: http://www.apemc2010.org

June 2010

CROWNCOM - 5th Int. Conference on Cognitive Radio Oriented Wireless Networks and Communications
Cannes, France, 16-18 June 2010
Contact : Jacques Palicot, SUPELEC, Avenue de la Bulaie, 53576 Cesson-Sévigné, France, Fax +33 299-844599, E-mail: jacques.palicot at supelec.fr

July 2010

SCOSTEP - STP12
Berlin, Germany, 12-16 July 2010
Contact: Prof. Dr. Franz-Josef Lübken, Leibniz Institute of Atmospheric Physics, Schloss-Straße 6, 18225 Kühlungsborn, Germany, Fax: +49-38293-6850, E-Mail: luebken@iap-kborn.de, http://www.iap-kborn.de/SCOSTEP2010/

COSPAR 2010 - 38th Scientific Assembly of the Committee on Space Research (COSPAR) and Associated Events
Bremen, Germany, 18 - 25 July 2010

cf. Announcement in the Radio Science Bulletin of December 2008, p. 73.
Contact: COSPAR Secretariat, c/o CNES, 2 place Maurice Quentin, 75039 Paris Cedex 01, France, Fax: +33 1 44 76 74 37, E-mail: cospar@cosparhq.cnes.fr, Web: http://www.cospar2010.org/ or http://www.cospar-assembly.org

August 2010

EMTS 2010 - International Symposium on Electromagnetic Theory (Commission B Open Symposium)
Berlin, Germany, 16-19 August 2010
Contact: EMTS 2010, Prof. Karl J. Langenberg, Universität Kassel, D-34109 Kassel, Germany, E-mail: info@emts2010.de , Web: http://www.emts2010.de

3rd International Communications in Underground and Confined Areas
Val d'Or, Québec, Canada, 23-25 August 2010
Contact: Paule Authier, Secrétaire, LRTCS-UQAT, Laboratoire de recherche Télébec en communications souterraines de l'UQAT, 450, 3e Avenue, local 103, Val-d'Or,Québec J9P 1S2, Fax: (1)(819) 874-7166, E-mail: lrcs@uqat.ca, Web: http://www.icwcuca.ca/welcome.asp

September 2010

ICEAA 2010 - International Conference on Electromagnetics in Advanced Applications
Sydney, Australia, 20-24 September 2010
cf. Announcement in the Radio Science Bulletin of September 2009, p. 62-63.
E-mail: Roberto.Graglia@polito.it, Web: http://www.iceaa-offshore.org/

AP-RASC - 2010 Asia-Pacific Radio Science Conference
Toyama, Japan, 22-26 September 2010
cf. Announcement in the Radio Science Bulletin of September 2009, p. .
Contact: Prof. K. Kobayashi, Vice President for International Affairs, Chuo University, 1-13-27 Kasuga, Bunkyo-ku, Tokyo 112-8551, JAPAN, Fax: +81-3-3817-1847, E-mail: kazuya@tamacc.chuo-u.ac.jp

October 2010

Commission F Microwave Signatures Symposium
Florence, Italy, 4-8 October 2010
Contact: Prof. P. Pampaloni, IFAC/CNR, Florence, Italy, Fax +390 555 226434, E-mail: p.pampaloni@ifac.cnr.it

News from the URSI Community

U.R.S.I.

NEWS FROM A MEMBER COMMITTEE

NIGERIA

2009 Annual Conference of the Nigerian Union of Radio Science

The Nigerian URSI Commitee will organise its Annual Conference of Nigerian Union of Radio Science, which will be held at Redeemer's University, km 46/48, Lagos/Ibadan Expressway, Redemption City, Mowe, Ogun State, Nigeria, from 2 tot 5 December 2009.

Topics

Theme : Satellite Technology and its Applications
Sub-Themes :

- GSM and Fixed Wireless Services
- Space Weather Effects and Risks in Satellite Communications
- Remote Sensing and Global Satellite Systems
- Military Communications
- Telemedicine and Health Implications

Call for Abstracts

We cordially invite academicians, scientists, engineers, and students from academia, national scientific institutes agencies and industry / Radio Science practitioners with experience and / or knowledge of the topics cover on the theme / sub-themes of the conference to submit 200 words abstract with 4-6 keywords electronically to: nurs2009@run.edu.ng.

More information can be found at http//www.run.edu.ng/nurs2009/index.html

BOOK PUBLISHED BY URSI RADIOSCIENTISTS

Tools of Radio Astronomy, Fifth Edition

by T. L. Wilson, K. Rohlfs and S. Hüttemeister, Berlin, Springer, 2009, ISBN 978-3-540-85121-9,515 pp, € 80.

The publication of a fifth edition of *Tools of Radio Astronomy* is in itself proof of the success of the book. Several significant changes with respect to earlier editions have been incorporated. This is reflected in the change in the order of the authors, and the addition of sets of problems for which an additional third author was mainly responsible. Most chapters have been adjusted to reflect both technological and astronomical developments. In this process, the authors have separated the stable basic theoretical material from the more practical aspects, which are continuously subject to change and expansion.

The authors believe that the book will be useful to communication engineers, a claim about which one can have some reservations. The chapters with basic electromagnetic, antenna, and signal processing theory (Chapters 2, 3, 4, and 6) are covered more completely in the standard books very likely already on the engineer's shelf. The second half of the book is devoted solely to astrophysical aspects of radio astronomy. In Chapters 5, 7, 8, and 9, the general observational techniques with single-aperture and interferometer systems are sufficiently described for the use of the radio astronomer. However, the aspects of antenna calibration with the aid of radio-astronomical methods, of particular interest to communication engineers, are more completely treated in other publications.

When I read the second edition of this book, I wondered why one would fill six pages with an analysis of the dipole, something found in any textbook on electromagnetics. Now with the re-emergence of dipole arrays in several new low-frequency radio telescopes (for instance, LOFAR and SKA), the presence of this material does not look so much out of place. The infinitesimally small dipole is used as the starting point for the derivation of the radiation integral of an aperture antenna. To quote Wolfgang Pauli, while "it is not even wrong," it is unusual.

In Chapter 3, a basic description of wave polarization is presented, up to the Stokes-parameter representation of quasi-monochromatic waves. Unfortunately, the more unified treatment in terms of coherency matrices is missing, and the important recent work by Hamaker, Bregman and Sault (*Astron. Astrophys. Supp.*, **117**, 1996) is not even mentioned.

Chapter 4, on receiver theory, introduces the subject well, despite some incomplete mathematics in the square-law-detector and receiver-sensitivity sections. The equation for the rms noise level of a Dicke-switched radiometer is missing a factor of $\sqrt{2}$, although the authors explain in the text that this multiplicative factor should be added. This is a highly confusing way to present this important formula. A user limiting himself to the numbers in the table will expect a factor of $\sqrt{2}$ better sensitivity than he or she achieves.

In Chapter 5, aspects of receiver systems are described with a view of the practical use of these. All currently used technologies, from bolometers, SIS-mixers and HEMT amplifiers, to analog and digital spectrometers and special pulsar systems, are mentioned.

At several places in Chapters 6 and 7, one notices the result of imperfect "cut and paste" from the earlier edition, in that sentences are incomplete, or just crooked. The two paragraphs of Section 6.4.4. are almost identical. It seems that the first paragraph from an earlier edition had been overlooked for removal. A similar occurrence can be found in Sections 9.3.4 and 9.3.5. The "Summary," Section 6.6, indicates how things can be done differently, rather than summarizing the chapter.

Section 7.4. "Practical Design of Parabolic Reflectors," is unsatisfactory. More than a page is used in discussing aspects of the Arecibo telescope, which has a spherical reflector, and is unique. The discussion of homology design is in error, in that it assumes that it can be applied only to circularly symmetrical reflectors. The choice of illustrations – the SOFIA airplane without telescope and Satellite Herschel – is odd. More interesting examples of recent telescopes would have been welcome.

Chapter 8, "Single Dish Observational Methods," contains a good description of the influence of the Earth's troposphere, as well as of calibration methods. In the reviewer's opinion, telescope calibration at mm wavelengths suffers from a lack of a generally accepted and well understood method. This book does not present one either, but limits itself to comparing two widely used methods. The treatment of the "confusion problem" is quite confusing, and a step backwards from the discussion in the second edition of the book.

The discussion of "Interferometers and Synthesis" in Chapter 9 is a useful introduction to the subject. There are some small lapses, such as the missing parts of Figure 9.11 and wrongly numbered figures. The description of the different aspects of observing with these systems is well done, although the reader will have to turn to the definitive book by Thompson, Moran, and Swenson for the details.

The second half of the book, Chapters 10 through 16, deals with the astrophysical aspects of radio astronomy. Emission mechanisms and examples of continuum radiation sources are treated in Chapters 10 and 11. The remaining chapters are devoted to all aspects of spectral-line radiation. Starting with a short chapter on fundamentals, spectral lines of neutral hydrogen (the 21-cm line), recombination lines of ionized atoms, and lines of molecules in interstellar space are presented. These offer an up-to-date summary of the current state, along with all essential formulae and expressions for the analysis of line observations.

It is clear that the authors are in full command of these astrophysical subjects. However, this is also reflected in the very compact presentation of the theoretical material. Here, the character of a "tool box" is most obvious. This is not a textbook with a complete, pedagogically justified development of the material. It provides complete information to the at least somewhat knowledgeable astronomer, who is faced with the interpretation of his or her observations. As the authors point out, deeper insight into the physics of the interstellar medium will require study of additional material, to which proper reference is made.

In this respect, the "radio" part of the book differs in character from the "astronomical" section. The extensive and rather basic treatment in Chapters 2, 3, and 6 could be shortened, with references to standard textbooks. The available space might advantageously be used for a more-detailed description of the practicalities of observations and data analysis: in other words, provide more tools. This would better match the structure of the second part of the book.

In their Preface, the authors admit that they have yielded to the "preference of the astronomical community to use their mixed set of units." Such a mixture of Gaussian cgs and SI units is unfortunate, and should be avoided. Fortunately, the authors always clearly indicate the units in the equations. The index could be improved, and a list of symbols and physical constants is sorely missed. Proofreading has not been performed with the required care, but most errors will not hinder the understanding of the reader.

In summary, this up-to-date version will be of invaluable use to the observational astronomer. It indeed provides a toolbox of essential material, in particular in the area of the analysis and interpretation of radio-astronomy observations. The book should be on the desk of every practicing radio astronomer, next to the classic *Radio Astronomy* by John Kraus.

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Future of Planet Earth

(seminar proceedings, 2009; available free of charge as a download from http://www.futurefoundation.org/; 28.8 MB PDF, 358 pp.)

The seminar "Future of Planet Earth" was jointly sponsored and conducted by the Foundation for the Future and UNESCO's Division of Ecological and Earth Sciences (SC/EES). The seminar was among the activities organized in the framework of the United Nations "International Year of Planet Earth" (IYPE), which focused on the importance of Earth sciences for society and human well-being. The intention behind the seminar was to provide a forum for dialogue amongst academics, scientists, conservationists, resource managers, and others dedicated to the wellbeing of the planet, to explore the human-environment relationship with significant implications for the future of planet Earth. The seminar was held at UNESCO in Paris, France, in June 2008. The goal of the seminar was to identify issues critical to the long-term future of planet Earth.

The Foundation for the Future was established with the mission of increasing and diffusing knowledge concerning the long-term future of humanity. Its founder and President, the Swiss-born Walter Kistler, is a recognized pioneer in the invention and development of high-performance instrumentation for aerospace as well as a number of other industries. He has played a key role in the creation of several high-technology companies. Mr. Kistler is the owner of more than 50 US and foreign patents, and the author of a number of papers published in scientific and trade journals.

Scholars from five continents participated in the seminar, apparently all distinguished experts in their fields. Specific countries represented were India (2), the UK (3), Australia (1), South Africa (1), Norway (1), Ghana (1), Spain (1), and the USA (9).

The seminar "Future of Planet Earth" considered the past, present, and long-term future of planet Earth from a variety of perspectives, most importantly the following six themes:

1. Changes and shifts that can be identified and observed.
2. Systems and networks that form the totality of the planet.
3. Planet Earth futures.
4. Resources and sustainability, including consideration of carrying capacity vis-à-vis sustainability.
5. Cultures, values, ethics, and morals that underlie human interactions with the planet and the consequences thereof.

6. Knowledge, science, and technology – impact and potential with respect to humankind's custody and care of the planet.

Two to three presentations were given on each of these themes, 13 in total. The proceedings contains the complete text of the presentations, and for nine of them, the corresponding *PowerPoint* vugraphs. Covered were the well-known topics such as climate change, disasters, Earth-ocean interaction, energy and power, education, food supply, pandemics, resource management, and sustainable development. Less popular themes, such as human behavioral change, species survival and evolution, and sustained decentralization were also covered. After each of the above themes, the complete discussion among the 19 participants is included in the book. The contributions are mostly well written and understandable for laymen. Some of the vugraphs are excellently designed. In the discussions, the presentations were commented on from a broad variety of aspects, reflecting the diversity of the participants' fields of expertise (anthropology, biology and biodiversity, ecology, economy, environmental policy, epidemiology, geology and geomorphology, heritage conservation, management, ocean and water science, philosophy, physics, science and technology policy).

Following the presentations is a detailed discussion on critical issues and the formulation of guidelines for the future (70 pages).

The book also contains the biographies of the participants, and a detailed index, which is very useful for finding information about a specific item.

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The SUMMA Graduate Fellowships in Advanced Electromagnetics

These fellowships are intended to promote exceptionally creative contributions to the advancement of electromagnetic theory and applications. They promote EM theory (conceptual, new ideas, new approaches), but with apparent potential applications (antennas, scattering, propagation, other EM devices). They will be awarded to graduate students studying electromagnetics for pursuit of the PhD. Such students may be currently MS or PhD candidates. It is expected that these fellowships will be of primary interest to students in electrical engineering specializing in electromagnetics, but in some universities such students may be in other departments (e.g., mathematics and physics). Eligibility for these fellowships is international, except where legal regulations or restrictions may apply. The award of this fellowship is nondiscriminatory and is not based on race, color, religion, sex, national origin, handicap, or family status. Immediate family members or close associates of any members of the evaluation committee or SUMMA officers/board members are ineligible.

The fellowship award is a grant of $15,000 (US dollars). This amount is to be spent in development of the ideas described in the proposal in pursuit of the PhD (or equivalent degree) at a schedule to be determined by the student and the university. If in the judgment of the evaluation committee there is no proposal of sufficiently high quality, no fellowship award will be made. It is anticipated that one such award will be given each year. The decisions of the evaluation committee and SUMMA, including but not limited to who receives a Fellowship, disbursement schedules, and whether a Fellowship will be awarded at all, are final.

The student should submit a proposal of at least five, but not more than twenty, double-spaced pages, explaining the concepts, originality, and potential applications. The page limitation must be observed. The proposal must be endorsed by the student's faculty advisor and supported by the university department or sub-department (e.g., antenna laboratory). The applicant should also include a curriculum vitae. The faculty advisor should submit an evaluation of the student's progress to date, including the student's graduate history and plans toward the PhD. Three other recommendations should be submitted by faculty members familiar with the student's work. All material submitted must be in English with standard EM units. All this material must be in the hands of the Chair of the evaluation committee no later than 1 February 2010 for consideration for an award in August 2010. Mail (or send by courier) all materials to the Chair:

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Information for authors

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